Mimo block spread CDMA: system architecture and performance analysis for high rate wireless systems

Tio Surya Dharma

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MIMO BLOCK SPREAD CDMA: SYSTEM ARCHITECTURE AND PERFORMANCE ANALYSIS FOR HIGH RATE WIRELESS SYSTEMS

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Abstract

High speed wireless access of the order of Gbps is envisioned for future mobile communication systems. However, symbol duration decreases as transmission rate increases. As a result, the large number of multipaths makes the communication channel to be severely frequency-selective. Recently, the use of frequency-domain (FD) processing has been receiving a lot of attention and is seen as one of the most promising technique to overcome the frequency-selectivity problem. FD processing can be realized using either multicarrier scheme that is based on the orthogonal-frequency-division-multiplexing (OFDM) technique or single-carrier scheme that employs cyclic-prefix or guard intervals.

Other than transmission rate enhancement, future mobile communication systems require multiple accesses that delivers higher system capacity and flexibility to support variable rate transmission. Code-division-multiple-access (CDMA) is widely recognized as one enabling technique that satisfies such criteria. However, CDMA deployment in most existing system today employs time-domain processing such as RAKE receiver or Time-domain Equalization (TDE). While both techniques work well for current rate requirement, the frequency-selectivity of the channel that characterizes future communication system requires high processing complexity and hence making both methods practically unsuitable. The combination of CDMA and FD processing looks promising to overcome this problem.

The other fundamental issue in realizing the high data-rate capability is the spectral efficiency. As it has been and will always be, frequency spectrum is a scarce resource. The amount of bandwidth allocated for future communication system only sees a modest increment as compared to the enhancement in the data rate. The most direct way of increasing the transmission power has been ruled out due to many concerns such as hardware complexity. Some of the possibilities in enhancing spectral efficiency is the use of higher modulation schemes (such as Quadrature-Amplitude Modulation (QAM)) and the incorporation of multiple antennae at both the transmitter and receiver. It has been shown that the use

of multiple antennae promises dramatic performance improvement.

The focus of this thesis is to design an advanced transceiver structure as a candidate for future mobile communication system. The use of frequency-domain processing in CDMA is studied. Conventional methods such as multicarrier CDMA (MC-CDMA) and cyclic-prefix assisted CDMA (CP-CDMA) are found to be either unsuitable or require high computation complexity for uplink recovery. A novel CDMA architecture that makes use of block processing and frequency-domain equalization has been proposed as an alternative for uplink and referred to as block spread CDMA (BS-CDMA). In MC-CDMA and CP-CDMA, chip alignment is lost during asynchronous transmission. This results in the destruction of code orthogonality and gives rise to multi-user interference (MUI). In BS-CDMA, the block transmission structure that results from chip-interleaving allows orthogonality to be maintained during asynchronous transmission and is virtually free from MUI during low mobility conditions. The BS-CDMA architecture is enhanced further by the use of multiple antennae, resulting in MIMO BS-CDMA architecture. Three MIMO architectures are studied, namely layered space-time (LST), space-time coded (STC) and space-frequency coded (SFC). LST BS-CDMA architecture enhances data rate by transmitting independent data simultaneously from different antenna. STC BS-CDMA architecture improves the error-rate performance by adopting space-time coding algorithms such as Alamouti (for 2 transmit antennae) code and hence allowing higher modulation scheme to be used. SFC BS-CDMA offers an alternative to STC BS-CDMA. Compared to STC BS-CDMA, SFC BS-CDMA shortens the processing delay, making it a more suitable choice for delay-critical applications. The performance of MIMO system relies on the accuracy of the channel state information (CSI) that is available. One part of this thesis studies the design of a good pilot structure and the estimation algorithm that can be used to obtain accurate CSI. Meanwhile, the performance of BS-CDMA deteriorates as the mobility condition increases. A key parameter known as channel coherence ratio (CCR) has been proposed in this work. Based on CCR value, it is possible to determine the performance of a BS-CDMA system. The last part of this work addresses the MUI that arises during high mobility condition. Interference cancellation methods have been proposed as a part of this work. In practical deployment, the mobility condition of transmitting users tends to vary. In BS-CDMA, the amount of interference inflicted by a user to the remaining users depends on its mobility condition. The proposed interference cancellation methods make use of the mobility condition and detection ordering is made based on the degree of mobility.

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List of Abbreviations

3G	Third Generation Mobile Communication Networks
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase-Shift Keying
BS	Base-Station
BS-CDMA	Block Spread CDMA
CDMA	Code-Division-Multiple-Access
CDS	Channel Delay Spread
CFO	Carrier Frequency Offset
CP	Cyclic Prefix
CP-CDMA	CP-Assisted Single-Carrier CDMA
CSI	Channel State Information
DAB	Digital Audio Broadcasting
DBLAST	Diagonal Bell Labs Layered Space-Time
DCO	DC Offset
DS-CDMA	Direct-Sequence CDMA
DVB-T	Digital Video Broadcasting Terrestrial
FDMA	Frequency-Division-Multiple-Access
FFT	Fast Fourier Transform
GCOD	Generalized Complex Orthogonal Design
GROD	Generalized Real Orthogonal Design
HIU	High-Interferer User
IFFT	Inverse Fast Fourier Transform
IBI	Inter-Block Interference
ICI	Inter-Channel Interference
IEEE802.11	IEEE Standardization Group for Wireless LAN

IF	Intermediate Frequency
IFI	Inter-Finger Interference
ILI	Inter-Layer Interference
ISI	Inter-Symbol Interference
ITU	International Telecommunication Union
LIU	Low-Interferer User
LST	Layered Space-Time
MEMS	Micro-Electro-Mechanical Systems
MC-CDMA	Multicarrier CDMA
MIMO	Multiple-Input-Multiple-Output
MISO	Multiple-Input-Single-Output
MLD	Maximum-Likelihood Decoder
MMSE	Minimum Mean-Square Error
MRC	Maximal-Ratio Combining
MS	Mobile-Station
MSE	Mean-Square Error
MSMC	Multi-stage SMC
MUI	Multi-User Interference
OFDM	Orthogonal-Frequency-Division-Multiplexing
ORC	Orthogonal-Restore Combining
PAPR	Peak-to-Average-Power-Ratio
PHY	Physical Layer (of a Communication System)
QAM	Quadrature-Amplitude Modulation
QPSK	Quadrature Phase-Shift Keying
RF	Radio-Frequency
SC-FDE	Single-Carrier with Frequency-Domain Equalization
SISO	Single-Input-Single-Output
SMC	Serial MUI Cancellation
SNR	Signal-to-Noise Ratio
SS	Spread-Spectrum
STBC	Space-Time Block Coding
STTC	Space-Time Trellis Coding
TDE	Time-Domain Equalization
TDMA	Time-Division-Multiple-Access

UMTS	Universal Mobile Terrestrial Standards
VBLAST	Vertical Bell Labs Layered Space-Time
VOSF	Variable Orthogonal Spreading Factor
ZF	Zero-Forcing
ZIF	Zero-IF
ZP	Zero-Padded

List of Symbol Notations

a	Variable
a^*	Conjugate of Variable a
$\mathbf{a}$	Vector
$\mathbf{a}^T$	Transpose of Vector $\mathbf{a}$
$\mathbf{a}^H$	Conjugate Transpose of Vector $\mathbf{a}$
$\mathbf{A}$	Matrix
$\mathbf{A}^T$	Transpose of Matrix $\mathbf{A}$
$\mathbf{A}^H$	Hermitian Transpose of Matrix $\mathbf{A}$
$\mathbf{A}^+$	Pseudoinverse of Matrix $\mathbf{A}$ defined as $(\mathbf{A}^H \mathbf{A})^{-1} \mathbf{A}^H$
$\mathbf{I}_M$	Identity Matrix of size $M \times M$
$\mathbf{F}_M$	M -Point FFT Matrix of size $M \times M$
$\otimes$	Kronecker Product
$\oplus$	Linear Convolution
$\odot$	Circular Convolution

List of Publications

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1. S.D. Tio, A.S. Madhukumar and A.B. Premkumar, "Layered Space-Time Architecture for MIMO Block Spread CDMA Systems," *IEEE Commun. Lett.*, vol. 10, no. 2, pp. 70-72, February 2006
2. S.D. Tio, A.S. Madhukumar and A.B. Premkumar, "MIMO Block Spread CDMA," *IEEE Trans. Wireless Commun.*, vol. 7, no. 6, pp. 1987-1992, June 2008
3. X. Peng, S.D. Tio, F. Chin and A.S. Madhukumar, "Novel interference cancellation methods for BS-CDMA in uplink broadband mobile communication systems," *IEEE Commun. Lett.*, vol. 12, no. 8, pp. 560-562, August 2008
4. S.D. Tio, A.S. Madhukumar, A.B. Premkumar and F. Adachi, "Estimation of CSI for LST BS-CDMA Systems using Special Sequences with Pre-defined Spectral Nulls," *Submitted to IEEE Trans. on Wireless Commun.*

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3. S. D. Tio, A. S. Madhukumar, A. B. Premkumar, and X. Peng, "On Channel Estimation for Layered Space-Time Block Spread CDMA Systems," *63rd Vehicular Technology Conference*, vol. 4, pp. 1722-1727, May 2006.

4. S. D. Tio, A.S. Madhukumar and A.B. Premkumar, "Space-Frequency Coded BS-CDMA for Broadband Mobile Communication Systems," *IEEE Wireless Communication and Networking Conference (WCNC)*, pp. 207-211, March 2007
5. S.D. Tio, X. Peng and F. Chin and A.S. Madhukumar, "MUI Cancellation for Uplink BS-CDMA in Broadband Mobile Communication Systems," *68th IEEE Vehicular Technology*, pp. 1-5, September 2008
6. S.D. Tio, A.S. Madhukumar and A.B. Premkumar, "MUI Cancellation for Uplink LST BS-CDMA in Broadband Mobile Communication Systems," *In Preparation*

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Chapter 1

Introduction

1.1 Background

In the early days, application of wireless communication was limited to receiving and transmitting voice data. Since then, rigorous research activity in wireless communication led to the understanding of some fundamental properties and limitations that arise because of the nature of the wireless propagation channels. Soon, this awareness led to the discovery of new techniques that allowed data to be transmitted at higher rate with lower transmission cost. While the development of wireless techniques has seen tremendous advancement, the growth of Internet subscribers is even more prolific. In just one to two decades, Internet applications have evolved substantially from the simple text-based email communication to multimedia-rich content based applications. In comparison to the fixed wired counterpart, the capabilities offered by mobile wireless communication systems are still lagging far behind. This phenomenon motivates our investigation to unravel new techniques that will enhance the capability of existing mobile communication systems in order to provide low-cost high-rate mobile data transmission.

In wireless communication, one needs to distinguish the difference between fixed and mobile wireless communication. The main difference is that the channel coherence time (CCT) in fixed wireless communication system is normally high which essentially means that the channel is slowly-varying, while CCT in mobile wireless communication system is normally low or rapidly-varying. To combat the rapidly-varying channel, pilot symbols or training

information has to be transmitted more frequently which would then require a more efficient way of obtaining the channel estimates.

It is widely established that the major impairment encountered by any wireless communication system is the impact caused by multipath phenomenon. When transmitted over the air, the signal will reach the intended recipients through not one but multiple different paths with different time delays. Depending on the terrain condition and the presence of obstruction such as trees, buildings etc, these paths can be of different length and cause different level of attenuation to a transmitted symbol. This gives rise to the problem of inter-symbol interference (ISI) where previously transmitted symbols arrive simultaneously with the current symbol resulting in superposition of multiple symbols. In an ideal scenario where full Channel State Information (CSI) such as the channel gain of each path and the arrival timing of each symbol are available, symbol detection with high accuracy is possible, at the cost of additional processing to compensate the channel effect. However, CSI is usually unknown in practical deployment. Hence, estimation of CSI is usually required prior to symbol detection. The severity of ISI is aggravated in high rate systems where the number of paths can be enormous and to consider all of them is sometimes prohibitive given the constraints of hardware capability. This means that a practical wireless communication system has to make use of the partial or imperfect CSI, which makes symbol detection a much challenging task.

In the previous and some of the existing mobile technologies, time-division multiple access (TDMA) or frequency-division multiple access (FDMA) divide the time or frequency resource equally among multiple users. However, both time and frequency are precious resources and there is a limit to the maximum number of users that a TDMA-based or FDMA-based system can support. The introduction of code-division multiple access (CDMA) opened up another domain that promised significant enhancement in system capacity along with other benefits such as robustness against jamming noise, enhanced security and privacy.

While discovery of CDMA adds the third dimension in wireless resource domain, recent advancement in digital processor technology allows efficient implementation of FFT/IFFT operations. This triggered much interest in orthogonal frequency division multiplexing (OFDM)-based multicarrier transmission that simplifies equalization process and offers robustness against channel with deep fades. Combining CDMA and OFDM techniques led to multicarrier CDMA (MC-CDMA) which proved to be highly effective and promising. How-

ever, MC-CDMA is prone to problems plaguing multicarrier schemes, namely the peak-to-average power ratio (PAPR) issue that requires expensive and inefficient amplifier and the susceptibility to carrier frequency offset (CFO) that requires constant tracking and correction. Single-carrier with frequency-domain equalization (SC-FDE) was proposed to simplify the hardware requirement while keeping the advantages of multicarrier schemes such as the simple equalization procedures and robustness against channel with long delay spread. A BS-CDMA transceiver structure that employs SC-FDE processing is often referred to as Cyclic-prefix assisted single-carrier CDMA (CP-CDMA), which has been proposed as an attractive alternative to MC-CDMA system.

Another advancement in the wireless communication research resulted from the advancement in microchip technology that allows multiple antenna to be fitted cheaply into small handheld devices. The exploitation of multiple transmit and receive antennae results in the creation of multiple input multiple output (MIMO) channels. MIMO system has been theoretically proven to enhance system performance and capacity dramatically and has attracted tremendous interest among the research community.

1.2 Problem Statement

The prolific growth in communication traffic and evolution of mobile communication applications present tremendous commercial opportunity for communication service providers. However, there are numerous challenges that require extensive research before communication system can successfully deliver its full potential. This thesis attempts to address few of the many challenges to design an advanced transceiver architecture that is capable of delivering the high transmission rate requirements of future applications, supporting large pool of subscribers while maintaining transmission cost low (affordable) and keeping device size small.

A fundamental issue in any communication system is the ability to allow multiple users to share the communication channel effectively. In the past, such multiple access capability was achieved using either time-division-multiple-access (TDMA) or frequency-division-multiple-access (FDMA) techniques. However, both schemes are considered to be ineffective when large number of users share the time/frequency resources. Moreover, both do not offer the flexibility to support asymmetric traffic between the forward and reverse link. Today,

CDMA is widely accepted as a more effective resource sharing mechanism than its two predecessors. Given that time and frequency are limited and precious resources, CDMA allows multiple users to transmit simultaneously at the same time using the same frequency. In downlink transmission where BS transmits data designated for numerous users, chip alignment is maintained and orthogonality allows perfect recovery of one's designated information while signals belonging to other users appear as noise. However, CDMA system faces a more daunting challenge during uplink transmission. Since CDMA allows all users to transmit all the time, it is safe to assume that there is minimal or no coordination among the different users before transmitting their signals. To make things worse, these transmitters are mobile or constantly on the move. This results in asynchronous communication where signals from different users arrive at the BS at different time. As a result, chip alignment is lost and code orthogonality is often destroyed. When this happens, the fundamental principle of CDMA is violated and causes serious implications to the signal recovery process. In addition, the complexity of traditional CDMA processing is directly proportional to the length of the spread code used. Longer spread code would increase the number of users supported and thus the overall system capacity. This means that the two goals, namely to keep processing complexity low and to increase system capacity contradict one another. This problem can be overcome by adopting block-processing instead of chip-processing. Block-spread CDMA (BS-CDMA) is one such scheme and thus forms an important part of this work.

The second fundamental issue of a mobile communication system is the ability to overcome the multipath phenomenon. As signal is transmitted from a transmit antenna, the surrounding objects cause refraction, reflection etc. This in turn lead to multiple reflections of the same symbols arriving at the receiver at different time with different attenuation. This gives rise to the problem of channel-induced inter-symbol interference (ISI). If ISI is not compensated appropriately, it would cause severe degradation to the system performance. In this context, OFDM is widely accepted as the most effective technique to combat multipath impairment. In OFDM system, the multipath impairment is turned into an advantage and the equalization consists only of a simple one-tap multiplication. However, OFDM system faces two major shortcomings. The first is the PAPR issue that necessitates the use of expensive and inefficient amplifier. The second is its performance sensitivity due to uncompensated CFO that demands constant frequency-offset tracking and additional processing to correct it. In this context, single-carrier system with frequency-domain equalization (SC-FDE) technique has emerged recently as an attractive alternative to OFDM.

This work focuses on combining the SC-FDE mechanism with BS-CDMA architecture in order to deliver an enhanced system performance.

The third fundamental issue of a mobile communication system is the capacity limitation that is bounded by *Shannon's* limit. In the past, state-of-the-art technology in areas such as Micro-Electro-Mechanical-Systems (MEMS), put a practical cap on the number of antenna that can be employed. This meant that regardless of the signal-processing algorithms used, the maximum amount of data that can be delivered is limited due to the availability of only a single data pipe. The recent advances in MEMS and antenna technology has opened up the possibility of fitting multiple antennae at both the transmitter and receiver. The use of multiple antennae at both transmitter and receiver promises dramatic enhancement in system performance and capacity. The later part of this work focuses on devising an efficient architecture that combines the three signal processing (CDMA with block processing, Frequency-Domain equalization and MIMO) capabilities to deliver the kind of performance that is in-line to meet the challenges of future mobile communication systems.

The contribution from this thesis has been published in the following journals and conferences.

1. The Layered Space-Time BS-CDMA (LST BS-CDMA) architecture discussed in Chapter 4 was published in IEEE Vehicular Technology Conference and IEEE Communication Letters;
2. The Space-Time Coded BS-CDMA (STC BS-CDMA) architecture discussed in Chapter 5 was published in IEEE Global Telecommunication Conference;
3. The Space-Frequency Coded BS-CDMA (SFC BS-CDMA) architecture discussed in Chapter 6 was published in IEEE Wireless Communication and Networking Conference;
4. The study on pilot design techniques and estimation algorithms to obtain channel state information (CSI) for LST BS-CDMA discussed in Chapter 4 was published in IEEE Vehicular Technology Conference and has been submitted to IEEE Transaction on Wireless Communications;
5. The practical challenges faced by LST and STC BS-CDMA systems as discussed in Chapter 4 and 5 were published in IEEE Transaction on Wireless Communications;

and

6. The study on interference cancellation techniques discussed in Chapter 7 was published in IEEE Communication Letters and IEEE Vehicular Technology Conference.

1.3 Thesis Outline

The rest of this thesis is organized as follows:

Chapter 2 presents an in-depth literature review of the existing techniques that are used to address the three fundamental issues discussed in Section 1.2 of this thesis. The first part of Chapter 2 discusses the signal processing blocks that form the basics of a CDMA system. This is followed by the different variants of CDMA schemes that employ advanced signal-processing techniques such as frequency-domain processing in order to enhance the robustness of the overall system performance. The later part of Chapter 2 discusses various ways of enhancing system performance and capacity with the use of multiple antennae at both the transmitter and the receiver.

Chapter 3 discusses the basic block-spread CDMA (BS-CDMA) architecture that forms the basic of the present research work. Details of the signal-processing mechanism employed at both the transmitter and the receiver are presented. This is followed by theoretical analysis on the system performance that serves as the benchmark to the simulation studies. The last part of Chapter 3 presents a detailed description of the simulation studies and the corresponding discussions based on the results.

Chapter 4 discusses an enhanced architecture that combines Layered Space-Time (LST) architecture with BS-CDMA and referred to as LST BS-CDMA. The processing arrangement at both the transmitter and the receiver is first presented. This is followed by a study on the channel estimation problem. A detailed description of the problem and the technique used to construct the pilot sequences is presented. This is followed by an elaborative description of the simulation studies and the numerical results obtained from the simulation exercises.

Chapter 5 presents Space-Time Coded architecture for BS-CDMA (STC BS-CDMA) system as an enhancement alternative to LST BS-CDMA system. The basic architecture of a STC BS-CDMA system with only two transmit antennae is first presented. This is followed

by generalization of the orthogonal code construction using arbitrary number of transmit antennae. The last part of Chapter 5 presents the details of the simulation investigation and the findings based on the numerical results obtained.

Chapter 6 presents Space-Frequency Coded architecture for BS-CDMA (SFC BS-CDMA) system. The proposed SFC BS-CDMA serves as an alternative to STC BS-CDMA that offers lower processing delay and improved performance during fast fading channel condition. Similar to Chapter 5, the basic architecture of a SFC BS-CDMA system with only two transmit antennae is first presented, followed by a generalization of the orthogonal code construction using arbitrary number of transmit antennae. A discussion of the various aspects affecting SFC BS-CDMA system is then presented. The later part of Chapter 6 presents the details of the simulation studies and the findings based on the numerical results.

Chapter 7 discusses some of the challenges facing practical deployment of the various BS-CDMA systems that are presented in Chapter 3-6. The main issue arises when users are mobile and they experience a rapidly fading communication channel. In some cases, this violates the fundamental principle of BS-CDMA system and hence causes imperfect removal of the multi-user interference (MUI). An improved MMSE equalizer is proposed that takes into consideration of the MUI component. Interference cancellation techniques that attempt to mitigate the degradation impact due to MUI are also presented.

Chapter 8 concludes the thesis and discusses some of the points for the future research directions in the related area.

Chapter 2

Background and Literature Review

2.1 Introduction

This chapter presents an in-depth review of code-division-multiple-access (CDMA) and its variations. Clearly, CDMA alone will not be able to overcome all the challenges and fulfill all the requirements of future mobile communication systems. The combination of CDMA with other advanced signal processing algorithms is inevitable. Several such algorithm like frequency-domain processing and MIMO signal processing are discussed. The challenges encountered in the attempt to combine these techniques with CDMA are also addressed in detail.

The organization of Chapter 2 is as follows. Section 2.2 reviews the origin of code-division-multiple-access (CDMA) and how it has evolved into the *de-facto* choice for commercial deployment targeting high data rate transmission. Section 2.2.1 presents the basic CDMA system, namely direct-sequence CDMA (DS-CDMA), and the different receiver processing techniques to recover the transmitted symbols. Section 2.2.2 presents the use of multicarrier transmission to transmit users' chip information. Section 2.2.3 presents an alternative to the multicarrier transmission, namely the use of frequency-domain equalization with single-carrier transmission. Section 2.3 discusses the use of antenna diversity as a technique to enhance spectral efficiency usage.

2.2 The Origin and Evolution of Code-Division-Multiple-Access (CDMA)

The origin of spread spectrum (SS) communication can be traced back to the early 1900s [1][2]. The exploitation of spread spectrum for transmitting information was mainly motivated by military needs for a communication system that is able to withstand enemy's jamming signal and to escape enemy's detection. Since then, multiple access scheme based on SS principle has come a long way and various forms of spread spectrum multiple access (SSMA) has been adopted in many of today's wireless communication standards.

A typical SS system is generally characterized by the following features:

1. The bandwidth used for transmission is much larger than the original bandwidth of the useful information symbols;
2. The carrier signal used in SS system is a wideband signal with a short pulse duration. However, the carrier must be of low duty cycle and should appear like a random noise (pseudorandom);
3. A receiver with knowledge of the pseudorandom carrier can recover the transmitted information symbols by performing correlation of the known pseudorandom sequence with the received signal.

Based on the general characteristics above, any SS system is associated with one important parameter, namely the 'multiplicity factor' which is commonly referred to as 'processing gain' nowadays. The processing gain defines the number of possible orthogonal signaling or carriers which is the maximum number of users that can be supported in a system. In many cases, the processing gain is simply the ratio between the expanded bandwidth to the original information bandwidth. Though larger processing gain means better performance and higher capacity, the maximum processing gain that can be used is limited by practical considerations such as the available bandwidth and hardware complexity in producing the carrier and processing involved to recover the transmitted information. With the use of wideband and pseudorandom noise-like carrier, a SS system offers three inherent advantages:

1. Anti-Interception Capability. The robustness of a SS system against detection depends on the processing gain used. To detect the presence of a SS transmission with

processing gain in the range of thousands, a surveillance receiver first needs to know all the thousands of orthogonal signaling format and secondly, determines which of those is being used. This makes interception of a properly designed SS system a daunting task or virtually impossible;

2. Anti-Jamming Capability. During World War II, the use of jamming and anti-jamming tactics was prevalent and significant amount of resources were dedicated to search for a transmission method with the best anti-jamming capability and SS technique emerged as the ultimate choice. Since carrier signal in SS system is 'random' and unpredictable, jamming methods that rely on signal observation to improve its performance will fail in this case;
3. Cryptographic Capability. In SS system, the carrier signaling can be seen as a kind of encryption process. In order to recover the transmitted symbols, a receiver needs to know the carrier signal that is being used. The demodulation process at the receiver that performs cross-correlation between the reference and received signal can be seen as the decryption process.

Today, SS multiple-access is more commonly referred to as Code-Division-Multiple-Access (CDMA). Though the use of SS technique was prevalent in USA's military communication by the end of World War II, the real breakthrough to adopt SS technique into commercial deployment came much later. It was only in 1995 that IS-95A [3], the first commercial CDMA standard, was published. Later on, IS-95A evolved and was known as cdmaOne [4][5] and was deployed worldwide in countries such as USA, Japan, Korea etc. The adoption of CDMA as an industry standard is more widespread in today's 3G standards that were developed under International Telecommunications Union's Universal Mobile Terrestrial Standards (ITU-UMTS) [6].

According to ITU's Open Systems Interconnection Reference Model (OSI Reference Model), a network communication architecture consists of 7-layers. Starting from the top to bottom, the 7-layers are the Application, Presentation, Session, Transport, Network, Data Link and Physical Layers. In this context, CDMA fits into the Physical Layer in the overall system design.

2.2.1 CDMA Basic: Direct-Sequence CDMA (DS-CDMA)

In wireless communication system, multiple access refers to the system capability in allowing simultaneous transmissions from multiple users to occur at a common transmission medium. Previously, multiple access communication was achieved by allowing different user to transmit during different time-slots (TDMA) or frequency-slots (FDMA). One major drawback encountered in TDMA and FDMA is the static allocation of resources to users. In voice communication system, the probability that all users are talking simultaneously is low. According to [7], voice signals are intermittent with a duty cycle of approximately $\frac{3}{8}$. Under such condition, static resource allocation can be very ineffective. Earlier, Shannon mentioned that CDMA is deemed as the fairest way of allocating resources as well as distributing noise or other channel impairments [8]. This is because in TDMA or FDMA system, 'simultaneous' transmission by multiple users occur during different time or frequency slots. This means that it is likely for certain users to experience poor or deep fading channel while some other users may be enjoying good and reliable channel conditions due to time or frequency selectivity [9][10]. However, this 'unfair' advantage is non-existent in CDMA system as all users get to transmit at the same time on the same frequency band and hence a common medium. As more and more users become active, performance of CDMA system starts to degrade as cross-talk interference starts to take effect. Such 'graceful degradation' is preferred than abrupt degradation such as call blocking and is a feature unique to CDMA. Another advantage of CDMA is the simple frequency planning as the frequency reuse factor is close to unity.

CDMA supports variable data rate transmission. With rapid evolution of services that is made available to mobile subscribers, it appears inevitable that variable transmission rate capability will be one key characteristic of future mobile communication system in order to support diverse nature of applications which may range from simple low-rate short text messaging services to bandwidth hungry multimedia streaming applications. Incorporation of variable rate capability in a CDMA system can be accomplished through two means. The first method is referred to as variable orthogonal spreading factor (VOSF). Construction of such code and its application for broadband wireless communication system has been studied in [11][12][13]. However, VOSF CDMA system is plagued by the problems of higher call blocking rate for high data rate users which may lead to lower spectral efficiency utilization. The use of dynamic code allocation has been proposed [14] to mitigate this issue, at the

expense of higher processing cost incurred. Alternatively, variable rate transmission can be achieved by assigning multiple codes to a user that requires high data rate transmission [15]. While such multicode assignment is free from the call blocking issue encountered in VOSF method, multiple receiver processing unit is required which complicates the receiver architecture and hence is usually less favoured than the VOSF technique.

2.2.1.1 DS-CDMA Transmit Signal Model

In direct-sequence CDMA (DS-CDMA) system [12], the information symbols are first spread into chips of smaller duration. The ratio between the symbol duration, T_S and chip duration, T_C is given by the processing gain, $G = \frac{T_S}{T_C}$. After spreading, the baseband model of the signal before transmission is given by:

$$\mathbf{x}_k(i) = \mathbf{c}_k \otimes d_k(i) \quad (2.1)$$

where $\mathbf{x}_k(i)$ refers to the chip vector belonging to the k^{th} user during i^{th} instance, $\mathbf{c}_k = [c_{k,1} \cdots c_{k,G}]^T$ is the k^{th} user's spreading code, $d_k(i)$ is the i^{th} symbol for the k^{th} user and $\otimes$ denotes *Kronecker* product operation. For brevity, the variable i will be omitted in subsequent discussions. For downlink transmission, the chips from all users are first summed up before transmission. In the presence of pure additive white gaussian noise (AWGN), the received signal model is identical for both downlink and uplink transmission and is given as:

$$\mathbf{r} = \sum_{k=1}^K \mathbf{x}_k + \mathbf{n} \quad (2.2)$$

where K is the number of active users in the system and $\mathbf{n}$ is the AWGN component. Based on Equation (2.2), recovery of symbol information consists of two stages. In the first stage, the test statistic is computed by correlating the received chips with the respective user's spreading code. The test statistics will then be used in the second stage, where demodulation or symbol detection is performed. The test statistic for user k_0 is given as:

$$\begin{aligned} v_{k_0} &= \mathbf{c}_{k_0}^T \mathbf{r} \\ &= \mathbf{c}_{k_0}^T \mathbf{c}_{k_0} d_{k_0} + \sum_{k=1, k \neq k_0}^K \mathbf{c}_{k_0}^T \mathbf{c}_k d_k + \mathbf{c}_{k_0}^T \mathbf{n} \\ &= d_{k_0} + \mathbf{c}_{k_0}^T \mathbf{n} \end{aligned} \quad (2.3)$$

where

$$\mathbf{c}_{k_0}^T \mathbf{c}_k = \begin{cases} 1 & \text{for } k = k_0 \\ 0 & \text{otherwise} \end{cases} \quad (2.4)$$

The property shown in Equation (2.4) assumes that orthogonal signaling or spreading code is used by all users and the auto-correlation is normalized to unity. It can also be shown that the despreading operation does not alter the noise characteristics. This forms the fundamental principle of any CDMA system, in that a particular user is able to decode (understand) only the information that is meant for their own, while treating information belonging to all other users as noise. The pure AWGN scenario presents the simplest case. In practical scenario where the chips are attenuated by channels with varying gain and delay spread, typical DS-CDMA receivers as described below are normally used.

2.2.1.2 DS-CDMA Receivers

In multipath environment, transmitted signal reaches the receiver through not one but many paths. This means that a receiver hears multiple echoes arriving at different and randomly varying delays and amplitude, depending on several factors such as the distance travelled by each path, the presence of obstruction and sometimes atmospheric conditions. As a result, delayed replicas of the current symbol will spillover to the next arriving symbols, giving rise to a phenomena referred to as inter-symbol interference (ISI). It has been widely recognized that if not mitigated properly, ISI due to multipath is one major factor limiting the transmission capability of time-domain receiver. Two time-domain equalization methods to recover transmitted information symbols and how both methods fail or become practically unsuitable in the presence of severe multipath is described in the following section.

DS-CDMA Receivers I: RAKE Receiver

RAKE receiver is one type of diversity combining receiver technique. In essence, a RAKE receiver tries to turn multipaths into an advantage rather than a disadvantage. A general RAKE arrangement is shown in Figure 2.1. As shown in the Figure, a RAKE receiver is simply made up of a series of matched filters along a tapped delay line. Each filter tap is commonly referred to as a RAKE finger and each finger corresponds to a single resolvable path. In other words, each tap plays the role of reconstructing the energy of the respective path and at the end of the line, energy from all paths are harnessed back which can then

be used as the test statistic for symbol detection.

Assume that the sequence of symbols to be transmitted is represented by the vector $\mathbf{d} = [d(0) \cdots d(L-1)]^T$ and the channel in which the symbols propagate is given by $\mathbf{h} = [h(0) \cdots h(L-1)]^T$. The signal received over L symbol periods is given as:

$$\begin{aligned} r(0) &= h(0)d(0) + n(0) \\ r(1) &= h(0)d(1) + h(1)d(0) + n(1) \\ &\vdots \\ r(L-1) &= h(0)d(L-1) + \cdots + h(L-1)d(0) + n(L-1) \end{aligned} \quad (2.5)$$

Figure 2.1: RAKE Receiver Architecture

Based on Figure 2.1, a tap-weight w_l is multiplied with the delayed version of the received signal. In this case, $w_l = \frac{h^*(l)}{\sum_{m=0}^{L-1} |h(m)|^2}$. Now, assume that the RAKE receiver is attempting to generate the test statistics for $d(0)$. At the output of the integrator, the test statistic

$v(0)$ is given by:

$$\begin{aligned} v(0) &= \sum_{l=0}^{L-1} w_l h(l) d(0) + \sum_{l=1}^{L-1} \sum_{n=0}^{l-1} w_l h(n) d(l-n) + \sum_{l=0}^{L-1} h^*(l) n(l) \\ &= d(0) + \sum_{l=1}^{L-1} \alpha_l + \sum_{l=0}^{L-1} h^*(l) n(l) \end{aligned} \quad (2.6)$$

where

$$\alpha_l = \frac{\sum_{m=0}^{l-1} h^*(l) h(m) d(l-m)}{\sum_{m=0}^{L-1} |h(m)|^2} \quad (2.7)$$

is the zero-mean interfering components when the multipaths are assumed to be uncorrelated.

At one time, RAKE receiver was widely recognized as a better solution for the multipath issue and it was adopted in IS-95[3], the first commercial CDMA standards launched in USA.

From a simple observation, it appears that since RAKE is a path diversity combiner, larger number of paths would be able to give a better performance. However, an increase in the number of RAKE fingers leads to two major consequences. Firstly, having more fingers translates to higher amount of inter-finger interference that would lead to serious performance degradation at some point [16]. Secondly, implementing a RAKE receiver with large number of fingers in hardware is a tremendous challenge and there is a limit on the maximum number of fingers before hardware cost becomes economically unjustifiable. Though there is a possibility of reducing the number of RAKE fingers by carefully selecting those paths with higher energy and resolving only those paths, such approach would incur a trade-off with system performance. In addition, additional processing and advanced algorithms are required to incorporate the path selection process. Instead of employing RAKE fingers at the receiver, the use of pre-RAKE filter at the transmitter has been proposed [17][18] to combat multipath effect of high data rate communication channel. With a pre-RAKE filter, a symbol is transmitted in multiple succession with the interval between consecutive re-transmission corresponding to a single path delay. Effectively, a pre-RAKE filter transmits a symbol filtered by the channel inverse and symbol processing at the receiver then becomes a trivial task of collecting the right samples. However, the use of pre-RAKE filter reduces the overall transmission rate capability. To mitigate the loss in transmission rate, partial pre-RAKE filter which considers only part of the channel impulse response has also

been proposed at the cost of performance inferiority. Nonetheless, practical implementation of pre-RAKE transceiver is challenging as it requires the receiver to determine the path selection and then send a feedback on channel information to the transmitter. Due to these shortcomings, the popularity of RAKE receiver is declining and this is especially true for high-rate communication systems where large number of multipaths (tens or hundreds) characterizes the channel propagation model. As such, a search for a more suitable alternative led to frequency-domain processing which will be discussed in the later part of this chapter.

DS-CDMA Receivers II: Time-Domain Equalization

Another school of thought in using time-domain processing to recover transmitted symbols is a class of time-domain equalizers (TDE). The basic idea behind TDE is to collect the received samples and stack them one after another to form a column vector that can then be equalized using various equalization techniques such as minimum mean square equalization (MMSE) or zero-forcing (ZF). The simplified signal model of a typical TDE is shown below.

$$\mathbf{r} = \begin{bmatrix} h(0) & \cdots & \cdots & \cdots & \cdots & 0 \\ h(1) & h(0) & \cdots & \cdots & \cdots & \vdots \\ \vdots & \ddots & \ddots & \cdots & \cdots & \vdots \\ h(L-1) & \ddots & h(1) & h(0) & \cdots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & \cdots & h(L-1) & \cdots & \cdots & h(0) \end{bmatrix} \begin{bmatrix} d(0) \\ \vdots \\ d(M-1) \end{bmatrix} + \begin{bmatrix} n(0) \\ \vdots \\ n(M-1) \end{bmatrix} \quad (2.8)$$

$$= \mathbf{H}\mathbf{d} + \mathbf{n}$$

where $\mathbf{H}$ is the $M \times M$ channel matrix. The equalized signal that forms the test statistics is obtained by pre-multiplying the stacked received signal with either MMSE matrix, $G_{MMSE} = (\mathbf{H}^H \mathbf{H} + \sigma^2)^{-1} \mathbf{H}^H$ or the ZF matrix, $G_{ZF} = \mathbf{H}^+$.

TDE receiver suffers from two major drawbacks. Firstly, performance of TDE is affected by inter-block interference. Mathematical proof of this can be easily derived from Equation (2.8), where it can be shown that the multipath effect of the last symbol in the block would spillover to the next block. Secondly, the matrix size for such TDE receiver grows proportionally with respect to the maximum number of paths in the system. Since matrix inverse operation is usually required to compute equalizer matrix such as in MMSE and

ZF techniques, such TDE receiver is difficult to implement when hardware constraints are taken into consideration. As a result, TDE is relatively unsuitable to combat channels with severe multipath impairments.

2.2.2 CDMA with Multicarrier Transmission

This section describes a generic multicarrier CDMA (MC-CDMA) transceiver architecture. First, OFDM technique as the basis of any multicarrier system is first discussed. After that, different architectural combinations of multicarrier and CDMA techniques are presented.

2.2.2.1 OFDM

Parallel communication system has long been studied as an alternative to serial transmission [19][20]. The whole idea behind it is to divide an available frequency bandwidth into smaller sub-bands where each sub-band is modulated using frequency that is different from the rest of the sub-bands. Each sub-band can be viewed as an independent channel or data pipe that can be used to transmit data independently. Efficient use of frequency spectrum is achieved when the spectral characteristics of the different sub-bands contain overlapping sidelobes. While spectral overlapping is a desirable feature, it has to be carefully designed such that the different sub-carrier frequencies used are orthogonal to each other and that these carriers do not cause inter-channel interference to one another. In its early days, such communication system did not attract much interest due to the technological limitations at that time and the complex hardware structure associated with it.

In [21], it was suggested that such parallel communication system can be realized through the use of Fourier transform generators and it was in [22] that the concept of orthogonal-frequency-division-multiplexing (OFDM) using discrete Fourier transform (IDFT/DFT) operation was first proposed. Since then, multicarrier transmission relying on OFDM modulator/demodulator has attracted significant research interest which is attributed partly due to the technological advances that allow for efficient and simple implementation of the Fast Fourier transform (FFT/IFFT) operation. To illustrate the idea, Figure 2.2 depicts a generalized and simplified arrangement of an OFDM transceiver. The Figure shows the case when the whole available bandwidth is divided into four sub-carriers and each sub-carrier signal can be seen as being generated with separate oscillators. At the receiver, each

oscillator will act as a matched filter for the respective sub-carrier signal from which the output can be used for equalization and detection.

In OFDM system, cyclic-prefix (CP) is usually appended before transmission. This will act as a guard interval that will prevent inter-block interference (IBI) from corrupting the received signal thus making signal recovery simpler. Though the introduction of CP reduces the overall transmission efficiency, the advantage of avoiding IBI makes such a tradeoff worthwhile. Nonetheless, the exploitation of CP in aiding signal detection and estimation of channel parameters [23][24][25] have been proposed which add to the advantages of using CP. However, it was highlighted in [26] that CP-based OFDM system is vulnerable to zero gain channel characteristic or deep narrowband fading in certain sub-carriers which may result in complete loss or destruction of the symbols carried by those sub-carriers. To address such vulnerability, zero-padding (ZP) can be used in place of CP. In [26], it is shown that ZP-OFDM is robust against such a channel which implies a guarantee of symbol recovery regardless of the channel zero location [27][28][29]. By sacrificing minimal loss in signal-to-noise ratio (SNR) due to the clipping in power amplifier (PA), ZP-OFDM system also offers improvement in tracking capability.

Despite all the advantages, there are a few inherent setbacks that come with any multi-carrier system. The first major shortcoming arises due to the high Peak-to-Average Power Ratio (PAPR) that results from the IDFT operation at the transmitter [30]. This requires the transmitter to employ a power amplifier that has a large linear range. However, equipping a mobile device transmitter with such an amplifier will make it inefficient in terms of cost, power consumption and size. Since this is a critical issue that determines the suitability of OFDM to be incorporated into mobile transmitter, extensive research has been done to analyse the performance degradation that is caused by limiting the transmitted power [31][32]. Meanwhile, numerous processing algorithms that attempt to reduce the PAPR without significantly affecting the quality of signal recovery at the receiver have been proposed [33][34][35][36].

The second principal disadvantage of any OFDM-based system is the performance sensitivity due to DC offset (DCO) and carrier frequency offset (CFO) [37]. The problem of DCO is a prominent issue in direct-conversion or Zero-IF (ZIF) architecture, where only one mixer stage is used to convert the baseband signal to and from RF frequency directly, thereby eliminating the IF stage. In ZIF architecture, DCO component is generated as

a result of the self-mixing between the local oscillators. The problem of DCO is usually avoided by leaving some of the sub-carriers unused. In IEEE802.11 PHY Specifications[38], only 52 sub-carriers have been used from the available 64 sub-carriers. Obviously, leaving sub-carriers unused reduces the overall system capacity. Meanwhile, the problem of CFO imposes a more serious threat. CFO arises due to the mismatch between the transmit and receive oscillators. As frequency separation between sub-carriers is small and OFDM-based system relies on the orthogonal property between sub-carriers, the frequency shift caused by CFO results in inter-carrier interference and destruction of the orthogonal property. The performance degradation due to CFO is usually measured in terms of the loss in SNR, ρ_{SNR} . According to [37][39], ρ_{SNR} of an OFDM-based system is given by:

$$\rho_{SNR} = \frac{10}{3\ln 10} (\pi f_0) M^2 \frac{E_s}{N_0} \text{dB} \quad (2.9)$$

where f_0 , M and $\frac{E_s}{N_0}$ corresponds to the normalized frequency offset component, number of OFDM sub-carriers and signal-to-noise ratio (SNR) respectively. Based on Equation (2.9), it can be seen that the degradation impact due to CFO increases with the number of sub-carriers and SNR value used. Various DCO and CFO estimation and correction techniques have been proposed in the literature [40][41][23].

2.2.2.2 Multicarrier CDMA (MC-CDMA)

Earlier, direct-sequence CDMA which performs spreading in time domain has been discussed. With time-domain spreading, the symbol is converted into multiple chips of much smaller duration. When the chip duration is smaller compared to the channel coherence time, the channel becomes frequency-selective and the worst channel path delay may span across a large number of symbols. As highlighted earlier, time-domain receiver becomes ineffective under such scenario due to the high processing complexity.

Over the recent years, there has been a growing interest in exploiting multicarrier transmission for CDMA systems. As the basic principle behind multicarrier transmission is the creation of parallel orthogonal sub-channels, there are two schools of thought on how these sub-channels can be used to carry the chip information. The first technique is to divide the single stream of symbols into multiple streams followed by spreading operation. The chips on each sub-stream will then be modulated using one of the sub-carrier frequencies. As

Figure 2.2: A General OFDM Transceiver Architecture

a result, each sub-stream will be of lower rate. This class of CDMA scheme is commonly referred to as MC-DS-CDMA [42] and its generalized baseband transceiver is shown in Figure 2.3(a). The second technique attempts to introduce frequency diversity to each symbol and its transceiver arrangement is shown in Figure 2.3(b). Unlike MC-DS-CDMA, spreading operation precedes the sub-stream multiplexing. It means that now, each sub-carrier is used to carry different chips from a single symbol. In this way, each symbol enjoys frequency diversity as it is highly unlikely that all chips will be destroyed simultaneously. This class of CDMA scheme is commonly referred to as MC-CDMA [43][44][45][46][47][48][49]. With superior performance due to the frequency diversity advantage, MC-CDMA is widely accepted as the better choice in combining the two technologies together.

Now, consider a MC-CDMA system with K number of users. In MC-CDMA transmitter, the result of the spreading operation is fed into an IFFT block. The output of such operation for the k^{th} user is given by:

$$\bar{\mathbf{x}}_k = \mathbf{F}_M^H (\mathbf{c}_k \otimes d_k) \quad (2.10)$$

where $\mathbf{F}_M$ corresponds to the $M \times M$ FFT matrix and $(.)^H$ refers to the transpose conjugate operation. For downlink transmission, the IFFT outputs from all K users are first summed up together before CP is appended and then transmitted simultaneously. On the other hand, CP is immediately appended to the IFFT outputs and information from each user is transmitted independently from the rest. At the receiver, CP removal is first carried out and the received signal can be expressed as

$$\mathbf{r} = \sum_{k=1}^K \mathbf{H}_k \bar{\mathbf{x}}_k + \mathbf{n} \quad (2.11)$$

where $\mathbf{H}_k$ is a circular Toeplitz channel matrix between k^{th} user's transmit and the receive antenna. After performing FFT operation, the received signal now becomes

$$\begin{aligned} \bar{\mathbf{r}} &= \sum_{k=1}^K \mathbf{F}_M \mathbf{H}_k \bar{\mathbf{x}}_k + \mathbf{F}_M \mathbf{n} \\ &= \sum_{k=1}^K \mathbf{F}_M \mathbf{H}_k \mathbf{F}_M^H \mathbf{x}_k + \bar{\mathbf{n}} \end{aligned} \quad (2.12)$$

According to [50], pre- and post-multiplying a circular Toeplitz matrix with FFT and IFFT matrix produce a diagonal matrix. The output of the FFT operation shown above can be

(a) MC-DS-CDMA

(b) MC-CDMA

Figure 2.3: Two General Multicarrier CDMA Transceiver Architectures

simplified as

$$\begin{aligned}\bar{\mathbf{r}} &= \bar{\mathbf{H}}_k \mathbf{x}_k + \bar{\mathbf{n}} \\ &= \sum_{k=1}^K \begin{bmatrix} \bar{h}_k(0) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \bar{h}_k(M-1) \end{bmatrix} \begin{bmatrix} c_{k,1}d_k \\ \vdots \\ c_{k,G}d_k \end{bmatrix} + \bar{\mathbf{n}}\end{aligned}\quad (2.13)$$

where $\bar{\mathbf{H}}_k = \mathbf{F}_M \mathbf{H}_k \mathbf{F}_M^H = \text{diag} \{ \bar{h}_k(0) \cdots \bar{h}_k(M-1) \}$ and $\bar{h}_k(l)$ is the l^{th} frequency tone of the channel experienced by the k^{th} user. Prior to symbol detection (demodulation), two signal processing tasks are first performed. The first task is an attempt to recover the channel-distorted chips. This is achieved through equalization that is performed on each individual chip on individual sub-carrier. After equalization, the next task is to convert the chips back into symbols by performing despreading operation. After despreading, symbol detection can then be carried out. While such equalization technique is suitable for downlink signal recovery, the same cannot be applied for uplink. Performing such equalization in the uplink will generate inter-chip interference and the subsequent performance degradation due to multi-user interference (MUI) component. The following analysis illustrates this problem.

For illustration purpose, assume that zero-forcing (ZF) technique is used. The ZF multiplier for the k_0^{th} user on the l^{th} tone is given as:

$$G_{ZF|k_0}(l) = \frac{\bar{h}_{k_0}^*(l)}{|\bar{h}_{k_0}(l)|^2} \quad (2.14)$$

After all the tones are multiplied with the respective ZF multipliers, the equalized signal can then be expressed as:

$$\bar{\mathbf{y}}_{k_0} = \sum_{k=1}^K \begin{bmatrix} \alpha_{k|k_0}(0) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \alpha_{k|k_0}(M-1) \end{bmatrix} \begin{bmatrix} c_{k,1}d_k \\ \vdots \\ c_{k,G}d_k \end{bmatrix} + \begin{bmatrix} G_{ZF|k_0}(0)\bar{n}(0) \\ \vdots \\ G_{ZF|k_0}(M-1)\bar{n}(M-1) \end{bmatrix} \quad (2.15)$$

where

$$\alpha_{k|k_0}(l) = \frac{\bar{h}_k(l)\bar{h}_{k_0}^*(l)}{|\bar{h}_{k_0}(l)|^2} \quad (2.16)$$

In downlink transmission, the information symbols belonging to all users propagate through a common or identical channel, i.e. $\bar{h}_k(l) \equiv \bar{h}(l) \forall k$ and l . This means that $\alpha_{k|k_0}(l) = 1 \forall k$ and hence, code orthogonality is guaranteed even after equalization.

In uplink transmission, the propagation channel experienced by symbols belonging to different users is no longer the same, i.e. $\bar{h}_1(l) \neq \dots \neq \bar{h}_K(l) \forall k$ and l . This makes $\alpha_{k|k_0} = 1$ only when $k = k_0$ and $\alpha_{k|k_0} \neq 1$ otherwise. In this case, the result of the despreading operation is given by:

$$\mathbf{y}_{k_0} = \underbrace{(\mathbf{c}_{k_0}^T \mathbf{c}_{k_0})}_1 d_{k_0} + \sum_{k=1, k \neq k_0}^K \underbrace{\left(\sum_{g=1}^G \alpha_{k|k_0}(g) c_{k_0,g} c_{k,g} \right)}_{\neq 0} d_k + \mathbf{G}_{ZF|k_0} \bar{\mathbf{n}} \quad (2.17)$$

where $\mathbf{G}_{ZF|k_0} = \text{diag} \{G_{ZF|k_0}(0) \dots G_{ZF|k_0}(M-1)\}$. From Equation (2.17), it can be seen that the code orthogonality is lost and hence the despreading operation is no longer effective in eliminating the signals belonging to users other than the desired one. This results in multiple-user interference (MUI) that will severely affect the system performance and hence limits the suitability of MC-CDMA for uplink deployment.

2.2.3 CDMA with Cyclic-Prefix Assisted Single-Carrier Transmission

This section describes a generic single-carrier based CDMA system that employs frequency-domain processing. First, single-carrier technique with frequency-domain equalization (SC-FDE) as the basis of any multicarrier system is first discussed. After that, architectural combination of SC-FDE with CDMA is presented.

2.2.3.1 SC-FDE: The Enabling Technology

While OFDM-based multicarrier system promises several advantages such as robustness against multipath impairments and efficient bandwidth utilization resulting from the creation of spectrally overlapping sub-channels, multicarrier systems are associated with two inherent problems. The first problem is the presence of high PAPR which affects the linear range of transmit amplifier. The second major consideration for adopting OFDM-based scheme is its sensitivity to RF phase noise caused by uncompensated CFO.

The search for better alternative to multicarrier system led to the use of block transmission structure with single-carrier transmission and frequency-domain processing (SC-FDE) [51]. SC-FDE scheme uses the basic principle of OFDM, that is to divide the entire bandwidth

into smaller parallel non-interfering sub-bands for simple equalization processing. A general SC-FDE architecture is shown in Figure 2.4. As can be seen in the Figure, a SC-FDE transmitter is largely identical to normal single-carrier based transmitter, except for the CP insertion during transmission. The insertion of CP is essential to allow frequency-domain processing at the receiver. The role of CP here is to convert the time-domain linear convolution between the channel and transmitted signals into circular convolution, while at the same time acting as guard interval to prevent IBI.

Figure 2.4: A General SC-FDE Transceiver Architecture

After CP removal, the received signal is subjected to FFT operation. The output of the FFT operation is the product of the frequency tones of the transmitted signal and the channel which can be mathematically expressed as

$$\begin{aligned} \bar{\mathbf{y}} &= \mathbf{F}_M \mathbf{H} \mathbf{F}_M^H \mathbf{F}_M \mathbf{x} + \mathbf{F}_M \mathbf{n} \\ &= \begin{bmatrix} \bar{h}(0) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \bar{h}(M-1) \end{bmatrix} \begin{bmatrix} \bar{x}(0) \\ \vdots \\ \bar{x}(M-1) \end{bmatrix} + \begin{bmatrix} \bar{n}(0) \\ \vdots \\ \bar{n}(M-1) \end{bmatrix} \end{aligned} \quad (2.18)$$

where $\mathbf{x} = [x(0) \cdots x(M-1)]^T$ is the vector of transmitted symbols, $\mathbf{H}$ is the $M \times M$ circular Toeplitz channel matrix, $\mathbf{F}_M$ is the M -point FFT matrix and $\mathbf{n}$ is the Additive White Gaussian Noise (AWGN). $\bar{x}(m)$, $\bar{h}(m)$ and $\bar{n}(m)$ corresponds to the m^{th} frequency tone of the transmitted signal, multipath channel and AWGN component respectively. Note that Eqn. (2.18) makes use of the property that circular Toeplitz channel matrix can be diagonalized by pre- and post-multiplying with FFT/IFFT matrices [50]. Equalization is then performed on each frequency tone individually. Notice that the equalized signals are

still in frequency domain and an IFFT operation is required to recover the transmitted signals in time-domain.

In comparison to its OFDM counterpart, SC-FDE offers several advantages. The structure of SC-FDE transmitter is relatively simple and largely similar to existing technology as no IFFT processor is required and thus doing away with the need of an expensive amplifier [52]. Since single-carrier transmission is employed, the impact due to uncompensated DCO and CFO is not substantial in SC-FDE compared to OFDM. Furthermore, it is highlighted in [53] that during equalization in SC-FDE, energy of individual bit is distributed over all subcarriers, thus making it more robust against deep narrowband fading channels. This is unlike multicarrier schemes where it is observed in [46] that error rate is dominated by symbols experiencing poor channel conditions. The receiver complexity of SC-FDE is slightly higher due to the presence of an additional IFFT operation to convert the symbols back into time-domain. Nevertheless, SC-FDE scheme is particularly attractive for uplink deployment where the Base Stations act as the receiver and higher complexity is more tolerable while it is more critical to keep the transmitter structure simple.

2.2.3.2 Cyclic-Prefix Assisted Single-Carrier CDMA (CP-CDMA)

As briefly discussed earlier, SC-FDE architecture appears to be an emerging alternative to OFDM-based multicarrier system due to several advantages such as simple transmitter structure, better performance in the presence of phase noise and CFO while retaining most of the advantages that comes with multicarrier systems. The simple transmitter structure makes SC-FDE a strong candidate for future uplink communication systems. It was in view of all these advantages that the combination of SC-FDE with CDMA attracted considerable attention among the research community.

Figure 2.5: A General CP-CDMA Transceiver Architecture

In [54], a cyclic-prefix assisted single-carrier CDMA scheme is proposed and commonly referred to as CP-CDMA system. Figure 2.5 illustrates a simplified baseband model of a general CP-CDMA system. From the Figure, it is obvious that a CP-CDMA transmitter is much simpler than a MC-CDMA transmitter. In a simplistic view, a CP-CDMA transmitter is nothing more than a DS-CDMA transmitter that is equipped with additional CP insertion module before symbol transmission occurs. The chip output resulting from the spreading operation is $\mathbf{x}_k = \mathbf{c}_k \otimes d_k$. In downlink transmitter, the chips from all users are first added together before CP is inserted. In uplink transmitter, CP is inserted immediately to the output of the spreading operation.

After CP is removed, the received signal is given by

$$\mathbf{r} = \sum_{k=1}^K \mathbf{H}_k \mathbf{x}_k + \mathbf{n} \quad (2.19)$$

where $\mathbf{H}_k$ is the circular Toeplitz channel matrix as defined previously. After performing FFT operation, the received signal now becomes

$$\begin{aligned} \bar{\mathbf{r}} &= \sum_{k=1}^K \mathbf{F}_M \mathbf{H}_k \mathbf{x}_k + \mathbf{F}_M \mathbf{n} \\ &= \sum_{k=1}^K \mathbf{F}_M \mathbf{H}_k \mathbf{F}_M^H \mathbf{F}_M \mathbf{x}_k + \bar{\mathbf{n}} \\ &= \sum_{k=1}^K \begin{bmatrix} \bar{h}_k(0) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \bar{h}_k(M-1) \end{bmatrix} \begin{bmatrix} \bar{x}_k(0) \\ \vdots \\ \bar{x}_k(M-1) \end{bmatrix} + \begin{bmatrix} \bar{n}(0) \\ \vdots \\ \bar{n}(M-1) \end{bmatrix} \end{aligned} \quad (2.20)$$

where $\bar{h}_k(l)$ is the l^{th} frequency tone of the channel experienced by the k^{th} user and $\bar{\mathbf{x}}_k = \mathbf{F}_M \mathbf{x}_k = [\bar{x}(0) \cdots \bar{x}(M-1)]^T$ is the frequency tone vector of the transmitted chips. Per-tone equalization (as in SC-FDE) is then performed followed by IFFT operation to transform the chips' frequency tones back into time-domain. Despreading operation is then performed to obtain the reconstructed symbols that will be used for detection.

As compared to MC-CDMA, the advantages offered by CP-CDMA system are evident. The first advantage is attributed to the single-carrier transmission scheme that CP-CDMA is leveraging on. This allows CP-CDMA to avoid issues related to multicarrier system such as PAPR and performance sensitivity due to DCO and CFO. At the same time, by employing

frequency-domain equalization technique, CP-CDMA is able to tap on the advantageous features of multicarrier systems such as robustness against channel with severe multipath impairments and the simple one-tap equalization process. The second advantage is the chip (instead of symbol) diversity that is enjoyed by CP-CDMA. It is highlighted in [46] that the presence of deep fading sub-channels in MC-CDMA may lead to some of the chips to be completely destroyed. This might disrupt the code orthogonality property which makes symbol recovery impossible. In such a system, the error performance in MC-CDMA system will be dominated by those chips in deep fading sub-channels. In CP-CDMA, the energy of each chip is distributed over all sub-carriers [53][16] due to the FFT operation and hence the system performance is not severely affected as in MC-CDMA. This means that each chip will now enjoy frequency-diversity and thus additional robustness against channels consisting of deep fade sub-bands.

2.3 Enhancement of System Performance Through MIMO Architecture

2.3.1 Single-Input-Single-Output (SISO) Architectures

In any wireless communication system equipped with a single antenna at both the transmitter and receiver, the maximum error-free transmission rate that can be achieved is limited by the *Shannon Limit* or *Shannon Capacity Theorem* given as:

$$C = W \log\left(1 + \frac{1}{W} \frac{P_{av}}{N_0}\right) \quad (2.21)$$

where C is the channel capacity defined as the number of bits transmitted per second (bits/s), W is the total bandwidth available for transmission, P_{av} is the average transmitted power and N_0 refers to the noise power spectral density.

The *Shannon Capacity* of a communication link against signal-to-noise ratio (SNR) for a bandlimited signal is shown in Figure 2.6. As shown in the Figure, the normalized system capacity grows logarithmically with increasing SNR when the transmission bandwidth is fixed. The above capacity theorem is first proposed in 1948 [55]. However, *Shannon* only showed the maximum possible transmission rate for reliable transmission, but not

the methodologies to achieve this. Since then, numerous error-correcting algorithms that attempt to deliver the *Shannon* capacity have been proposed. Some of the successful examples are Reed-Solomon Code [56], Turbo Code [57] and Low Density Parity Check (LDPC) Codes [58]. However, the challenge of realizing the ultimate limit still remains as of today. As some of these codes approach the maximum possible limit, the task of simplifying the algorithms for efficient implementation given the current hardware constraints and technological limitations, is yet another challenge on its own.

Figure 2.6: Normalized Channel Capacity

Instead of using a single communication link and trying to improve the error-rate, another way of enhancing system capacity is by introducing additional links in which transmission can take place. This fundamental philosophy leads to the use of multiple antennae at both the transmitter and the receiver that form multiple-input-multiple-output (MIMO) channels. The following sections describe such architecture and the different signal processing techniques that can be employed.

2.3.2 Multiple-Input-Multiple-Output (MIMO) Architectures

Future generation of mobile communication systems face numerous challenges. Some of the challenges include the need to support high-rate applications, good Quality of Service (QoS) that demands consistently good and reliable signal quality and large system capacity to support the ever-increasing pool of subscribers. This is despite the fact that frequency spectrum remains a scarce commodity while power consumption and device size remains a tightly constrained design issue.

In a non Line-of-Sight (non-LOS) propagation environment, there are times when the receiver cannot recover an information symbol that is affected by deep fade. One way to overcome this problem is to do an error-checking and then request for a re-transmission. Another more effective way is to increase the signal quality by providing the receiver with different replicas of the transmitted symbol. This can be done in two ways: 1. By using multiple antennae at the transmitter; and 2. By using multiple antennae at the receiver. However, it is important to note that such methods will work well only when the propagation channels between different pair of transmit and receive antennae are unique and independent to each other. In other words, the multiple antennae at the receiver or the transmitter need to be uncorrelated and hence separated by a minimal distance [59][60]. Due to this requirement and limitations in the handheld devices, it was only possible to build multiple antennae at the BS. In the past, major effort was focused in building a BS with receiver diversity, resulting in the creation of single-input-multiple-output (SIMO) system. It was only with recent advancement in Micro-Electro-Mechanical-Systems (MEMS) technology [61] that it is now possible to fit multiple antennae in portable devices at relatively low cost and without significant enlargement in the equipment size. This triggered an enormous interest in the use of multiple antennae at both the BS and remote units. When diversity is employed only at the transmitter, a multiple-input-single-output (MISO) system is created and when diversity is employed at both the transmitter and receiver, a multiple-input-multiple-output (MIMO) system is created.

2.3.2.1 Fundamentals of MIMO Systems

It was first shown in [62] that the formation of MIMO channels through the use of multiple transmit and receive antennae promises tremendous capacity improvement. The fundamen-

tal capacity limit of any MIMO system is still bounded by *Shannon Theorem* which is given as

$$\mathcal{C} = \mathcal{E}(\log_2(\det(\mathbf{I}_{N_R} + (\frac{\gamma}{N_T})\mathbf{H}\mathbf{H}^H))) \quad (\text{bits/s/Hz}) \quad (2.22)$$

where $\mathbf{I}_{N_R}$ is a $N_R \times N_R$ identity matrix, γ is the SNR at each antenna, N_R and N_T refers to the number of antenna at the receiver and transmitter respectively and $\det(\mathbf{A})$ refers to the determinant of matrix $\mathbf{A}$. Here, $\mathcal{E}(\cdot)$ refers to the expectation or mean. Under certain set of assumptions, the overall system capacity would be growing linearly, instead of logarithmically, with increasing SNR. However, in this thesis, the channel model that is most relevant and that has been used extensively is *Rayleigh* fading channel. Therefore, understanding the fundamental capacity limit for such channel is important. When each of the MIMO channel is assumed to be independent and *Rayleigh* distributed, Equation (2.22) now becomes [63]

$$\mathcal{C} = \int_0^\infty \log(1 + P\lambda/N_T) \sum_{k=0}^{m-1} \frac{k!}{(k+n-m)!} [L_k^{n-m}(\lambda)]^2 \lambda^{n-m} e^{-\lambda} d\lambda \quad (2.23)$$

where $m = \min\{N_R, N_T\}$ and $n = \max\{N_R, N_T\}$ and $L_j^i(\lambda)$ is the associated *Laguerre* as described in Appendix B. Here, P corresponds to the total power transmitted from all antennae and λ refers to all possible eigenvalues of $(\mathbf{H}\mathbf{H}^H)$. To obtain the solution for the above expression using direct computation is not a trivial problem. Instead, an approximate solution can be obtained by first using the logarithmic approximation $\log_2(1+x) \approx \sqrt{x}$ which leads to the following expression:

$$\begin{aligned} \mathcal{C} &= \sum_{k=0}^{m-1} \int_0^\infty \left(\frac{P}{N_T}\lambda\right)^{\frac{1}{2}} \frac{k!}{(k+n-m)!} L_k^{n-m}(\lambda) L_k^{n-m}(\lambda) \lambda^{n-m} e^{-\lambda} d\lambda \\ &= \sum_{k=0}^{m-1} \left(\frac{P}{N_T}\right)^{\frac{1}{2}} \frac{k!}{(k+n-m)!} \int_0^\infty \lambda^{n-m+\frac{1}{2}} L_k^{n-m}(\lambda) L_k^{n-m}(\lambda) e^{-\lambda} d\lambda \end{aligned} \quad (2.24)$$

According to [64] with regards to *Laguerre* polynomial, the following definite integral property is true:

$$\int_0^\infty t^{\alpha-1} e^{-pt} L_a^\beta(pt) L_b^\gamma(pt) dt = \frac{p^{-\alpha} \Gamma(\alpha) \Gamma(b-\alpha+\gamma+1) \Gamma(a+\beta+1)}{a! b! \Gamma(1-\alpha+\gamma) \Gamma(\beta+1)} \quad (2.25)$$

$${}_3F_2(-a, \alpha, \alpha-\gamma; -b+\alpha-\gamma, \beta+1; 1)$$

where $\Gamma(n)$ and ${}_pF_q(a_1, \dots, a_p; b_1, \dots, b_q; 1)$ is the gamma function and generalized hypergeometric function as defined in Appendix B. By substituting Equation (2.25) into (2.24), the capacity equation can be simplified to the following:

$$\begin{aligned}
 C &= \left(\frac{P}{N_T} \right)^{\frac{1}{2}} \sum_{k=0}^{m-1} \frac{k!}{(k+n-m)!} \frac{\Gamma(n-m+\frac{3}{2})\Gamma(k-n+m+\frac{3}{2}+n-m+1)\Gamma(k+n-m+1)}{(k!)^2\Gamma(1-n+m-\frac{3}{2}+n-m)\Gamma(n-m+1)} \\
 &\quad {}_3F_2\left(-k, n-m+\frac{3}{2}, n-m+\frac{3}{2}-n+m; -k+n-m+\frac{3}{2}-n+m, n-m+1; 1\right) \\
 &= \left(\frac{P}{N_T} \right)^{\frac{1}{2}} \sum_{k=0}^{m-1} \frac{k!}{(k+n-m)!} \frac{\Gamma(n-m+\frac{3}{2})\Gamma(k+\frac{3}{2}+1)\Gamma(k+n-m+1)}{(k!)^2\Gamma(1-n+m-\frac{3}{2}+n-m)\Gamma(n-m+1)} \\
 &\quad {}_3F_2\left(-k, n-m+\frac{3}{2}, \frac{3}{2}; -k+\frac{3}{2}, n-m+1; 1\right)
 \end{aligned} \tag{2.26}$$

Figure 2.7 plots this theoretical capacity bounds that are achievable using different number of N_R and N_T for different level of power constraints. Note that the order of diversity at both the transmitter and the receiver is assumed to be equal, i.e. $N_T = N_R$ and $n - m = 0$. Further works addressing this issue include [65][66][67][68][69][70]. In addition, the capacity of MIMO system with correlated channels has been studied extensively in [71][72][73][74][75][76][77].

Motivated by the capacity improvement promised due to the formation of MIMO channels, the research community has dedicated tremendous effort in designing a MIMO architecture that is capable of realizing the theoretical bound and yet structurally simple. Fundamentally, the availability of additional antenna at the transmitter can either be used to transmit data independently in parallel or to introduce spatial diversity by transmitting the same symbol from all antennae possibly over consecutive symbol periods. A general overview of these two architectures is given in the following, followed by a brief overview of the hybrid architecture that combines the two architectures together.

2.3.2.2 MIMO I: Layered Space-Time Coding

The term layered space-time (LST) is first coined by Foschini and the first such architecture is first proposed in [78]. The main idea behind LST design is to treat the stream of information symbols as layer. Previously when there is only one antenna available, there is only one pipe in which this layer of data can be transmitted. Now, with the availability of additional antennae, there is a flexibility as to how the data can be transmitted. In LST

Figure 2.7: Capacity of MIMO Systems with Independent Rayleigh Fading Channels

architecture, this single layer is broken down into multiple layers of data. At the first symbol period, only the first symbol of the first layer will be transmitted from the first antenna. At the second symbol period, the second symbol from the first layer is transmitted from the second antenna while the first symbol from the second layer is transmitted from the first antenna. Transmission during subsequent symbol periods follow the same pattern. In other words, each layer is continuously rotated and symbols from each layer are fed into different antenna each time. This results in the creation of diagonal-like structure as shown in Figure 2.8 and hence referred to as Diagonal Bell Labs lAyered space-time (DBLAST)

Figure 2.8: Transmission Pattern of a DBLAST Transceiver

architecture. The received signal model can be represented mathematically as

$$\begin{bmatrix} r_1 \\ r_2 \\ r_3 \\ \vdots \\ \vdots \end{bmatrix} = \begin{bmatrix} x_{1,1} & 0 & 0 & 0 & 0 & \cdots & \cdots \\ x_{2,1} & x_{1,2} & 0 & 0 & 0 & \ddots & \ddots \\ x_{3,1} & x_{2,2} & x_{1,3} & 0 & 0 & \ddots & \ddots \\ x_{4,1} & x_{3,2} & x_{2,3} & x_{1,4} & 0 & \ddots & \ddots \\ 0 & x_{4,2} & x_{3,3} & x_{2,4} & x_{1,5} & \ddots & \ddots \\ 0 & 0 & x_{4,3} & x_{3,4} & x_{2,5} & x_{1,6} & \ddots \\ 0 & 0 & 0 & x_{4,4} & x_{3,5} & x_{2,6} & \ddots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots \end{bmatrix} \begin{bmatrix} h_{q1}(0) \\ h_{q2}(0) \\ h_{q3}(0) \\ h_{q4}(0) \\ h_{q1}(1) \\ h_{q2}(1) \\ h_{q3}(1) \\ h_{q4}(1) \\ \vdots \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ n_3 \\ \vdots \\ \vdots \end{bmatrix} \quad (2.27)$$

where r_t is the received signal during t^{th} symbol period, $x_{p,t}$ is the t^{th} symbol belonging to the p^{th} layer, $h_p(t_1)$ is the channel pair between the receiver and p^{th} transmit antenna during $t_1 = \text{mod}(t, 4)$ block period and $\text{mod}(a, b)$ refers to the modulus division by b . For illustration purpose, the number of antenna (and thus layer) is assumed to be 4 and it has been assumed that the channel for each antenna remains stationary during 4 symbol periods. Based on Equation (2.27), DBLAST detection is normally carried out one layer after another and interference contribution from previously detected layers are eliminated.

Since data from each layer in a DBLAST architecture is transmitted from multiple antennae (instead of a single dedicated antenna), each layer enjoys spatial diversity and thus the overall system performance is not dependant on the channel condition of any particular transmit antenna. However, DBLAST transmitter is not very popular due to two major implementation issues. The first issue is the high processing complexity that is associated with DBLAST transmitter. The hardware structure involved to constantly rotate (or multiplex) the output of each layer to be fed into different RF chains and transmit antenna is non-trivial and expensive. The second issue is related to the detection process at the receiver. As detection is carried out one layer after another, the presence of interference from undetected layers causes certain degree of inaccuracy in the detection process. As such, the use of error-correcting codes at the transmitter is necessary to aid the overall performance detection.

Figure 2.9: A Generalized VBLAST Architecture

Vertical Bell Labs lAyered space-time (VBLAST) architecture is proposed in [79] as an alternative to DBLAST architecture. In VBLAST, each RF chain (transmit antenna) is dedicated to a single layer. Obviously, VBLAST offers a tradeoff between spatial diversity and processing complexity. While the error of a particular layer is now dependant on the channel quality between its associated transmit chain and the receiver, there is no need to perform the rotational (multiplexing) operation at the transmitter.

A generalized VBLAST architecture is shown in Figure 2.9. For illustration purpose, assume that there are 4 symbols to be transmitted and there are 4 antennae ($N_T = 4$) available for transmission. The vertical multiplexer takes in the single stream of symbols and outputs them onto 4 independent sub-streams for 4 different RF chains, where symbol on the p^{th} sub-stream is denoted by $x_p, p \in [1, N_T]$. To begin with, assume that the channel is flat-

fading. The received signal is given by

$$\begin{aligned} \mathbf{r} &= \begin{bmatrix} h_{11} & \cdots & \cdots & h_{1N_T} \\ \vdots & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ h_{N_R1} & \cdots & \cdots & h_{N_RN_T} \end{bmatrix} \begin{bmatrix} x_1 \\ \vdots \\ \vdots \\ x_{N_T} \end{bmatrix} + \begin{bmatrix} n_1 \\ \vdots \\ \vdots \\ n_{N_R} \end{bmatrix} \\ &= \mathbf{H}\mathbf{x} + \mathbf{n} \end{aligned} \quad (2.28)$$

where N_R corresponds to the number of receive antenna, h_{qp} is the channel between the q^{th} receive and p^{th} transmit antenna and n_q is the AWGN component on the q^{th} receive antenna. There are few different ways of processing the received signal.

The first processing technique is the maximum likelihood detection (MLD) algorithm which has been widely recognized as the optimum solution for such detection problem. The detection problem arising from such flat-fading VBLAST channel is similar to the multiuser detection problem as first discussed in [80] and can be expressed as

$$\hat{\mathbf{x}} = \underset{\mathbf{x}}{\arg \min} \|\mathbf{r} - \mathbf{H}\mathbf{x}\|^2 \quad (2.29)$$

where $\hat{\mathbf{x}} = [\hat{x}_1 \cdots \hat{x}_{N_T}]^T$ is the estimated symbol vector. Though MLD offers the optimum performance, it is rarely adopted as a practical solution because of its higher computation complexity. The complexity cost associated with M -ary modulation scheme is $2^{N_T(M-1)}$. Instead, various sub-optimum approaches have been proposed. The sub-optimum methods can be classified into either linear or non-linear technique. For linear technique, the received signal in Equation (2.28) is pre-multiplied with equalization matrix to produce the test statistics for symbol detection to be carried out. Two most popular equalization techniques are ZF or MMSE algorithm in which the multiplier matrix is given by

$$\mathbf{G}_{ZF} = \mathbf{H}^+ \quad \mathbf{G}_{MMSE} = (\mathbf{H}^H \mathbf{H} + \sigma^2 \mathbf{I})^{-1} \mathbf{H}^H \quad (2.30)$$

where $[\cdot]^+$ and $[\cdot]^H$ corresponds to pseudoinverse and Hermitian matrix operation respectively. Another commonly used linear equalization method is based on QR decomposition technique [81][82][83]. The fundamental principle behind QR decomposition technique is

that any given matrices, such as the $N_R \times N_T$ channel matrix $\mathbf{H}$, can be decomposed into

$$\mathbf{H} = \mathbf{Q}\mathbf{R} \quad (2.31)$$

where $\mathbf{Q}$ is a $N_R \times N_R$ orthonormal matrix (i.e. $\mathbf{Q}^T\mathbf{Q} = \mathbf{I}$) and $\mathbf{R}$ is a $N_R \times N_T$ with the first N_T rows forming an upper triangular matrix and the remaining $N_R - N_T$ rows consisting of all 0. Pre-multiplying the received signal $\mathbf{r}$ with $\mathbf{Q}^T$ results in

$$\begin{aligned} \mathbf{y} &= \mathbf{Q}^T \mathbf{r} \\ &= \mathbf{Q}^T (\mathbf{Q}\mathbf{R})\mathbf{x} + \mathbf{Q}^T \mathbf{n} \\ &= \mathbf{R}\mathbf{x} + \mathbf{Q}^T \mathbf{n} \\ &= \begin{bmatrix} \alpha_{11} & \alpha_{12} & \cdots & \alpha_{1N_T} \\ 0 & \alpha_{22} & \cdots & \alpha_{2N_T} \\ \vdots & \vdots & \ddots & \vdots \\ \vdots & \vdots & \ddots & \alpha_{N_T N_T} \\ 0 & 0 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & 0 \end{bmatrix} \begin{bmatrix} x_1 \\ \vdots \\ x_{N_T} \end{bmatrix} + \mathbf{Q}^T \mathbf{n} \end{aligned} \quad (2.32)$$

Detection can then be carried out layer-by-layer, starting from the N_T^{th} layer which will then be subtracted during subsequent layer detection.

The second approach is based on non-linear algorithm. Instead of detecting all layers at one time, detection is carried out iteratively one layer at a time. The iterative VBLAST detection is given by the following procedures:

1. $\mathbf{r}_0 = \mathbf{r}$, $\hat{a} = 0$ and $[\mathbf{H}]_0 = \mathbf{0}$
2. for $i = 1$ to I ,
3. for $l = 1$ to N_T ,
4. $\mathbf{r}_l = \mathbf{r}_{l-1} - \hat{a}_{l-1} [\mathbf{H}]_{l-1}$
5. $y_l = \mathbf{w}_l^T \mathbf{r}_l$
6. $\hat{a}_l = \text{demod} \langle y_l \rangle$
7. end
6. end

where I is the number of iteration stage to be carried out and $\text{demod}(\cdot)$ is the demodulation function that depends on the modulation scheme used. The weight vector $\mathbf{w}_l$ depends on the equalization technique used. When ZF or MMSE technique is used, $\mathbf{w}_l$ is simply the l^{th} row of $\mathbf{G}_{ZF}$ or $\mathbf{G}_{MMSE}$ as given in Equation (2.30). However, the key criteria for iterative VBLAST algorithm to work well is the detection ordering that is determined based on the channel rank information. Detection order starts from the transmit antenna with the best channel condition. When no detection ordering is considered, significant performance degradation is likely to occur.

The original VBLAST architecture proposed in [79] considers only flat-fading MIMO channels. Since multicarrier system is robust against channel with severe multipath impairment, VBLAST OFDM [84][85][86][87] has been proposed to increase the system capacity of multicarrier systems. After CP removal and FFT operation, the received signal is given as

$$\begin{aligned} \bar{\mathbf{r}} &= \begin{bmatrix} \mathbf{F}_M \mathbf{H}_{11} \mathbf{F}_M^H & \cdots & \mathbf{H}_{1N_T} \mathbf{F}_M^H \\ \vdots & \ddots & \vdots \\ \mathbf{F}_M \mathbf{H}_{N_R1} \mathbf{F}_M^H & \cdots & \mathbf{F}_M \mathbf{H}_{N_R N_T} \mathbf{F}_M^H \end{bmatrix} \begin{bmatrix} \mathbf{x}_1 \\ \vdots \\ \mathbf{x}_{N_T} \end{bmatrix} + \begin{bmatrix} \mathbf{F}_M \mathbf{n}_1 \\ \vdots \\ \mathbf{F}_M \mathbf{n}_{N_R} \end{bmatrix} \\ &= \begin{bmatrix} \bar{\mathbf{H}}_{11} & \cdots & \bar{\mathbf{H}}_{1N_T} \\ \vdots & \ddots & \vdots \\ \bar{\mathbf{H}}_{N_R1} & \cdots & \bar{\mathbf{H}}_{N_R N_T} \end{bmatrix} \begin{bmatrix} \mathbf{x}_1 \\ \vdots \\ \mathbf{x}_{N_T} \end{bmatrix} + \begin{bmatrix} \bar{\mathbf{n}}_1 \\ \vdots \\ \bar{\mathbf{n}}_{N_R} \end{bmatrix} \end{aligned} \quad (2.33)$$

where $\mathbf{H}_{qp}$ is the $M \times M$ circular Toeplitz channel matrix between the q^{th} receive and p^{th} transmit antenna, $\mathbf{x}_p$ is the $M \times 1$ symbol vector transmitted from the p^{th} antenna and $\mathbf{n}_q$ is the $M \times 1$ AWGN component on the q^{th} receive antenna. After FFT transformation, $\bar{\mathbf{H}}_{qp}$ represents the diagonal frequency channel response of $\mathbf{H}_{qp}$ and $\bar{\mathbf{n}}_q$ is the frequency response of the AWGN component $\mathbf{n}_q$.

It can be seen that performing MLD detection in VBLAST OFDM is practically impossible since exhaustive search on $2^{N_T(N-1)M}$ possible sequences is required. Alternatively, linear equalization such as in the flat-fading case can be performed. However, as ZF and MMSE techniques involve matrix multiplication and the processing complexity of computing the inverse of a matrix of size $n \times n$ is $\mathcal{O}(n^3)$, the processing complexity incurred would be $\mathcal{O}((N_T M)^3)$. In light of this problem, per-tone equalization method [84] that offer significant complexity reduction at the cost of slight performance degradation has been proposed. Per-tone equalization is performed by collecting the l^{th} taps of the FFT output from N_R

antennae and stack them as

$$\bar{\mathbf{r}}(l) = \begin{bmatrix} \bar{h}_{11}(l) & \cdots & \bar{h}_{1N_T}(l) \\ \vdots & \ddots & \vdots \\ \bar{h}_{N_R1}(l) & \cdots & \bar{h}_{N_RN_T}(l) \end{bmatrix} \begin{bmatrix} x_1(l) \\ \vdots \\ x_{N_T}(l) \end{bmatrix} + \begin{bmatrix} \bar{n}_1(l) \\ \vdots \\ \bar{n}_{N_R}(l) \end{bmatrix} \quad (2.34)$$

Given the earlier assumption that $N_R = N_T$, the channel matrix to be equalized on individual subcarrier is now a square matrix of size $N_T \times N_T$. Hence, the computation cost required to obtain the inverse matrix for each subcarrier is now $\mathcal{O}((N_T)^3)$ or a total of $M\mathcal{O}((N_T)^3)$ which is a linear function of the block size used.

It is highlighted in [88] that VBLAST transmission is not capable of exploiting the full receiver diversity. As such, [88] proposes the construction of lattice code [89][90] based on the received signals and then used sphere decoding (SD) technique to detect the transmitted information symbols. It is shown that at the cost of slight increase in processing complexity, it is possible to achieve considerable performance improvement and full diversity order.

One fundamental disadvantage associated with LST MIMO architecture is that the number of antenna employed at the receiver needs to be equal to or more than the number of antenna at the transmitter. Even though it is possible to equip a BS transmitter with as many antennae as possible in order to enhance the transmission rate, the maximum number of antenna is strictly limited by the size and power constraints of mobile devices.

2.3.2.3 MIMO II: Space-Time Block Coding

The main motivation behind space-time coding is to improve signal transmission quality instead of relying on the receiver diversity to improve reception quality. The first approach to achieve transmitter diversity is by transmitting copies of the same symbol through multiple antennae at different times. Few such architectures can be found in [91][92][93][94]. Based on similar delay diversity concept, space-time trellis code (STTC) is proposed in [95][96][97][98][99] that effectively combines the benefits of error-correcting codes with transmitter diversity to achieve performance gains. Though STTC is able to exploit both temporal and spatial diversity, the maximum likelihood decoder (MLD) at the receiver grows exponentially as a function of the bandwidth efficiency and the diversity order [100].

This problem leads to the second approach that is commonly referred to as space-time

block code (STBC). It is Alamouti [101] who first proposed a remarkable scheme to exploit transmitter diversity with 2 antennae without incurring expensive processing complexity at both the transmitter and receiver. The Alamouti scheme is depicted in Fig 2.10. The basic idea is to transmit two symbols s_1 and s_2 using two transmit antennae over two consecutive symbol intervals. During the first time epoch, s_1 and s_2 are transmitted from antenna 1 and 2 respectively. During the second time epoch, $-s_2^*$ and s_1^* are transmitted. The signal received during the two time epochs is represented by r_1 and r_2 . To recover s_1 and s_2 , only a simple maximal ratio combining (MRC) operation is required. The test statistics y_1 and y_2 (for s_1 and s_2 respectively) obtained as a result of the MRC operation are given by

$$\begin{aligned} y_1 &= h_1^* r_1 + h_2 r_2^* = (|h_1|^2 + |h_2|^2) s_1 \\ y_2 &= h_2^* r_1 - h_1 r_2^* = (|h_1|^2 + |h_2|^2) s_2 \end{aligned} \quad (2.35)$$

where h_1 and h_2 are the channel gains of the first and second antenna respectively. Maximum likelihood detection (MLD) can then be carried out to recover the transmitted information. In [102], a method was devised that generalized the construction of orthogonal codes for system with higher diversity order. Unfortunately, the effective transmission rate drops when transmitter diversity is greater than two. In STBC, the transmit rate is defined as the number of information symbols transmitted in a block and the number of symbol periods required to transmit the entire block. It was shown in [102] that rate-1 STBC for complex constellation is only possible with diversity order of two and the rate drops to $\frac{3}{4}$ or $\frac{1}{2}$ for higher diversity order. In general, it has been shown that given a transmit diversity order of $n = 2m - 1$ or $n = 2m$, the maximum effective transmission rate is $\frac{m+1}{2m}$. Theoretical study on the existence of rate 1 orthogonal code with general diversity order is further addressed in [103]. Since then, the problem of constructing optimal orthogonal codes attracted enormous interest. In [104][105][106][107][108], the use of algebraic approach to construct STBC has been addressed. Meanwhile, [109] proposes the use of number theory to construct STBC. Alternatively, the use of quasi-orthogonal code has been proposed in order to minimize the loss in data rate [110][111]. The improvement in the rate is traded off with other advantages such as diversity and processing complexity. For example, decorrelation operation in the received signal is usually required before detection is carried out and such non-orthogonal design usually requires the MIMO channels to be weakly correlated.

It has been assumed so far that the MIMO channels are flat-fading. For mobile communication system experiencing frequency-selective fading channels, Alamouti-coded multicarrier

Figure 2.10: Alamouti's Simple Transmit Diversity Scheme with 2 antennae

system is proposed in [112] to further enhance the system performance of OFDM-based multicarrier system. Now, let $\mathbf{d}_1 = [d_1(0) \cdots d_1(M-1)]^T$ and $\mathbf{d}_2 = [d_2(0) \cdots d_2(M-1)]^T$ be the symbol data vectors. The STBC encoding outputs the following matrix

$$\begin{bmatrix} \mathbf{x}_1^1 & \mathbf{x}_1^2 \\ \mathbf{x}_2^1 & \mathbf{x}_2^2 \end{bmatrix} = \begin{bmatrix} \mathbf{d}_1 & -\mathbf{d}_2^* \\ \mathbf{d}_2 & \mathbf{d}_1^* \end{bmatrix} \quad (2.36)$$

where $\mathbf{x}_p^t$ is the encoded vector to be transmitted from the p^{th} antenna during the t^{th} symbol period and $(p, t) \in \{1, 2\}$. FFT operation is then performed on the encoded vectors followed by CP insertion before transmission.

After CP removal and OFDM demodulation, the received signal is given by:

$$\begin{aligned} \bar{\mathbf{r}}^1 &= \mathbf{F}_M \mathbf{H}_1 \mathbf{F}_M^H \mathbf{d}_1 + \mathbf{F}_M \mathbf{H}_2 \mathbf{F}_M^H \mathbf{d}_2 = \bar{\mathbf{H}}_1 \mathbf{d}_1 + \bar{\mathbf{H}}_2 \mathbf{d}_2 \\ \bar{\mathbf{r}}^2 &= -\mathbf{F}_M \mathbf{H}_1 \mathbf{F}_M^H \mathbf{d}_2^* + \mathbf{F}_M \mathbf{H}_2 \mathbf{F}_M^H \mathbf{d}_1^* = -\bar{\mathbf{H}}_1 \mathbf{d}_2^* + \bar{\mathbf{H}}_2 \mathbf{d}_1^* \end{aligned} \quad (2.37)$$

where $\bar{\mathbf{H}}_p = \text{diag}(\bar{h}_p(0) \cdots \bar{h}_p(M-1))$ and $\bar{h}_p(l)$ is the l^{th} tone of the channel frequency response from the p^{th} transmit antenna. Per-carrier equalization can then be performed.

The issue of system capacity in STBC system is addressed in [70] and it is shown that a loss of capacity occurs when two or more receive antennae are employed. As such, an architecture that concatenates STTC as an outer code to STBC is proposed in [113][100]. Theoretical analysis comparing STBC and STTC are discussed in [114]. Hybrid architectures that attempt to combine the advantages of both STBC and STTC have also been proposed [113]

Figure 2.11: Multi-Layer STBC Architecture

2.3.2.4 MIMO III: Hybrid Multi-Layer Space-Time Block Codes

Based on earlier discussion, it is clear that the two MIMO architectures namely, LST and STBC both aim to enhance overall system capacity using different strategies in exploiting the additional available antenna. In LST architecture, direct enhancement in the transmission rate is achieved by transmitting different data in parallel. However, error contribution from the inter-layer interference limits the choice of modulation technique that can be used. On the other hand, STBC architecture accomplishes error performance improvement through the exploitation of spatial transmitter diversity and thus allowing higher modulation scheme (such as higher QAM) to be used and hence delivering higher overall transmission rate. However, rate-1 orthogonal STBC is only possible with two transmit antennae and the rate drops to $\frac{3}{4}$ with four antennae and to $\frac{1}{2}$ for higher number of antennae.

It is in this context that a hybrid version of these two architectures is proposed in [115][116][117] and is referred to as multi-layer STBC. The general multi-layer STBC architecture is shown in Figure 2.11 where for simplicity, the number of transmit and receive antennae are assumed as four and two respectively. The four transmit antennae are grouped into two groups or essentially divided into two layers. For each layer, the normal STBC scheme such

as Alamouti scheme is employed. This means that the symbols transmitted from antenna 1, 2, 3 and 4 during the first symbol epoch are s_1 , s_2 , s_3 and s_4 respectively. During the second symbol epoch, $-s_2^*$, s_1^* , $-s_4^*$ and s_3^* are transmitted respectively. The received signal on the q^{th} antenna (r_q^1 and r_q^2) during the first and second symbol epoch is given as

$$\begin{aligned} r_q^1 &= [h_{q1}s_1 + h_{q2}s_2] + [h_{q3}s_3 + h_{q4}s_4] + n_q^1 \\ r_q^2 &= [-h_{q1}s_2^* + h_{q2}s_1^*] + [-h_{q3}s_4^* + h_{q4}s_3^*] + n_q^2 \end{aligned} \quad (2.38)$$

where h_{qp} is the channel between the q^{th} receive and p^{th} transmit antenna and n_q^t is the additive white Gaussian noise (AWGN) component on the q^{th} receive antenna during the t^{th} symbol epoch. The channel is assumed to be flat-fading and stationary during the two symbol epochs. The received signal from N_R receive antennae is then stacked as

$$\begin{aligned} \mathbf{r} &= [r_1^1 \quad \cdots \quad r_{N_R}^1 \quad r_1^2 \quad \cdots \quad r_{N_R}^2]^T \\ &= \begin{bmatrix} h_{11} & h_{12} & h_{13} & h_{14} \\ \vdots & \vdots & \vdots & \vdots \\ h_{N_R1} & h_{N_R2} & h_{N_R3} & h_{N_R4} \\ h_{12}^* & -h_{11}^* & h_{14}^* & -h_{13}^* \\ \vdots & \vdots & \vdots & \vdots \\ h_{N_R2}^* & -h_{N_R1}^* & h_{N_R4}^* & -h_{N_R3}^* \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} \end{aligned} \quad (2.39)$$

Based on Equation (2.39), various equalization techniques such as ZF, MMSE or eigen-value decomposition-based algorithm can be used to recover the transmitted symbols.

At this point, it can be seen that such hybrid architecture offers a compromise between diversity and rate enhancement. Instead of transmitting 8 different symbols during two epochs, only 4 symbols are transmitted. From LST perspective, the effective transmission rate is only half of what is possible when pure LST architecture is used. Meanwhile, recall that by STBC definition, a transmission rate-1 is achieved when two symbols are transmitted on two time epochs ($N_T = 2$). This essentially means that by transmitting four symbols over two time epochs, transmission rate of 2 is achieved with the hybrid architecture. Another advantage of such an architecture is that the minimum number of antennae required at the receiver is only two instead of four as in conventional LST architecture.

2.4 Summary

In this chapter, a literature review of some of the technologies that enable today's wireless communication systems is presented. It appears highly likely that spread spectrum multiple access (such as CDMA) will be the critical building block in future mobile communication systems. This is due to the inherent advantages such as efficient bandwidth sharing, capability to support asymmetric traffic between up and down-link and etc. The choice between single-carrier and multicarrier transmission is still very much debatable. While multicarrier approach offers robustness against multipath channel, it comes with the price of requiring inefficient power amplifier at the transmitter and higher sensitivity to carrier frequency offset. On the other hand, single-carrier approach has been receiving serious attention and has been considered in some of the emerging industry standards. Lastly, it is obvious that to deliver the kind of data rate expected for future mobile communication systems, the use of multiple antennae at both the transmitter and receiver is necessary in order to boost the spectral efficiency. The additional antennae can be used directly to enhance the transmission rate by transmitting data streams parallelly or by enhancing the system's error performance and thereby exploit higher modulation schemes to improve system throughput.

Chapter 3

A Novel Architecture for CDMA Systems: Block Spread CDMA (BS-CDMA)

3.1 Introduction

A brief survey on CDMA and its different variations have been discussed in chapter 2. the evolution of CDMA as the preferred multiple access scheme is presented. The availability of a novel CDMA processing paradigm is critical in enabling mobile broadband systems to compete with the fixed-wired counterpart for transmitting high-rate multimedia data. A novel architecture called block-spread CDMA (BS-CDMA) is proposed as an alternative to existing conventional approaches, in particular for mobile communication systems that require high data rate capability.

Chapter 3 is organized as follows. Section 3.2 begins by highlighting some of the major limitations encountered by today's conventional CDMA schemes and how Block Spread CDMA (BS-CDMA) (the focus of this thesis) attempts to overcome those challenges. Section 3.3 presents a general BS-CDMA transceiver architecture. Section 3.4 presents a theoretical framework that is used to obtain the theoretical performance benchmark. Section 3.5 presents the details of the investigation exercise and results obtained through simulation studies.

3.2 Motivation

CDMA has been accepted worldwide as an efficient multiple access scheme over conventional methods such as TDMA and FDMA and has been adopted in today's 3G standards due to its numerous advantages. The rapidly evolving mobile applications are shifting towards high bandwidth and this has a broad implication on the overall system design. Firstly, there is a growing demand for packet-based switching in addition to the circuit-based switching that was sufficient to support voice-only communication in the past. Secondly, there is a need to support asymmetric traffic between the uplink and downlink transmission. Moreover, dynamic resource allocation to facilitate on-demand services is essential to allow profit-optimization for mobile communication providers. Clearly, TDMA/FDMA are not suited to these needs as resource allocation (time or frequency) is static. In CDMA, dynamic resource allocation can be achieved using various mechanisms such as the use of Variable Orthogonal Spreading Factor (VOSF) and multi-code assignment [14][11]. Other advantages offered by CDMA include simpler frequency reuse planning, inherent privacy, stronger resistance against jamming etc.

Despite the benefits that have been highlighted, CDMA system has its own set of challenges. For example, a DS-CDMA receiver with time-domain processing typically employs RAKE receiver. However, designing a good RAKE receiver for high rate communication system is a challenging task. In high-rate systems, the channel is usually characterized by severe multipaths with long delay spread. This means that ISI usually spans across large number of symbols. On one hand, since each RAKE finger corresponds to a single resolvable path, employing large number of such fingers is required to resolve the large number of paths. On the other hand, equipping a RAKE receiver with high number of fingers leads to higher degree of inter-finger interference (IFI) that would cause significant degradation in the overall system performance. Furthermore, the maximum number of fingers that can be built in a RAKE receiver is limited by the hardware constraints. Meanwhile, it is widely accepted that the suitability of DS-CDMA for uplink deployment is limited. This is due to the fact that chips from the different users now undergo different delay and attenuation. As a result, the simple receiver structure such as RAKE equalizer is no longer effective to suppress the MUI that arises due to the loss in code orthogonality. This requires a more sophisticated joint multi-user detection in order to achieve good symbol recovery.

Meanwhile, MC-CDMA has been proposed as an alternative to DS-CDMA in frequency selective channels as such. However, being an OFDM-based architecture, MC-CDMA is subjected to high PAPR and greater sensitivity to CFO. In [118], the advantage of SC-based system compared to OFDM-based system are measured in terms of the number of bits required by ADC/DAC and the efficiency of the power amplifier. For example, an OFDM system employing 64-QAM modulation requires 8-bits ADC/DAC while its single-carrier counterpart employing the same modulation scheme requires only 3-bits. In terms of power amplifier efficiency, single-carrier system outperforms its OFDM-based counterpart by as much as 4 times. In this context, CP-CDMA has been proposed as a suitable alternative due to its inherent capability in retaining the advantages of MC-CDMA and alleviating the problems that plague multi-carrier systems.

It is highlighted in [16] that the above conventional CDMA techniques do not perform well during asynchronous uplink transmission. During uplink transmission, mobile devices of different users are assumed to be scattered at different geographical locations and since CDMA users are allowed to transmit all the time, it is normally assumed that there is little or no coordination among users. As symbols are transmitted asynchronously and information signal transmitted from different users propagate through different paths to reach a common base station (BS), the channel characteristics of each user such as the attenuation factor and delay spread profile is unique. During such asynchronous reception, received chips from different users are likely to be misaligned which results in the destruction of code orthogonality. As a result, signals from different users are indistinguishable from one another and hence performance degradation is inevitable as symbol detection is now carried out on symbols that are corrupted by multi-user interference (MUI).

Research efforts directed at mitigating the effect of MUI in uplink CDMA systems have focused on several major areas, namely code waveform design and multi-user detection algorithms [119]. In practical channel environment, the degree of asynchronism is varying and unpredictable. The optimum code design is one that maintains orthogonality over all possible delays. Clearly, designing orthogonal set of codes with such a property is not possible. Sub-optimal code design that is nearly orthogonal in synchronous scenario and possesses low cross-correlation property during asynchronous scenario has been proposed in [120]. Such a design is referred to as quasi-orthogonal design. The second major approach is to devise an advanced algorithm that is capable of reducing the presence of MUI. Such detectors are called the optimum multi-user detectors. They are based on joint maximum-likelihood

detection [80]. However, the detection complexity grows exponentially as a function of the total number of users and hence, is prohibitively complex. Due to this reason, various sub-optimal algorithms have been proposed. The first class of such algorithm exploits higher order statistics to design a filter that is capable of multi-user identification even in asynchronous multipath channels [121][122]. Another class of sub-optimal algorithm is based on subspace-based approach to carry out the detection [123][124][125]. Alternatively, multi-user detector that is based on full or partial feedback has been proposed in [126][127][128].

Recently, another alternative approach for overcoming the degradation impact due to MUI has been proposed. It is shown in [129][130] that by performing chip-interleaving, the resulting transmission structure is robust against MUI even in the presence of frequency-selective channels. A simplified block spread CDMA (BS-CDMA) system that leverages on the block transmission structure created by the chip-interleaving is proposed in [131]. The main novelty in BS-CDMA stems from the observation that the transmission structure created as a result of the chip-interleaving operation [131] can be exploited to eliminate MUI and reduce the processing complexity at the receiver. In BS-CDMA, the chip transmission sequence maintains code alignment even during asynchronous reception and hence preserving orthogonality between the codes. It is thus possible to obtain nearly MUI-free performance during certain operating environment. At BS-CDMA receiver, the processing sequence is re-arranged such that block despreading operation is performed prior to other processing such as FFT/IFFT and channel equalization. This leads to a significant reduction in the processing complexity. Detailed description of the BS-CDMA transceiver structure is presented in the remaining part of this chapter.

3.3 System Model

Figure 3.1: Generalized Baseband BS-CDMA Transceiver Structure

The general BS-CDMA architecture is illustrated in Figure 3.1. In BS-CDMA system,

transmission of information symbols leverages on the single-carrier (SC) scheme. This allows a BS-CDMA transmitter to inherit two distinct advantages. The first advantage is its ability to use the existing SC-based transmission technology which has been proven. The second advantage is a simplified transmitter architecture due to the absence of signal processing modules such as IFFT operation and other pre-transmission modules for multicarrier systems. As shown in Figure 3.1, this makes a BS-CDMA transmitter simple.

Now, the main difference between a BS-CDMA transmitter and a CP-CDMA or DS-CDMA (conventional SC-based CDMA schemes) transmitter lies in the way the spreading operation is carried out. A detailed description of the end-to-end processing in BS-CDMA system is elaborated as follows.

Instead of spreading a single information bit one at a time, BS-CDMA spreading operation takes a group of symbols and then performs the spreading operation collectively. In other words, the information bit $d_k(n)$ belonging to the k^{th} user is first grouped into blocks of size M , $\mathbf{d}_k = [d_k(iM) \cdots d_k(iM + M - 1)]^T$. Each of these blocks is then fed into a BS-CDMA spreader which produces the following $G \times M$ chip matrix

$$\begin{aligned} \mathbf{X}_k &= \mathbf{c}_k \otimes \mathbf{d}_k^T \\ &= \begin{bmatrix} d_k(1)c_{k,1} & \cdots & d_k(M)c_{k,1} \\ \vdots & \ddots & \vdots \\ d_k(1)c_{k,G} & \cdots & d_k(M)c_{k,G} \end{bmatrix} \end{aligned} \quad (3.1)$$

where $\mathbf{c}_k = [c_{k,1} \cdots c_{k,G}]^T$ is k^{th} user's spreading code, c_{kg} is the g^{th} chip of the code and G corresponds to the spreading factor and $\mathbf{X}_k$ is a $G \times M$ chip matrix. The chips are then read out row-by-row and each row is defined as one chip-block, $\mathbf{x}_k(g)$

$$\mathbf{x}_k(g) = [d_k(1)c_{k,g} \cdots d_k(M)c_{k,g}]^T \quad (3.2)$$

Cyclic-prefix (CP) is then inserted by duplicating the last L_{CP} chip entry of $\mathbf{x}_{k,g}$ and then appending them as entries to create the new CP-inserted chip-block, $\mathbf{x}_{k,g}^{(CP)}$ as

$$\mathbf{x}_{k,g}^{(CP)} = [\mathbf{I}_M^{(CP)T} \quad \mathbf{I}_M]^T \mathbf{x}_k(g) \quad (3.3)$$

where $\mathbf{I}_M^{(CP)}$ is an $L_{CP} \times M$ matrix that is made up of the last L_{CP} rows of $\mathbf{I}_M$ and $\mathbf{I}_M$ is a

$M \times M$ identity matrix. The role of these guard chips is to prevent inter-block interference (IBI) and to facilitate the frequency-domain processing at the receiver. Transmission is then carried out block-by-block, one chip-block after another. Another way of looking at a BS-

(a) Block Spreading Operation

(b) Block Transmission Structure after Parallel-to-Serial Conversion

Figure 3.2: Block Spread CDMA Spreading and Transmission Structure

CDMA spreader is by performing chip-interleaving to a conventional DS-CDMA spreader. This is illustrated in Figure 3.2. In DS-CDMA system, chips from a particular symbol are transmitted consecutively one after another and the transmitted sequence of chips from M symbols is given as

$$\begin{aligned}
 \mathbf{x}_{k|DS-CDMA} &= \left[d_k(1)c_{k,1} \quad \dots \quad d_k(1)c_{k,G} \quad \dots \quad d_k(M)c_{k,1} \quad \dots \quad d_k(M)c_{k,G} \right]^T \\
 &= \left[\mathbf{x}_{k,1|DS-CDMA}^T \quad \dots \quad \mathbf{x}_{k,M|DS-CDMA}^T \right]^T
 \end{aligned} \tag{3.4}$$

From Equation (3.4), it can be seen that each mini-block in DS-CDMA, $\mathbf{x}_{k,m|DS-CDMA} = \left[d_k(m)c_{k,1} \quad \dots \quad d_k(m)c_{k,G} \right]^T$ now consists of all the G chips belonging to the m^{th} symbol. In high-rate wireless communication systems, the channel is typically characterized by large number of paths and long delay spread. As chips from a symbol are transmitted consecutively, the presence of multipath gives rise to inter-symbol interference. This self-imposed interference might lead to significant performance degradation.

Unlike DS-CDMA, CP-CDMA and MC-CDMA, BS-CDMA transmits a single chip (say

the g^{th} chip) from M symbols consecutively. The fundamental principle behind BS-CDMA originates from the interesting observation highlighted in [131]. In high-rate CDMA communication system, the chip duration turns out to be much smaller than the channel coherence time, T_{CO} . According to [132], the channel coherence time can be approximated by the formula, $T_{CO} = \frac{9}{16\pi F_D}$, where F_D corresponds to the Doppler frequency [133] that indicates the mobility condition. Based on this formula, the channel coherence time for a Doppler frequency of 40Hz and 200Hz (low to medium vehicular mobility), the channel coherence time is found to be 4.5msecs and 0.9msecs respectively. On the other hand, given a typical mobile communication system with a chip rate of 4Mcps, the chip duration is only 250nsecs.

$$\begin{aligned}
 & \sum_{k=1}^K \mathbf{h}_k \odot \begin{array}{|c|c|c|} \hline d_k(0)c_{k1} & \dots & d_k(M-1)c_{k1} \\ \hline \vdots & \ddots & \vdots \\ \hline d_k(0)c_{kG} & \dots & d_k(M-1)c_{kG} \\ \hline \end{array} \begin{array}{l} \times c_{11} \\ + \\ \vdots \\ + \\ \times c_{1G} \end{array} \\
 &= \mathbf{h}_1 \odot \begin{array}{|c|c|c|} \hline d_1(0) & \dots & d_1(M-1) \\ \hline \end{array} \\
 & \quad \text{(a) Synchronous} \\
 & \mathbf{h}_k \odot \begin{array}{|c|c|c|} \hline d_k(\tau_k)c_{k1} & \dots & d_k(M-1+\tau_k)c_{k1} \\ \hline \vdots & \ddots & \vdots \\ \hline d_k(\tau_k)c_{kG} & \dots & d_k(M-1+\tau_k)c_{kG} \\ \hline \end{array} \begin{array}{l} \times c_{11} \\ + \\ \vdots \\ + \\ \times c_{1G} \end{array} \\
 &= \mathbf{h}_k \odot \begin{array}{|c|c|c|} \hline d_k(\tau_k) & \dots & d_k(M-1+\tau_k) \\ \hline \end{array} \left(\sum_{g=1}^G c_{kg}c_{1g} \right) \\
 & \quad \text{(b) Asynchronous}
 \end{aligned}$$

Figure 3.3: Block Spread CDMA Despreading Operation

After CP removal is completed, the received signal at the BS-CDMA receiver during the g^{th} chip-block interval, $\mathbf{r}(g)$, can be expressed as

$$\mathbf{r}(g) = \sum_{k=1}^K \mathbf{h}_k(g) \otimes \mathbf{x}_{k,g} + \mathbf{n}(g) \quad (3.5)$$

where $\mathbf{h}_k(g) = [h_k(0) \ \dots \ h_k(L-1)]^T$ is the channel between the receive antenna and k^{th}

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user's antenna and $\mathbf{n}(g)$ is the AWGN component. Based on the timing analysis presented above, it is reasonably safe to conclude that the channel is likely to remain stationary or at least largely invariant throughout the transmission of G chip-blocks, i.e. $\mathbf{h}_k(1) \approx \dots \approx \mathbf{h}_k(G) \equiv \mathbf{h}_k$. The received signal is then stacked up to construct a $G \times M$ matrix, $\mathbf{R}$. Each CP-removed chip-block now makes up a single row in $\mathbf{R}$ and $\mathbf{R}$ can be expressed as

$$\begin{aligned} \mathbf{R} &= [\mathbf{r}(1) \dots \mathbf{r}(G)]^T \\ &\approx \sum_{k=1}^K \begin{bmatrix} \mathbf{h}_k \otimes \mathbf{x}_{k,1} \\ \vdots \\ \mathbf{h}_k \otimes \mathbf{x}_{k,G} \end{bmatrix}^T + \begin{bmatrix} \mathbf{n}^T(1) \\ \vdots \\ \mathbf{n}^T(G) \end{bmatrix} \\ &= \sum_{k=1}^K \begin{bmatrix} c_{k,1} \mathbf{d}_k^T \mathbf{H}_k^T \\ \vdots \\ c_{k,G} \mathbf{d}_k^T \mathbf{H}_k^T \end{bmatrix} + \mathbf{N} \end{aligned} \quad (3.6)$$

where the circular convolution between the transmitted vectors and channel has been replaced with $\mathbf{H}_k \mathbf{d}_k$, $\mathbf{H}_k$ is a circular Toeplitz matrix with $\mathbf{h}_k$ as the first column and $\mathbf{N} = [\mathbf{n}(1) \dots \mathbf{n}(G)]^T$. Block despreading is performed by multiplying each row with the corresponding chip and then summing up all the rows together. This operation can be mathematically represented as

$$\begin{aligned} \mathbf{y}_{k_0} &= (\mathbf{c}_{k_0}^T \mathbf{R})^T \\ &= \sum_{k=1}^K \mathbf{c}_{k_0}^T \mathbf{c}_k \otimes (\mathbf{d}_k^T \mathbf{H}_k^T) + \mathbf{c}_{k_0}^T \mathbf{N} \end{aligned} \quad (3.7)$$

where k_0 is assumed to be the desired user. When perfectly orthogonal codes (such as *Walsh-Hadamard Code*) is used, the following property is true

$$\mathbf{c}_{k_0}^T \mathbf{c}_k = \begin{cases} 1 & \text{when } k = k_0 \\ 0 & \text{otherwise} \end{cases} \quad (3.8)$$

This means that the result of the block despreading operation can be simplified to

$$\mathbf{y}_{k_0} = \mathbf{H} \mathbf{d}_{k_0} + \mathbf{v}_{k_0} \quad (3.9)$$

where $\mathbf{v}_{k_0} = \mathbf{c}_{k_0}^T \mathbf{N}$. After the chips are converted back to symbols, the remaining processing

blocks are the same as that of a SC-FDE receiver. FFT operation is performed immediately after the despreading operation. This is followed by frequency-domain equalization that consists of one-tap multiplication of each tone. After equalization is completed, the equalized tones are transformed back into time-domain through IFFT operation which is followed by symbol detection.

One major advantage that comes with such despreading operation is the ability to preserve code orthogonality even during asynchronous transmission, where late / early timing mismatch between different users' signal is inevitable. In order to give a clearer view, Figure 3.3 illustrates the despreading operation during different transmission environment. Figure 3.3(a) shows the despreading operation that occurs when there is no timing misalignment, which is what has been described earlier. Figure 3.3(b) shows how the orthogonal property is preserved even when symbols transmitted by different users do not arrive synchronously. This illustration shows the robustness of a BS-CDMA scheme as it allows symbol mis-alignment without affecting the chip alignment that is critical in orthogonal design. It is this capability that makes BS-CDMA suitable for uplink transmission, where most conventional CDMA schemes have performed poorly.

The other advantage that comes with BS-CDMA is a significant reduction in receiver processing complexity. By performing the despreading operation immediately after CP-removal, the chips are converted into symbols. It can be seen clearly that the subsequent operations would occur at symbol-level instead of chip-level that happens in conventional CDMA schemes. This means that the processing complexity of those blocks is now reduced by a factor of the spreading factor, G . For downlink deployment, such reduction is vital as it will mitigate the additional complexity required to perform the IFFT operation to recover the time-domain symbols. For uplink deployment, the complexity reduction serves as an advantage to the MUI-free reception.

BS-CDMA offers receiver design that is low in complexity but yet robust against multi-user interference, in particular during the asynchronous uplink transmission. The advantages that come with BS-CDMA at the physical layer could be related to the subsequent higher layer protocol. For example, the smaller sensitivity of BS-CDMA performance against timing misalignment means that timing or frame synchronization can be performed less frequent. In addition, if a transmitting station is stationary, it causes minimal interference to other users. In addition, such station is likely to experience good error-rate performance

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compared to another station that is moving at high vehicular mobility. This means it is possible to support features such as adaptive modulation and coding, variable rate transmission etc. These features are usually done at the higher layer protocol. However, the block despreading operation has to wait for all chip-blocks of a single block of symbols to arrive before it outputs the despread block of symbols. This means that a BS-CDMA receiver introduces slightly longer delay in-between block of symbols. This delay needs to be considered when designing the higher layer protocol for BS-CDMA.

From hardware design point of view, the considerably reduced number of FFT operations translates to several advantages. The smaller number of operations required leads to less amount of power required to carry out the processing. In addition to the lower power consumption, the lower amount of processing involved might make it possible for some of the hardware units to be reused and hence reducing the overall hardware complexity. When asynchronous uplink transmission is considered, conventional CDMA schemes are expected to perform poorly unless the highly complex joint multi-user detector is used. Meanwhile, recent multiple access techniques proposed for uplink such as OFDMA / SCFDMA either requires power inefficient amplifier or vulnerable to physical impairments such as CFO or both. While the problem of power efficiency in OFDMA system can be mitigated using PAPR reduction techniques, additional hardware complexity is usually involved and certain amount of performance degradation can be expected. The performance sensitivity of OFDMA / SCFDMA to the presence of CFO requires frequent CFO tracking and correction that again incur additional complexity. However, from software design point of view, a BS-CDMA system is likely to incur slightly larger buffer size. This is because processing of a single block of information symbols will take place only after all the G chip-blocks have arrived. In other words, a BS-CDMA system could potentially offer hardware architecture that is more power efficient and costs lower compared to conventional CDMA and existing uplink multiple access schemes.

3.4 Theroretical Analysis

3.4.1 Slow Fading Channel

It was shown earlier that if the assumption about channel stationarity is true, the output of the block despreading operation will be MUI-free and the output of the FFT operation is then given as

$$\bar{\mathbf{y}}_{k_0} = \bar{\mathbf{H}}_{k_0} \bar{\mathbf{s}}_{k_0} + \bar{\mathbf{v}} \quad (3.10)$$

When ZF technique is employed, the equalized frequency-tones of the transmitted signal can be expressed as

$$\hat{\bar{\mathbf{s}}}_{k_0} = \bar{\mathbf{s}}_{k_0} + \bar{\mathbf{G}}_{ZF} \bar{\mathbf{v}} \quad (3.11)$$

where $\bar{\mathbf{G}}_{ZF} = (\bar{\mathbf{H}}_{k_0}^H \bar{\mathbf{H}}_{k_0})^{-1} \bar{\mathbf{H}}_{k_0}^H$. The recovered time-domain symbols can be represented as

$$\begin{aligned} \hat{\mathbf{s}}_{k_0} &= \mathbf{s}_{k_0} + \mathbf{F}^H \bar{\mathbf{G}}_{ZF} \bar{\mathbf{v}} \\ &= \mathbf{s}_{k_0} + \mathbf{F}^H \gamma (\gamma \mathbf{F} \mathbf{H}_{k_0}^H \mathbf{H}_{k_0} \mathbf{F}^H)^{-1} \mathbf{F} \mathbf{H}_{k_0}^H \mathbf{F}^H \mathbf{F} \mathbf{v} \\ &= \mathbf{s}_{k_0} + (\mathbf{H}_{k_0}^H \mathbf{H}_{k_0})^{-1} \mathbf{H}_{k_0}^H \mathbf{v} \\ &= \mathbf{s}_{k_0} + \mathbf{w} \end{aligned} \quad (3.12)$$

where $\mathbf{w} = (\mathbf{H}_{k_0}^H \mathbf{H}_{k_0})^{-1} \mathbf{H}_{k_0}^H \mathbf{v}$ and γ corresponds to the SNR. Now, the covariance matrix of $\mathbf{v}$ is found to be

$$\begin{aligned} E[\mathbf{v} \mathbf{v}^H] &= E \left[\left(\sum_{g=1}^G c_{k_0,g} \mathbf{n}(g) \right) \left(\sum_{g=1}^G c_{k_0,g}^* \mathbf{n}^H(g) \right) \right] \\ &= E \left[\frac{1}{G} \sum_{g=1}^G \mathbf{n}(g) \mathbf{n}^H(g) \right] \\ &= \left[\frac{1}{G} \sum_{g=1}^G \sigma_n^2 \mathbf{I}_M \right] \\ &= \sigma_n^2 \mathbf{I}_M \end{aligned} \quad (3.13)$$

which means that the noise characteristics remain unchanged at the output of the despreading operation. With this, the covariance matrix of $\mathbf{w}$ can be derived as

$$\begin{aligned} E[\mathbf{w}\mathbf{w}^H] &= \gamma(\gamma\mathbf{H}_{k_0}^H\mathbf{H}_{k_0})^{-1}\mathbf{H}_{k_0}^H\mathbf{v}\mathbf{v}^H\mathbf{H}_{k_0}(\gamma\mathbf{H}_{k_0}^H\mathbf{H}_{k_0})^{-1} \\ &= \gamma\sigma_n^2(\gamma\mathbf{H}_{k_0}^H\mathbf{H}_{k_0})^{-1} \\ &= \sigma_s^2(\gamma\mathbf{H}_{k_0}^H\mathbf{H}_{k_0})^{-1} \end{aligned} \quad (3.14)$$

This means that the noise power that is corrupting the m^{th} symbol is $\sigma_s^2\lambda_{k_0,m}$, where $\lambda_{k_0,m} = [(\gamma\mathbf{H}_{k_0}^H\mathbf{H}_{k_0})^{-1}]_{m,m}$. Therefore, the SNR for the m^{th} symbol is given as

$$SNR_m = \frac{\sigma_s^2}{\sigma_s^2\lambda_{k_0,m}} = \frac{1}{\lambda_{k_0,m}} \quad (3.15)$$

Meanwhile, when MMSE equalizer is employed, the recovered time-domain symbol vector obtained can be expressed as

$$\hat{\mathbf{s}}_{k_0} = \mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\bar{\mathbf{H}}_{k_0}\mathbf{F}\mathbf{s}_{k_0} + \mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\bar{\mathbf{v}} \quad (3.16)$$

The covariance matrix for the recovered symbols can then be derived as

$$\begin{aligned} E[\hat{\mathbf{s}}_{k_0}\hat{\mathbf{s}}_{k_0}^H] &= E[(\mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\bar{\mathbf{H}}_{k_0}\mathbf{F}\mathbf{s}_{k_0} + \mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\bar{\mathbf{v}}) \\ &\quad (\mathbf{s}_{k_0}^H\mathbf{F}^H\bar{\mathbf{H}}_{k_0}^H\bar{\mathbf{G}}_{k_0,MMSE}^H\mathbf{F} + \bar{\mathbf{v}}^H\mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}^H\mathbf{F})] \\ &= \sigma_s^2\mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\mathbf{F}\mathbf{H}_{k_0}\mathbf{F}^H\mathbf{F}\mathbf{H}_{k_0}\mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}^H\mathbf{F} + \sigma_n^2\mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\bar{\mathbf{G}}_{k_0,MMSE}^H\mathbf{F} \\ &= \sigma_s^2\gamma\mathbf{H}_{k_0}^H(\gamma\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \mathbf{I}_M)^{-1}\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H(\gamma\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \mathbf{I}_M)^{-1}\mathbf{H}_{k_0} \\ &\quad + \sigma_n^2\gamma\mathbf{H}_{k_0}^H(\gamma\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \mathbf{I}_M)^{-1}(\gamma\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \mathbf{I}_M)^{-1}\gamma\mathbf{H}_{k_0} \\ &= \gamma^2\mathbf{H}_{k_0}(\gamma\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \mathbf{I}_M)^{-1}(\sigma_s^2\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \sigma_n^2\mathbf{I}_M)(\gamma\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \mathbf{I}_M)^{-1}\mathbf{H}_{k_0} \\ &= \sigma_s^2\mathbf{G}_{k_0,MMSE}\mathbf{H}_{k_0} \end{aligned} \quad (3.17)$$

On the other hand, the equalized symbol vector can be expressed as

$$\begin{aligned} \hat{\mathbf{s}} &= \mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\bar{\mathbf{H}}_{k_0}\bar{\mathbf{s}}_{k_0} + \mathbf{F}^H\bar{\mathbf{G}}_{k_0,MMSE}\bar{\mathbf{v}} \\ &= \mathbf{F}^H\{\gamma\mathbf{F}\mathbf{H}_{k_0}^H(\gamma\mathbf{H}_{k_0}\mathbf{H}_{k_0}^H + \mathbf{I}_M)^{-1}\mathbf{F}^H\}\mathbf{H}_{k_0}\mathbf{s}_{k_0} \\ &= \mathbf{G}_{k_0,MMSE}\mathbf{H}_{k_0}\mathbf{s}_{k_0} + \mathbf{G}_{k_0,MMSE}\bar{\mathbf{v}} \end{aligned} \quad (3.18)$$

Now, let $a_m = [\mathbf{G}_{k_0, MMSE} \mathbf{H}_{k_0}]_{m,m}$. From here, the m^{th} equalized symbol and its variance can be simplified as

$$\hat{s}_{k_0,m} = a_m s_{k_0,m} + w_m \quad (3.19)$$

$$E[|\hat{s}_{k_0,m}|^2] = a_m^2 \sigma_s^2 + E[|w_m|^2] \quad (3.20)$$

By substituting Equation (3.18) into (3.20), the power of the noise corrupting the m^{th} symbol can be derived as

$$\begin{aligned} a_m \sigma_s^2 &= a_m^2 \sigma_s^2 + E[|w_m|^2] \\ E[|w_m|^2] &= \sigma_s^2 a_m (1 - a_m) \end{aligned} \quad (3.21)$$

Eventually, the SNR for the received m^{th} symbol is given as

$$SNR_m = \frac{\sigma_s^2 a_m^2}{\sigma_s^2 a_m (1 - a_m)} = \frac{a_m}{1 - a_m} \quad (3.22)$$

Once the SNR value is obtained, the theoretical BER and capacity expression can be derived by averaging over the M recovered symbols.

3.4.2 Fast Fading Channel

A fast-fading channel is used to describe the situation where the propagation channel is rapidly varying during a certain interval. In the context of BS-CDMA, the channel is considered to be fast fading if the channel coherence time is much smaller than the duration in which G chip-blocks are transmitted. Such condition occurs only when the vehicular mobility of the transmitting device is high. A study on the relationship between vehicular mobility and BER performance of BS-CDMA system has been conducted and presented in Chapter 7

Under fast-fading condition, $\mathbf{H}_k(1) \neq \dots \neq \mathbf{H}_k(G)$. The channel during the g^{th} chip-block interval can be expressed as

$$\mathbf{H}_k(g) = \tilde{\mathbf{H}}_k + \dot{\mathbf{H}}_k(g) \quad (3.23)$$

where $\tilde{\mathbf{H}}_k$ is the channel mean averaged over G chip-blocks and $\dot{\mathbf{H}}_k(g)$ is the offset between the channel mean and the channel during the g^{th} chip-block. This means that the result of

the block despreading operation is now given as

$$\begin{aligned}
 \mathbf{y}_{k_0} &= \sum_{g=1}^G \sum_{k=1}^K (\tilde{\mathbf{H}}_k + \dot{\mathbf{H}}_k(g)) \mathbf{s}_k c_{k,g} c_{k_0,g} + \mathbf{v} \\
 &= \underbrace{\tilde{\mathbf{H}}_{k_0} \mathbf{s}_{k_0} \left(\sum_{g=1}^G c_{k_0,g}^2 \right)}_1 + \sum_{k \neq k_0} \underbrace{\tilde{\mathbf{H}}_k \mathbf{s}_k \left(\sum_{g=1}^G c_{k_0,g} c_{k,g} \right)}_0 + \sum_{k=1}^K \underbrace{\mathbf{s}_k \left(\sum_{g=1}^G \dot{\mathbf{H}}_k(g) c_{k_0,g} c_{k,g} \right)}_{\neq 0} + \mathbf{v} \\
 &= \tilde{\mathbf{H}}_{k_0} \mathbf{s}_{k_0} + \underbrace{\sum_{k=1}^K \mathbf{s}_k \left(\sum_{g=1}^G \dot{\mathbf{H}}_k(g) c_{k_0,g} c_{k,g} \right)}_{\text{Residual MUI}} + \mathbf{v}
 \end{aligned} \tag{3.24}$$

When ZF equalizer is used, the ZF multiplier matrix $\mathbf{G}_{k_0,ZF}$ is

$$\mathbf{G}_{k_0,ZF} = \gamma (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \tilde{\mathbf{H}}_{k_0} \tag{3.25}$$

After equalization is performed, it can be shown that the SNR value on the m^{th} symbol can be expressed as

$$\begin{aligned}
 SNR_m &= \frac{\sigma_s^2}{\sigma_s^2 \left\{ \lambda_{k_0,m} + \frac{1}{G^2} \sum_{k=1}^K \sum_g |\beta_{kk_0g,m}|^2 \right\}} \\
 &= \frac{1}{\lambda_{k_0,m} + \frac{1}{G^2} \sum_{k=1}^K \sum_g |\beta_{kk_0g,m}|^2}
 \end{aligned} \tag{3.26}$$

where $\lambda_{k_0,m} = \left[\gamma (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \right]_{m,m}$ and $\beta_{kk_0g,m} = \left[\mathbf{G}_{k_0,ZF} \tilde{\mathbf{H}}_k(g) \right]_{m,m}$. Similarly, the SNR_m for MMSE can be shown to be

$$SNR_m = \frac{\sigma_s^2 a_{k_0,m}}{\sigma_s^2 a_{k_0,m} (1 - a_{k_0,m}) + \frac{1}{G^2} b_{k_0,m}} \tag{3.27}$$

where $a_{k_0,m}$ and $b_{k_0,m}$ are defined as follow:

$$\begin{aligned}
 a_{k_0,m} &= \left[\mathbf{G}_{k_0,MMSE} \tilde{\mathbf{H}}_{k_0} \right]_{m,m} \\
 b_{k_0,m} &= \left[\mathbf{G}_{k_0,MMSE} \left\{ \sum_{k \neq k_0} \sum_g^{PG} \dot{\mathbf{H}}_k(g) \dot{\mathbf{H}}_k^H(g) \right\} \mathbf{G}_{k_0,MMSE}^H \right]_{m,m}
 \end{aligned} \tag{3.28}$$

Derivation for Equation (3.26) and (3.27) can be found in Appendix C.

Now that the SNR value on each symbol has been found, the theoretical probability of error rate can be found by substituting SNR_m into the equation below

$$\begin{aligned} P_e &= \frac{1}{M} \sum_{m=1}^M \frac{1}{\sqrt{2\pi}} \int_{\sqrt{SNR_m}}^{\infty} e^{-\frac{t^2}{2}} dt \\ &= \frac{1}{M} \sum_{m=1}^M Q(\sqrt{SNR_m}) \end{aligned} \quad (3.29)$$

3.5 Numerical Results

The BS-CDMA architecture is modeled and its performance was investigated using monte-carlo simulation for both uplink and downlink deployment. The performance obtained is then compared against the benchmark theoretical curve and against conventional CDMA systems, namely MC-CDMA and CP-CDMA that were being deployed using similar configurations.

The whole investigation exercise targets high-rate high-mobility communication systems that is capable of addressing the needs of future generation of mobile communication systems. The transmission carrier frequency is 5GHz, with a chip-rate of 40Mcps. Two different block sizes are being considered, namely $M = 256$ and $M = 512$. The length of CP employed is $L_{CP} = \frac{1}{4}M$. This effectively means that a single chip-block is of size 320 and 640. For $M = 256$, the duration of each chip-block, T , is $8\mu\text{secs}$, where $6.4\mu\text{secs}$ is the duration of information chips and $1.6\mu\text{secs}$ is the duration of guard interval. At the receiver, the system bandwidth of 40MHz is divided into M subchannels and equalized individually. The sizes of processing gain used are 8, 16 and 32. The number of active users is varied accordingly to simulate different loading conditions in the system. The channel model assumes *Jakes' Rayleigh* distribution [132] and the number of paths is assumed to be 24. It is assumed that the maximum delay spread is less than the guard duration and hence the inter-symbol interference (ISI) due to the presence of multipath is avoided. During the simulation exercise, different vehicular mobility conditions (0Hz (stationary), 463Hz (medium-high) and 926Hz (high)) are simulated to model the practical deployment of mobile communication systems. The modulation scheme used is BPSK and QPSK. The complete set of parameter configuration that was used is tabulated in Table 3.1.

Figure 3.4 shows the theoretical BER that is plotted based on the expression given in

Table 3.1: CDMA System Models: Simulation Parameters

	MC – and CP – CDMA	BS – CDMA
Carrier Frequency, F	5 GHz	5GHz
Chip rate, F_C	40.96 Mcps	40.96 Mcps
Modulation	BPSK — QPSK	BPSK — QPSK
OFDM / Chip-Block Size, M	256 (512)	256 (512)
FFT / IFFT Size, M	256 (512)	256 (512)
OFDM Block / Chip-Block Duration, T	8 μ secs (16 μ secs)	8 μ secs (16 μ secs)
Data Duration per OFDM Block, T_D	6.4 μ secs (12.8 μ secs)	6.4 μ secs (12.8 μ secs)
Guard Duration per OFDM Block, T_{CP}	1.6 μ secs (3.2 μ secs)	1.6 μ secs (3.2 μ secs)
Super-Block Size, M_{SB}	1 OFDM Block	PG OFDM Blocks
Super-Block Duration, T_{SB}	8 μ secs (16 μ secs)	PG $\times$ 8 μ secs (PG $\times$ 16 μ secs)
Processing Gain, PG	8 16 32	8 16 32
Symbols per Super-Block, N_{SB}	32 16 8 (64 32 16)	256 (512)
Load Condition	1 user Half-Load Full-Load	1 user Half-Load Full-Load
Doppler Spread, F_D	0 Hz 463 Hz 926 Hz	0 Hz 463 Hz 926 Hz
Vehicular Mobility, V	0 kph 100 kph 200 kph	0 kph 100 kph 200 kph
Channel Coherence Time, T_{CO}	∞ 387 μ secs 193 μ secs	∞ 387 μ secs 193 μ secs

(a) Performance with ZF Equalizer

(b) Performance with MMSE Equalizer

Figure 3.4: Theoretical BER Performance of BS-CDMA System

Equation (3.29) and substituting the SNR values found from Equation (3.26) and (3.27). The block size is assumed to be 256 and the processing gain is chosen to be 32. Figure 3.4(a) and Figure 3.4(b) depict the BER curve for the ZF and MMSE equalizer respectively. As expected, the MMSE equalizer delivers superior performance compared to the ZF equalizer. This is due to the well-known fact that ZF technique tends to amplify the noise at weak sub-carriers.

When ZF technique is employed and the channel is stationary ($F_D = 0\text{Hz}$), a degradation of about 2dB is observed when the number of users increased from 1 to 16 at the BER level of 1%. However, no further significant degradation is observed when the number of users is increased from 16 to 32. When the Doppler spread is 463Hz and ZF technique is used, about 1dB degradation is observed when the number of users is increased from 1 to 16 and another 1dB is observed when the number of users is increased to 32. During extremely high mobility condition (926Hz), about 5dB degradation occurs when the loading condition is increased from 1 user to half-load and an error-floor of above 10% is observed.

When MMSE technique is used, the system performance is totally unaffected by the loading condition. When the Doppler spread is 463Hz , performance degradation of less than 1dB is observed when the number of users increased to 16 and a further 1dB degradation occurs when it is increased to 32 for BER of 1%. When the Doppler spread is 926Hz , an error floor close to 1% and 6% is observed under half-load and full-load conditions respectively.

In Figure 3.5, the performance of BS-CDMA is compared against the two conventional CDMA schemes, namely MC-CDMA and CP-CDMA. The mobility condition is chosen to be stationary. From Figure 3.5(a), it can be seen that when there is only 1 user, a MC-CDMA system employing ZF technique performs approximately 1dB better than that of both BS-CDMA and CP-CDMA systems. In this case, both BS-CDMA and CP-CDMA systems show identical performance. However, both MC-CDMA and CP-CDMA completely fail in the presence of multiple users in the uplink channel. Meanwhile, the performance of BS-CDMA shows robustness against multi-user degradation and does not show noticeable degradation even under heavy loading conditions. Similar trend is observed even when MMSE technique is used. The performance of MC-CDMA (and CP-CDMA) is about 4dB (and 1.5dB) superior to that of BS-CDMA when there is only one user transmitting in the system. However, both MC-CDMA and CP-CDMA perform poorly in the presence of multiple users in the system. This confirms BS-CDMA as the most suitable multiple-access

(a) Performance of ZF Equalizer

(b) Performance of MMSE Equalizer

Figure 3.5: BER Performance Comparison of Different CDMA Schemes
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scheme for uplink when no complicated multi-user detection is performed.

Figure 3.6 shows the BER performance of BS-CDMA system when ZF equalizer is used. In Figure 3.6(a), the mobility condition is set as 100kmph while for Figure 3.6(b), the mobility is set to 200kmph. In the figures, the theoretical curves are plotted along with the simulated ones. Based on the observation from Figure 3.6(a) and 3.6(b), the performance curves obtained through simulation study closely match those of the theoretical ones. Based on Figure 3.6(a), it can be seen that when $F_D = 463\text{Hz}$, no error-floor is observed. However, Figure 3.6(b) shows that when $F_D = 926\text{Hz}$, the *residual MUI* during full-load condition causes an error-floor that lies above the 10% level.

Figure 3.7 shows the BER performance of BS-CDMA system when MMSE equalizer is used. Similarly, the performance curves plotted in Figure 3.7(a) and 3.7(b) show close match between the curves obtained by theoretical and simulated means. From Figure 3.7(a), it can be seen that when the mobility condition is 100kmph, noticeable performance degradation is observed but no error-floor is seen. When the mobility condition is set to 200kmph, Figure 3.7(a) shows that the error-floor for half-load condition is still lower than 1%. During full-load condition, the error-floor goes up to about 10% level.

3.6 Summary

In this chapter, the basic architecture of a BS-CDMA system is presented. When single-user detector is used at an uplink receiver, it can be shown analytically that BS-CDMA scheme offers near-optimal performance compared to conventional CDMA schemes such as MC-CDMA and CP-CDMA. This is due to its ability in preserving code orthogonality even during asynchronous reception when chip misalignment occurs. A theoretical study to derive the probability of error of a BS-CDMA system is presented. BS-CDMA system for a multi-user environment is extensively simulated for different channel parameters. From the simulation studies, the results obtained confirm the suitability of BS-CDMA over conventional CDMA schemes for uplink transmission. However, though BS-CDMA is virtually free from MUI during low mobility environment, its performance is prone to the channel variation that occurs during high mobility conditions. This issue will be addressed in the later part of this thesis.

(a) (PG=32, V=100kph)

(b) (PG=32, V=200kph)

Figure 3.6: BER Performance of Uplink BS-CDMA System with ZF Equalizer
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(a) (PG=32, V=100kph)

(b) (PG=32, V=200kph)

Figure 3.7: BER Performance of Uplink BS-CDMA System with MMSE Equalizer
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The contents in this chapter can be summarized by the following points:

- basic architecture of BS-CDMA system is presented;
- theoretical performance of BS-CDMA system is presented and supported with results from simulation studies;
- uplink performance comparison with conventional CDMA schemes, namely MC-CDMA and CP-CDMA is presented;
- results from simulation studies show that BS-CDMA system is capable of preserving code orthogonality during asynchronous transmission and is capable of outperforming conventional schemes; and
- virtually free MUI performance is achieved in BS-CDMA during low mobility transmission while performance degradation starts to occur during high mobility conditions.

Chapter 4

Layered Space-Time Block Spread CDMA Systems (LST BS-CDMA)

4.1 Introduction

Future mobile communication envisage enhancement of data rate by about ten times or more, while the available frequency spectrum is expected to grow at a much smaller scale. Increasing data rate within available bandwidth to match the demands of the future mobile communication systems is one of the primary requirement for a successful transmission technology. The proposed BS-CDMA architecture is studied in this context and a MIMO BS-CDMA transmission structure with layered space-time (LST) architecture is proposed to enhance the performance of the BS-CDMA system. The present chapter summarizes the results of this study.

Chapter 4 is organized as follows. Section 4.2 highlights the motivation behind the LST BS-CDMA proposal. Section 4.3 presents the system model for LST BS-CDMA architecture. Section 4.4 describes the algorithms that can be used to estimate the channel state information (CSI) for LST BS-CDMA systems with the help of specialized sequences with pre-defined frequency tone nulling. Section 4.5 discusses the results from the simulation studies.

4.2 Motivation

It is important to recall the fundamental principle in which a BS-CDMA system would be operating at the optimum condition. BS-CDMA architecture was initially proposed based on the observation that if the duration of a single chip is much lower than the coherence time, the channel would remain largely invariant over many chip periods or stationary. Leveraging on this property and the structure in which chips are transmitted in BS-CDMA, the block processing approach results in MUI-free reception and considerable reduction in processing complexity at the receiver. As the data rate increases, the chip duration decreases proportionally. This is highly favourable towards BS-CDMA as it means that a higher transmission rate translates to a lower degree of channel variation during the transmission of a single block. On the other hand, the channel coherence time is inversely related to the mobility condition of the transmitting and receiving units. This means that a BS-CDMA system with high data rate is more robust with respect to the mobility environment.

Meanwhile, the effective symbol rate of a CDMA user decreases as the processing gain increases. On the other hand, a large processing gain is required to support high number of users. This contradiction imposes a serious challenge for system design that requires to support large number of users and high data rate transmission. As there is a limit to the amount of bandwidth expansion that is allowable, it is thus important to make up for the rate loss by enhancing the spectral efficiency.

The first possible way to compensate for the loss in symbol rate is by using higher modulation scheme. For example, a 256-QAM modulation offers a symbol rate that is four times higher than a QPSK modulation. Such higher modulation schemes have been used in IEEE802.11[38] wireless LAN standards. However, such schemes incur additional constraints such as higher peak-to-average power ratio (PAPR) values and higher demodulation complexity. In BS-CDMA, the use of single-carrier transmission generates considerably lower PAPR as compared to a multi-carrier based system. Therefore, for a particular power amplifier with certain linear range, a BS-CDMA system can accommodate higher modulation schemes than multi-carrier systems. In addition, the use of simpler single-user (instead of maximum-likelihood multi-user) detector involves much lower processing complexity and is practically attractive.

However, a communication system that employs single antenna at both the receiver and

transmitter has its system capacity limited by the *Shannon Capacity* and it grows only logarithmically as a function of SNR. As discussed in chapter 2, the possibility of using multiple antennae at both the BS and MS triggered explosive growth in research activity pertaining to MIMO systems. MIMO transmission is known as a promising technique to increase the transmission rate without bandwidth expansion. The issue of system capacity enhancement in MIMO cellular system is addressed in [134]. One MIMO architecture that achieves high spectral efficiency without the need of additional power or bandwidth is Vertical Bell Labs layered Space Time (VBLAST) coding [135].

The idea behind VBLAST coding [79] is mainly to demultiplex a single data stream into several substreams, each of which is fed into a single transmitter. This transmission scheme reverses the destructive effect of multipath to create channels with spatial diversity and thus improving the system performance. Layered space-time architecture for single-carrier BS-CDMA (LST BS-CDMA) was proposed in [136] which attempts to combine the advantage of a MIMO system with BS-CDMA. Details about the LST BS-CDMA system model and its performance is presented in the following sections.

4.3 System Model

Figure 4.1: A Generalized Baseband LST BS-CDMA Transmitter Architecture

A simplified general baseband model of a LST BS-CDMA transmitter is depicted in Figure 4.1. The single continuous stream of information symbols is fed into a VBLAST mapper. The task of the mapper is essentially to divide the single stream of symbols into multiple sub-streams, where the number of sub-streams output by the VBLAST mapper is denoted by N_T , or effectively equivalent to the number of layers or antenna employed at the trans-

mitter. At any one time, the VBLAST mapper takes in N_TM symbols, $\mathbf{d}_k(i)$:

$$\mathbf{d}_k(i) = \begin{bmatrix} d_k((i-1)N_TM + 1) & \cdots & d_k(iN_TM) \end{bmatrix}^T \quad (4.1)$$

where M represents the block transmission size. Let $d_k(n)$ denote the n^{th} symbol belonging to the k^{th} user and n is related to i by $n = (i-1)N_TM + m$ with $0 < m \leq N_TM$. For brevity, the index i will be removed from subsequent discussions. The mapper arranges the input symbol vector into matrix $\mathbf{D}_k$, of size $M \times N_T$ which can be expressed as:

$$\begin{aligned} \mathbf{D}_k &= \begin{bmatrix} d_k(1) & \cdots & d_k(M) \\ \vdots & \ddots & \vdots \\ d_k((N_T-1)M + 1) & \cdots & d_k(N_TM) \end{bmatrix} \\ &= \begin{bmatrix} \mathbf{d}_{k,1} & \cdots & \mathbf{d}_{k,N_T} \end{bmatrix}^T \end{aligned} \quad (4.2)$$

where $\mathbf{d}_{k,p} = \begin{bmatrix} d_k((p-1)M + 1) & \cdots & d_k(pM) \end{bmatrix}^T$ is the stream of symbols on the p^{th} stream and will be eventually transmitted on the p^{th} antenna. Based on the output of the VBLAST mapper, block spreading operation is carried out separately on each sub-stream. (Pictorial representation of the process of the block spreading operation can be found in Figure 3.2). The block spreading operation for the p^{th} stream for the k^{th} user is given by the following mathematical operation:

$$\begin{aligned} \mathbf{X}_{k,p} &= \mathbf{c}_k \otimes \mathbf{d}_{k,p}^T \\ &= \begin{bmatrix} c_{k,1}d_k((p-1)M + 1) & \cdots & c_{k,1}d_k(pM) \\ \vdots & \ddots & \vdots \\ c_{k,G}d_k((p-1)M + 1) & \cdots & c_{k,G}d_k(pM) \end{bmatrix} \end{aligned} \quad (4.3)$$

where $\otimes$ corresponds to the *Kronecker* product operation, $\mathbf{c}_k = \begin{bmatrix} c_{k,1} & \cdots & c_{k,G} \end{bmatrix}^T$ is the spreading code belonging to the k^{th} user, G is the size of the spreading gain and $\mathbf{X}_{k,p}$ is a $G \times M$ chip matrix. The chips are then output row-by-row and each row is defined as one chip-block, $\mathbf{x}_{k,p}(g)$:

$$\mathbf{x}_{k,p}(g) = \begin{bmatrix} c_{k,g}d_k((p-1)M + 1) & \cdots & c_{k,g}d_k(pM) \end{bmatrix}^T \quad (4.4)$$

To facilitate frequency-domain processing at the receiver, CP is inserted and the g^{th} CP-

Figure 4.2: A Generalized Baseband LST BS-CDMA Receiver Architecture

appended chip-block is given as:

$$\mathbf{x}_{k,p}^{(CP)}(g) = \begin{bmatrix} \mathbf{I}_M^{(CP)T} & \mathbf{I}_M \end{bmatrix}^T \mathbf{x}_{k,p}(g) \quad (4.5)$$

where $\mathbf{I}_M^{(CP)}$ consists of the last L_{CP} rows of $\mathbf{I}_M$.

The generalized BS-CDMA receiver is depicted in Figure 4.2. After CP removal, the received signal at the q^{th} antenna during the g^{th} chip-block period is given by:

$$\begin{aligned} \mathbf{r}_q(g) &= \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{h}_{k,qp}(g) \otimes \mathbf{x}_{k,p}(g) + \mathbf{n}_q(g) \\ &= \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{H}_{k,qp}(g) \mathbf{x}_{k,p}(g) + \mathbf{n}_q(g) \end{aligned} \quad (4.6)$$

where $\mathbf{h}_{k,qp}(g)$ is the channel pair between the p^{th} antenna belonging to the k^{th} user and the q^{th} receive antenna during the g^{th} chip-block period. $\mathbf{H}_{k,qp}(g)$ is a circular *Toeplitz* matrix (see Appendix A) with $\mathbf{h}_{k,qp}(g)$ as the first column, $\mathbf{n}_q(g)$ corresponds to the AWGN component at the q^{th} antenna during the g^{th} chip-block interval and K refers to the total number of users who are actively transmitting.

The block despreading module considers the G chip-blocks that are convolved with the channel and performs despreading operation prior to other processing. Despreading operation is carried out by multiplying each chip-block with the corresponding chip and then

summing all the G chip-blocks together. Mathematically, this can be expressed as:

$$\begin{aligned} \mathbf{y}_{k_0,q} &= \sum_{g=1}^G c_{k_0,g} \mathbf{r}_q(g) \\ &= \sum_{g=1}^G c_{k_0,g}^2 \sum_{p=1}^{N_T} \mathbf{H}_{k_0,qp}(g) \mathbf{d}_{k_0,p} + \sum_{k \neq k_0} \sum_{g=1}^G c_{k_0,g} c_{k,g} \sum_{p=1}^{N_T} \mathbf{H}_{k,qp}(g) \mathbf{d}_{k,p} + \sum_{g=1}^G \mathbf{n}_q(g) c_{k_0,g} \end{aligned} \quad (4.7)$$

In the case of downlink reception, the propagation channels for all users are identical, i.e. $\mathbf{H}_{k,qp}(g) \equiv \mathbf{H}_{qp}(g) \forall k$. In addition, if the channel is assumed to be stationary, then $\mathbf{H}_{qp} \approx \mathbf{H}_{qp}(g) \forall g$. Hence, Equation (4.7) translates to:

$$\begin{aligned} \mathbf{y}_{k_0,q} &= \underbrace{\sum_{g=1}^G c_{k_0,g}^2}_{1} \sum_{p=1}^{N_T} \mathbf{H}_{qp} \mathbf{d}_{k_0,p} + \underbrace{\sum_{k \neq k_0} \sum_{g=1}^G c_{k_0,g} c_{k,g}}_0 \sum_{p=1}^{N_T} \mathbf{H}_{qp} \mathbf{d}_{k,p} + \mathbf{n}_q(g) c_{k_0,g} \\ &= \sum_{p=1}^{N_T} \mathbf{H}_{qp} \mathbf{d}_{k_0,p} + \mathbf{v}_{k_0,q} \end{aligned} \quad (4.8)$$

where $\mathbf{v}_{k,q} = \sum_{g=1}^G c_{k,g} \mathbf{n}_q(g)$. As can be seen from Equation (4.8), the interferences from other users are cancelled off due to the block despreading operation. The output of the block despreading operation consists of the desired symbols that are corrupted with irreducible noise. It is then fed into FFT module, which outputs $\bar{\mathbf{y}}_{k_0,q}$ as follows:

$$\bar{\mathbf{y}}_{k_0,q} = \sum_{p=1}^{N_T} \bar{\mathbf{H}}_{qp} \bar{\mathbf{d}}_{k_0,p} + \bar{\mathbf{v}}_{k_0,q} \quad (4.9)$$

where $\bar{\mathbf{v}}_{k_0,q} = \mathbf{F} \mathbf{v}_{k_0,q}$. There are two different approaches to carry out the equalization operation. The first approach directly processes all the tones together. However, this would involve high computational cost as both ZF and MMSE equalizers now require the inverse computation of $N_T M \times N_T M$ matrix. The second alternative is to apply per-tone equalization, which is of much lower complexity. This is accomplished by first arranging

the l^{th} tone from all N_R antennae stack-by-stack as:

$$\begin{aligned}\bar{\mathbf{y}}_{k_0}(l) &= \left[\left[\bar{\mathbf{y}}_{k_0,1} \right]_l \quad \cdots \quad \left[\bar{\mathbf{y}}_{k_0,N_R} \right]_l \right]^T \\ &= \begin{bmatrix} \bar{h}_{11}(l) & \cdots & \bar{h}_{1N_T}(l) \\ \vdots & \ddots & \vdots \\ \bar{h}_{N_R1}(l) & \cdots & \bar{h}_{N_RN_T}(l) \end{bmatrix} \begin{bmatrix} \bar{d}_{k_0,1}(l) \\ \vdots \\ \bar{d}_{k_0,N_T}(l) \end{bmatrix} + \begin{bmatrix} \bar{v}_{k_0,1}(l) \\ \vdots \\ \bar{v}_{k_0,N_R}(l) \end{bmatrix} \\ &= \bar{\mathbf{H}}(l)\bar{\mathbf{d}}_{k_0}(l) + \bar{\mathbf{v}}(l)\end{aligned}\quad (4.10)$$

where $\bar{\mathbf{H}}(l)$, $\bar{\mathbf{d}}_{k_0}(l)$ and $\bar{\mathbf{v}}(l)$ are of size $N_R \times N_T$, $N_T \times 1$ and $N_R \times 1$ respectively, with $\left[\bar{\mathbf{H}}(l) \right]_{q,p} = \bar{h}_{qp}(l)$, $\left[\bar{\mathbf{d}}_{k_0}(l) \right]_p = \bar{d}_{k_0,p}(l)$ and $\left[\bar{\mathbf{v}}(l) \right]_q = \bar{v}_{k_0,q}(l)$ respectively. The ZF and MMSE equalizer matrices for the l^{th} tone are given as:

$$\bar{\mathbf{G}}_{ZF}(l) = \gamma(\gamma\bar{\mathbf{H}}^H(l)\bar{\mathbf{H}}(l))^{-1}\bar{\mathbf{H}}^H(l) \quad (4.11)$$

$$\bar{\mathbf{G}}_{MMSE}(l) = \gamma\bar{\mathbf{H}}^H(l)(\gamma\bar{\mathbf{H}}(l)\bar{\mathbf{H}}^H(l) + \mathbf{I}_{N_T})^{-1} \quad (4.12)$$

After equalization is completed, the M tones belonging to the p^{th} transmit antenna are fed into a single IFFT module to transform the tones back into time-domain for detection.

In the case of uplink reception, conventional CDMA schemes with single-user detector would perform poorly due to the destruction in code orthogonality. It can be shown that given the channel stationarity assumption remains true, the single-user detector for uplink LST BS-CDMA performs equally well with its downlink counterpart. However, the performance of LST BS-CDMA system is susceptible to the presence of high Doppler spread. More detailed description about this and the attempt to overcome its detrimental effect is presented in Chapter 7.

4.4 Estimation of Channel State Information (CSI) for LST BS-CDMA Systems

4.4.1 System Model

In the channel estimation problem, pilot or training symbols known to both the transmitter and receiver, are usually transmitted periodically to facilitate the task of estimating the

channel state information (CSI). In essence, the channel estimation problem is one that attempts to obtain a solution to minimize a given cost function.

Recall that after CP removal, the received signal on the q^{th} antenna is given in Equation (4.6). The channel estimation method studied here is an overlay approach, where a dedicated spread code is assigned for pilot and the pilot chips are transmitted simultaneously with users' data. When the pilot symbols are transmitted together with the data, the received downlink signal on the q^{th} antenna becomes:

$$\mathbf{r}_q^{(g)} = \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{h}_{qp}(g) \otimes \mathbf{x}_{k,p}(g) + \sum_{p=1}^{N_T} \mathbf{h}_{qp}(g) \otimes \tilde{\mathbf{t}}_p(g) + \mathbf{n}_q(g) \quad (4.13)$$

where $\mathbf{h}_{1,qp} = \dots = \mathbf{h}_{K,qp} = \mathbf{h}_{qp}$, $\mathbf{t}_p(g) = c_{t,g} \mathbf{t}_p$. $\mathbf{t}_p$ is the $M \times 1$ pilot chips to be transmitted from antenna p and $\mathbf{c}_{t,g} = [c_{t,1} \dots c_{t,G}]^T$ is the spreading code assigned for the pilot sequence.

The result of performing despreading on the received downlink signal using pilot-assigned spread code, followed by FFT operation is given as:

$$\bar{\mathbf{z}}_q = \sum_{p=1}^{N_T} \mathbf{F}_M \mathbf{T}_p \mathbf{F}_M^H \mathbf{F}_M \mathbf{h}_{qp} + \bar{\mathbf{v}}_q \quad (4.14)$$

where $\mathbf{T}_p$ is a circular Toeplitz matrix with the first column as $\tilde{\mathbf{t}}_p$. Using the special property of a circular Toeplitz matrix when pre- and post-multiplied with FFT/IFFT matrix, Equation (4.14) can be simplified into:

$$\bar{\mathbf{z}}_q = \sum_{p=1}^{N_T} \bar{\mathbf{T}}_p \mathbf{F}_{M|L} \mathbf{h}_{qp|L} + \bar{\mathbf{n}}_q \quad (4.15)$$

where $\bar{\mathbf{T}}_p$ is a diagonal matrix with its diagonal elements as tones of the pilot symbols transmitted from the p^{th} antenna, $\mathbf{F}_{M|L}$ is now a $M \times L$ matrix containing only the first L columns of $\mathbf{F}_M$ and $\mathbf{h}_{qp|L}$ is the channel parameters with all the trailing $(M - L)$ zeros omitted. L is the length of guard duration or CP. In the following, two approaches that minimize the MSE across all subcarriers are described.

4.4.2 Special Sequence with Pre-Defined Frequency Tone Nulling

Designing pilot sequences that have null frequency components at pre-defined indices exhibit desirable features for a SC-FDE based system. A special class of repeated finite-length sequence [137] is considered in this context. Interestingly, phase-rotation of this repeated sequence allows us to define the indices of non-null tones. The detailed procedures are laid out below.

4.4.2.1 Repeated Finite-Length Sequence

Since the pilot sequence discussed in this paper is to be used in LST BS-CDMA, symbol representation will follow that of Section 4.3. M corresponds to the block size, N_T represents the number of transmit antenna and M is assumed to be divisible by N_T .

1. Given an arbitrary sequence of length- $(\frac{M}{N_T})$
 $\mathbf{t} = [t(0) \cdots t(\frac{M}{N_T} - 1)]^T$
2. Repeat or append the sequence with itself N_T times to generate sequence of length M .
 $\hat{\mathbf{t}} = [\mathbf{t} \cdots \mathbf{t}]^T$

It was shown in [137] that the m^{th} frequency tone, $\bar{t}(m)$ of $\hat{\mathbf{t}}$ is given as

$$\bar{t}(m) = \begin{cases} N_T \sum_{l=0}^{\frac{M}{N_T}-1} t(l) e^{-j2\pi ml} & \text{for } m = 0, N_T, \dots \\ 0 & \text{Otherwise} \end{cases} \quad (4.16)$$

4.4.2.2 Sequence with Pre-Defined Null Tones

Earlier, non-null tones of a repeated finite-length sequence was shown to exist only at index kN_T (for $k = 0$ to $\frac{M}{N_T} - 1$). However, in order to facilitate channel estimation in SC-FDE based system, it is necessary to generate such sequence with non-null components at different index set. This can be achieved by rotating the phase of the sequence generated using the above procedure. Now, let $\Omega_p = \{kN_T + p - 1\}$ for $k = 0$ to $\frac{M}{N_T} - 1$. A repeated

finite-length sequence $\hat{\mathbf{t}}_p$ will have its non-null tones at index $m \in \Omega_p$ after it is phase-rotated by $\theta_p = \frac{2\pi(p-1)}{M}$ and $\tilde{\mathbf{t}}_p$, the phase-rotated copy of $\hat{\mathbf{t}}_p$ is given by:

$$\tilde{\mathbf{t}}_p = \begin{pmatrix} 1 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & e^{\frac{j2\pi(p-1)(M-1)}{M}} \end{pmatrix} \begin{pmatrix} \hat{t}_p(0) \\ \vdots \\ \hat{t}_p(M-1) \end{pmatrix} \quad (4.17)$$

In Appendix D, it is shown that the m^{th} frequency tone of $\tilde{\mathbf{t}}_p$ is given by:

$$\tilde{t}_p(m) = \begin{cases} N_T \sum_{l=0}^{\frac{M}{N_T}-1} t_p(l) e^{-\frac{j2\pi(m-p-1)l}{M}} & \text{for } m = p-1, N_T + p-1, \dots \\ 0 & \text{Otherwise} \end{cases} \quad (4.18)$$

4.4.3 LST BS-CDMA CSI Estimator I: Tap Selection-Based

4.4.3.1 Estimation Problem: The Cost Function

The channel estimation problem is one that minimizes the following cost function:

$$\xi = \left\| \bar{\mathbf{z}}_q - \sum_{p=1}^{N_T} \bar{\mathbf{T}}_p \mathbf{F}_{M|L} \mathbf{h}_{qp|L} \right\|^2 \quad (4.19)$$

Differentiating the cost function w.r.t. $\mathbf{h}_{qp|L}$

$$\frac{\partial \xi}{\partial \mathbf{h}_{qp|L}} = -2 \left(\mathbf{F}_{M|L}^H \bar{\mathbf{T}}_p^H \right) \left(\bar{\mathbf{z}}_q - \sum_{p=1}^{N_T} \bar{\mathbf{T}}_p \mathbf{F}_{M|L} \mathbf{h}_{qp|L} \right) \quad (4.20)$$

and the estimate of channel parameters is obtained based on the following:

$$\begin{aligned} \left(\frac{\partial \xi}{\partial \mathbf{h}_{q1|L}} \cdots \frac{\partial \xi}{\partial \mathbf{h}_{qN_T|L}} \right)^T &= \mathbf{0}^T \\ \Rightarrow \mathbf{u}_q - \mathbf{Q} \mathbf{h}_q &= \mathbf{0} \\ \mathbf{h}_q &= \mathbf{Q}^{-1} \mathbf{u}_q \end{aligned} \quad (4.21)$$

where $\mathbf{u}_q$ and $\mathbf{Q}$ is defined respectively as

$$\mathbf{u}_q = \begin{pmatrix} \mathbf{u}_{q1} \\ \vdots \\ \mathbf{u}_{qN_T} \end{pmatrix} \text{ and } \mathbf{Q} = \begin{pmatrix} \mathbf{Q}_{11} & \cdots & \mathbf{Q}_{1N_T} \\ \vdots & \ddots & \vdots \\ \mathbf{Q}_{N_T1} & \cdots & \mathbf{Q}_{N_TN_T} \end{pmatrix} \quad (4.22)$$

and $\mathbf{u}_{qp} = \mathbf{F}_{M|L}^H \bar{\mathbf{T}}_p^H \bar{\mathbf{z}}_q$, $\mathbf{Q}_{p_1 p_2} = \mathbf{F}_{M|L}^H \bar{\mathbf{T}}_{p_1}^H \bar{\mathbf{T}}_{p_2} \mathbf{F}_{M|L}$, $\mathbf{h}_q = [\mathbf{h}_{q1}^T \cdots \mathbf{h}_{qN_T}^T]^T$ and $\tilde{\mathbf{h}}_q = [\tilde{\mathbf{h}}_{q1}^T \cdots \tilde{\mathbf{h}}_{qN_T}^T]^T$. Here, $\tilde{\mathbf{h}}$ is the channel estimate that minimizes the MSE. It can be seen that one major limitation of this technique is the inversion of a $N_T L \times N_T L$ size matrix. This means that computation complexity is proportional to the length of CP and transmitter diversity used.

4.4.3.2 Estimation Problem: The Solution Algorithm

In a practical deployment, the channel vector $\mathbf{h}_{qp|L}$ contains mostly null, except at few indices. When information about channel delay profile is available, those indices containing non-null coefficients are known and can be used to reduce the size of matrix inversion considerably. Let the number of significant taps be L_0 . The cost function in Eqn. (4.19) can be re-expressed as:

$$\xi = \|\bar{\mathbf{z}}_q - \sum_{p=1}^{N_T} \bar{\mathbf{T}}_p \hat{\mathbf{F}}_M \hat{\mathbf{h}}_{qp}\|^2 \quad (4.23)$$

where $\hat{\mathbf{h}}_{qp} = (\hat{h}_{qp}(l_0) \cdots \hat{h}_{qp}(l_{L_0-1}))^T$ is the channel vector consisting only L_0 significant taps, $\hat{\mathbf{F}}_M$ is of size $M \times L$ and $[\hat{\mathbf{F}}_M]_{ab} = e^{-\frac{j2\pi a l_b}{M}}$. Following similar steps described earlier, the new estimated channel parameter that minimizes MSE is given by:

$$\begin{pmatrix} \hat{\mathbf{h}}_{q1} \\ \vdots \\ \hat{\mathbf{h}}_{qN_T} \end{pmatrix} = \begin{pmatrix} \hat{\mathbf{Q}}_{11} & \cdots & \hat{\mathbf{Q}}_{1N_T} \\ \vdots & \ddots & \vdots \\ \hat{\mathbf{Q}}_{N_T1} & \cdots & \hat{\mathbf{Q}}_{N_TN_T} \end{pmatrix}^{-1} \begin{pmatrix} \hat{\mathbf{u}}_{q1} \\ \vdots \\ \hat{\mathbf{u}}_{qN_T} \end{pmatrix} \quad (4.24)$$

where $\hat{\mathbf{u}}_{qp} = \hat{\mathbf{F}}_M^H \bar{\mathbf{T}}_p^H \bar{\mathbf{z}}_q$ and $\hat{\mathbf{Q}}_{p_1 p_2} = \hat{\mathbf{F}}_M^H \bar{\mathbf{T}}_{p_1}^H \bar{\mathbf{T}}_{p_2} \hat{\mathbf{F}}_M$. The size of the matrix to be inverted is now considerably reduced to $N_T L_0 \times N_T L_0$.

4.4.3.3 Performance Analysis: Error Bound Evaluation

Since the received signal is made up of superposition between signals transmitted from N_T antennae, the correlation vector $\mathbf{u}_{qp}$ can be decomposed into:

$$\begin{aligned}\mathbf{u}_{qp} &= \mathbf{F}_{M|L}^H \bar{\mathbf{T}}_p^H \left(\sum_{p1=1}^{N_T} \bar{\mathbf{T}}_{p1} \mathbf{F}_{M|L} \mathbf{h}_{qp1|L} + \mathbf{n}_q \right) \\ &= \left(\mathbf{Q}_{p1} \cdots \mathbf{Q}_{pN_T} \right) \left(\mathbf{h}_{q1|L}^T \cdots \mathbf{h}_{qN_T|L}^T \right)^T + \mathcal{N}_p\end{aligned}\quad (4.25)$$

where $\mathcal{N}_p = \mathbf{F}_{M|L}^H \bar{\mathbf{T}}_p^H \bar{\mathbf{n}}_q$ is the noise term. With this and by letting $\mathcal{N} = \left(\mathcal{N}_1^T \cdots \mathcal{N}_{N_T}^T \right)^T$, $\mathbf{u}_q$ and $\tilde{\mathbf{h}}_q$ can be expressed as:

$$\begin{aligned}\mathbf{u}_q &= \left(\mathbf{u}_{q1}^T \cdots \mathbf{u}_{qN_T}^T \right)^T \\ &= \mathbf{Q} \mathbf{h}_q + \mathcal{N}\end{aligned}\quad (4.26)$$

$$\tilde{\mathbf{h}}_q = \mathbf{Q}^{-1} \mathbf{u}_q = \mathbf{h}_q + \mathbf{Q}^{-1} \mathcal{N} \quad (4.27)$$

The mean of the channel estimate is

$$E \left\{ \tilde{\mathbf{h}}_q \right\} = E \left\{ \mathbf{h}_q \right\} + E \left\{ \mathbf{Q}^{-1} \mathcal{N} \right\} = E \left\{ \mathbf{h}_q \right\} \quad (4.28)$$

Similarly to [138], the MSE of the channel estimate is given as:

$$MSE = \frac{1}{N_T L} E \left\{ \left\| \tilde{\mathbf{h}}_q - \mathbf{h}_q \right\|^2 \right\} = \frac{\sigma^2}{N_T L} \text{Tr} \left\{ \mathbf{Q}^{-1} \right\} \quad (4.29)$$

where $\text{Tr} \{ \cdot \}$ corresponds to the Trace of matrix $\{ \cdot \}$ and σ^2 is the variance of the AWGN component.

4.4.3.4 Performance Analysis: Optimum Pilot Design

For the general case of $N_T > 2$, derivation of $\mathbf{Q}^{-1}$ is non-trivial. However, for $N_T = 2$, if the pilot tones are assumed to be constant, i.e. $|\bar{t}_p(l)|^2 = \gamma$ and thus $\mathbf{Q}_{pp} = \gamma \mathbf{I}_{L \times L}$, $\mathbf{Q}^{-1}$

can easily be derived as:

$$\mathbf{Q}^{-1} = \frac{1}{M\gamma} \begin{pmatrix} \mathbf{I}_{L \times L} + \frac{\mathbf{Q}_{12}}{M\gamma} \left(\mathbf{I}_{L \times L} - \frac{\mathbf{Q}_{12}^H \mathbf{Q}_{12}}{M^2 \gamma^2} \right)^{-1} \frac{\mathbf{Q}_{12}^H}{M\gamma} & \frac{\mathbf{Q}_{12}}{M\gamma} \left(\mathbf{I}_{L \times L} - \frac{\mathbf{Q}_{12}^H \mathbf{Q}_{12}}{M^2 \gamma^2} \right)^{-1} \\ - \left(\mathbf{I}_{L \times L} - \frac{\mathbf{Q}_{12}^H \mathbf{Q}_{12}}{M^2 \gamma^2} \right)^{-1} \frac{\mathbf{Q}_{12}^H}{M\gamma} & \left(\mathbf{I}_{L \times L} - \frac{\mathbf{Q}_{12}^H \mathbf{Q}_{12}}{M^2 \gamma^2} \right)^{-1} \end{pmatrix} \quad (4.30)$$

Let $\{\lambda_l^2\}$ for $l = 0$ to $L - 1$ be the set of eigenvalues for $\mathbf{Q}_{12}^H \mathbf{Q}_{12}$. The MSE is then found to be

$$MSE = \frac{\sigma^2}{2M\gamma L} \left(L + \sum_{l=0}^{L-1} \frac{M^2 \gamma^2 + \lambda_l^2}{M^2 \gamma^2 - \lambda_l^2} \right) \quad (4.31)$$

It can easily be deduced that minimum MSE occurs when $\lambda_l = 0$ for $l = 0$ to $L - 1$ or equivalently

$$MSE \geq \frac{\sigma^2}{2M\gamma L} (2L) = \frac{\sigma^2}{M\gamma} \quad (4.32)$$

when pilots with such property are used, they are said to be optimum because since $\mathbf{Q}_{12}^H \mathbf{Q}_{12} = \mathbf{0}$, ICI is totally eliminated from the system. In addition, matrix inversion consists only of trivial division as $\mathbf{Q}$ is now a diagonal matrix, $\mathbf{Q} = \text{diag}\{\mathbf{Q}_{11} \mathbf{Q}_{22}\}$. One possible pilot design is to let $\bar{t}_2(m) = (-1)^m \bar{t}_1(m)$. To construct such pilots, the phase-rotated repeated finite-length sequence presented earlier can be used by adopting the following procedure:

1. Generate two arbitrary sequences of length $(\frac{M}{2})$ and let the two sequences be denoted $\mathbf{t}_1^1$ and $\mathbf{t}_1^2$ respectively.
2. Repeat or append $\mathbf{t}_1^1$ with itself to generate $\hat{\mathbf{t}}_1^1$ of length M .

$$\hat{\mathbf{t}}_1^1 = \left((\mathbf{t}_1^1)^T (\mathbf{t}_1^1)^T \right)^T$$
3. Perform step 2 on $\mathbf{t}_1^2$ to generate $\hat{\mathbf{t}}_1^2$ $\hat{\mathbf{t}}_1^2 = \left((\mathbf{t}_1^2)^T (\mathbf{t}_1^2)^T \right)^T$
4. Phase-rotate $\hat{\mathbf{t}}_1^1$ and $\hat{\mathbf{t}}_1^2$ by θ_0 and θ_1 to generate $\tilde{\mathbf{t}}_1^1$ and $\tilde{\mathbf{t}}_1^2$ respectively.
5. Let $\tilde{\mathbf{t}}_2^1 = \tilde{\mathbf{t}}_1^1$ and $\tilde{\mathbf{t}}_2^2 = -\tilde{\mathbf{t}}_1^2$.
6. Pilot symbols to be transmitted from antenna 1 and 2 are

$$\tilde{\mathbf{t}}_1 = \tilde{\mathbf{t}}_1^1 + \tilde{\mathbf{t}}_1^2 \quad \tilde{\mathbf{t}}_2 = \tilde{\mathbf{t}}_2^1 + \tilde{\mathbf{t}}_2^2$$

Instead of any arbitrary sequence, Chu's sequence [139] can be used to generate constant amplitude tone $|\bar{t}_p(m)|^2 = \gamma$ for $m = 0$ to $M - 1$. This is to ensure that all the pilot tones are the same and hence uniform attenuation is applied to the noise across all sub-carriers.

By constructing the pilot sequence according to the above procedure, it can be easily derived that $\bar{t}_2(m) = (-1)^m \bar{t}_1(m)$. The error performance and pilot design in the present analysis assumes that $N_T = 2$. For $N_T > 2$, the optimum pilot design can be obtained by first deriving $\mathbf{Q}^{-1}$ and then minimizing the ICI components.

4.4.4 LST BS-CDMA CSI Estimator II: Improved Channel Estimation

In the two methods described earlier, each antenna transmits pilot sequences with non-null components at all frequency tones. It is in this context, a new method is suggested for each antenna to transmit pilot sequence with non-null components only at pre-defined indices. The indices defined should be mutually exclusive among all antennae. The construction procedure for defining such sequences has been defined in Section 4.4.2. More details on this modified channel estimator is discussed in the following section.

4.4.4.1 Mathematical Model

Let the pre-defined non-null tone index for the p^{th} antenna be $\Omega_p = \{kN_T + p\}$ for $k = 0$ to $\frac{M}{N_T} - 1$ and $\bar{\mathbf{t}}_p$ be the pilot sequence generated using the procedure suggested in Section 4.4.2. After despreading and FFT operation, the received signal is "clear" from ICI and the received signal at subcarrier $k \in \Omega_p$ can be written as:

$$\bar{\mathbf{z}}_{q|\Omega_p} = \bar{\mathbf{T}}_{p|\Omega_p} \mathbf{F}_{M|L|\Omega_p} \mathbf{h}_{qp|L} \quad (4.33)$$

where $\bar{\mathbf{z}}_{q|\Omega_p}$ and $\mathbf{F}_{M|L|\Omega_p}$ consists of rows (row index $\in \Omega_p$) from $\bar{\mathbf{z}}_q$ and $\mathbf{F}_{M|L}$ respectively. $\bar{\mathbf{T}}_{p|\Omega_p} = \text{diag} \{\bar{t}_p(kN_T + p)\}$ for $k = 0$ to $\frac{M}{N_T} - 1$. Since non-null tones from different antenna exist at different indices, channel estimation can be carried out individually for each antenna without interference from other antennae. The cost function can then be defined individually for each transmit antenna as:

$$\xi_p = \| \bar{\mathbf{z}}_{q|\Omega_p} - \bar{\mathbf{T}}_{p|\Omega_p} \mathbf{F}_{M|L|\Omega_p} \mathbf{h}_{qp|L} \|^2 \quad (4.34)$$

After differentiating the cost function, estimate of the channel parameter is obtained as:

$$\begin{aligned} \mathbf{u}_{qp|\Omega_p} - \mathbf{Q}_{pp|\Omega_p} \mathbf{h}_{qp|L} &= 0 \\ \mathbf{h}_{qp|L} &= \mathbf{Q}_{pp|\Omega_p}^{-1} \mathbf{u}_{qp|\Omega_p} \end{aligned} \quad (4.35)$$

where $\mathbf{Q}_{pp|\Omega_p} = \mathbf{F}_{M|L|\Omega_p}^H \bar{\mathbf{T}}_{p|\Omega_p}^H \bar{\mathbf{T}}_{p|\Omega_p} \mathbf{F}_{M|L|\Omega_p}$ and $\mathbf{u}_{qp|\Omega_p} = \mathbf{F}_{M|L|\Omega_p}^H \bar{\mathbf{T}}_{p|\Omega_p}^H \bar{\mathbf{z}}_{q|\Omega_p}$. It can be seen that the size of matrix to be inverted is $L \times L$ or independent of the number of transmit antenna. If knowledge of channel delay spread (CDS) is available, the simplified tap-selection approach can be used to reduce computational complexity further.

4.4.4.2 Pilot Design

Figure 4.3: Pilot Tones Arrangement

In the improved method, each antenna transmits a pilot sequence with non-null components at pre-defined indices, unique to itself. The index set assigned to p^{th} transmit antenna is $\Omega_p = \{p, N_T + p, \dots\}$. The pilot sequences are repeated finite-length sequences constructed from the steps described in Section 4.4.2.1. Before transmission, pilot sequence to be transmitted from the p^{th} antenna is phase-rotated by θ_{p-1} , $p = 1$ to N_T as described

in Section 4.4.2.2. For illustration purpose, the pilot tones transmitted from N_T ($N_T = 4$) antennae are given as shown in Figure 4.3.

4.4.4.3 Performance Analysis

The estimator proposed here minimizes MSE across all subcarriers w.r.t. all N_T channels. However, derivation of MSE for $N_T > 2$ is non-trivial and only the special case when $N_T = 2$ is considered. In the improved method, pilot tones from different transmit antennae do not superimpose or interfere with one another. The size of matrix inversion is now reduced by a factor of N_T and no ICI is present. This section presents the error performance of the improved estimator and the analysis applies for the general case of $N_T \geq 1$. The MSE of the channel estimator is given as:

$$\begin{aligned}
 MSE &= \frac{1}{L} E \left\{ \left\| \tilde{\mathbf{h}}_{qp|L} - \mathbf{h}_{qp|L} \right\|^2 \right\} \\
 &= \frac{1}{L} E \left\{ (\mathbf{Q}_{pp}^{-1} \mathcal{N}_p)^H (\mathbf{Q}_{pp}^{-1} \mathcal{N}_p) \right\} \\
 &= \frac{\sigma^2}{L} \text{Tr} \left\{ \mathbf{Q}_{pp}^{-1} \right\} \\
 &= \frac{\sigma^2}{M\gamma}
 \end{aligned} \tag{4.36}$$

Now, assume that pilot sequence of length $(\frac{M}{N_T})$ for each antenna is generated based on Chu's sequence with constant amplitude γ_0 . It can easily be derived from Eqn (4.18) that $\bar{t}_p(m) = \gamma = N_T \gamma_0$ for $m \in \Omega_p$. Thus, estimation accuracy improves when transmitter diversity increases.

4.5 Numerical Results

4.5.1 Perfect CSI

The LST BS-CDMA architecture was modelled and its performance was investigated using monte-carlo simulation for both uplink and downlink deployment. The entire simulation studies target high data rate and high mobility communication systems that deliver better system throughput compared to the basic single antenna BS-CDMA system. The transmission carrier frequency considered here is 5GHz, with a chip-rate of 40Mcps. As highlighted

earlier, a system with LST arrangement benefits from the transmission rate enhancement as additional data are transmitted from the additional antenna. Two different antenna configurations are used for the LST BS-CDMA architecture, namely 2×2 ($N_T = N_R = 2$) and 4×4 ($N_T = N_R = 4$). Two different block sizes are considered, namely $M = 256$ and $M = 512$. The length of CP here is $L_{CP} = \frac{1}{4}M$. This effectively means that a single chip-block is of size 320 and 640 respectively. For $M = 256$, the duration of each chip-block, T , is $8\mu\text{secs}$, where $6.4\mu\text{secs}$ is the duration of information chips and $1.6\mu\text{secs}$ is the duration of guard interval. At the receiver, the system bandwidth of 40MHz is divided into M subchannels and equalized individually. The processing gain used are 8, 16 and 32. The number of actively transmitting users is varied accordingly to simulate different loading conditions in the system. The channel model assumes *Jakes' Rayleigh* distribution [132] and the number of path is assumed to be 24. It is assumed that the maximum delay spread is less than the guard duration so that IBI is avoided completely. Different vehicular mobility conditions are also used to simulate the practical deployment of typical mobile communication system, namely 0Hz (stationary), 463Hz (medium-high) and 926Hz (high). The modulation scheme used is BPSK and QPSK. When the processing gain is 32, this translates to each user transmitting at bandwidth efficiency of 0.125bps/Hz and 0.25bps/Hz. The complete set of parameters used for simulation studies is tabulated in Table 4.1.

The performance of 2×2 LST BS-CDMA system is depicted in Figure 4.4. In Figure 4.4(a), the performance of the system with per-tone ZF equalizer is plotted. Under non-mobile environment, the loading condition does not have any implication on the error rate performance of the system. Under moderately high mobility environment ($F_D = 463\text{Hz}$), significant performance degradation is observed. However, an error rate of 9% is still achievable with SNR value of 20dB. From the Figure, it can be seen that the performance of half-load condition under severely high mobility ($F_D = 926\text{Hz}$) is almost identical to that of full-load condition with Doppler spread of 463Hz . However, under full-load condition, the system exhibits an error rate that is greater than 10% when $F_D = 926\text{Hz}$.

Figure 4.4(b) shows the performance of 2×2 LST BS-CDMA system with per-tone MMSE equalizer. Again, the curves show that the system performance is independent of the system load when the channel is stationary. Under moderately high mobility, the system performance degrades by about 1dB when the system load is half and an error floor of about 3% is observed. Similarly, the half-load performance with Doppler spread of 926Hz is close to identical to the full-load performance with Doppler spread of 463Hz . Under full-load

(a) Performance of ZF Equalizer

(b) Performance of MMSE Equalizer

Figure 4.4: BER Performance of LST BS-CDMA System (2×2)

Table 4.1: LST CDMA Systems: Simulation Parameters

	System I	System II
Carrier Frequency, F	5 GHz	5GHz
Chip rate, F_C	40.96 Mcps	40.96 Mcps
Number of Tx Antenna, N_T	2	4
Number of Rx Antenna, N_R	2	4
Modulation	BPSK — QPSK	BPSK — QPSK
OFDM / Chip-Block Size, M	256	256
OFDM / Chip-Block Size, M	256	256
	(512)	(512)
FFT / IFFT Size, M	256	256
	(512)	(512)
OFDM Block / Chip-Block Duration, T	8 μsecs	8 μsecs
	(16 μsecs)	(16 μsecs)
Data Duration per OFDM Block, T_D	6.4 μsecs	6.4 μsecs
	(12.8 μsecs)	(12.8 μsecs)
Guard Duration per OFDM Block, T_{CP}	1.6 μsecs	1.6 μsecs
	(3.2 μsecs)	(3.2 μsecs)
Super-Block Size, M_{SB}	1 OFDM Block	PG OFDM Blocks
Super-Block Duration, T_{SB}	8 μsecs	PG $\times$ 8 μsecs
	(16 μsecs)	(PG $\times$ 16 μsecs)
Processing Gain, PG	8 16 32	8 16 32
Symbols per Super-Block, N_{SB}	32 16 8	256
	(64 32 16)	(512)
Load Condition	1 user Half-Load Full-Load	1 user Half-Load Full-Load
Doppler Spread, F_D	0 Hz 463 Hz 926 Hz	0 Hz 463 Hz 926 Hz
Vehicular Mobility, V	0 kph 100 kph 200 kph	0 kph 100 kph 200 kph
Channel Coherence Time, T_{CO}	∞ 387 μsecs 193 μsecs	∞ 387 μsecs 193 μsecs

condition, the MMSE equalizer exhibits an error floor that is slightly above 10%.

The performance of 4×4 LST BS-CDMA system is depicted in Figure 4.5. As shown in Figure 4.5(a), the system performance with ZF equalizer is generally poor. This is attributed to the fact that neither error-correcting codes nor advanced iterative algorithm (such as the one proposed in [79]) is used. As a result, the interference from the multiple transmitting antennae leads to the poor performance.

Figure 4.5(b) shows the performance of 4×4 LST BS-CDMA system with per-tone MMSE equalizer. During stationary condition, the error rate of 1% is achievable with SNR of approximately 17dB. During moderately high mobility environment, performance degradation of about 4dB is observed when there are 16 transmitting users for BER of 1%. When

(a) Performance of ZF Equalizer

(b) Performance of MMSE Equalizer

Figure 4.5: BER Performance of LST BS-CDMA System (4×4)

32 users are transmitting, an error floor above 1% is observed. Meanwhile, the system performance experiencing high mobility has an error floor that is above 10%.

4.5.2 Imperfect CSI

Extensive simulation of LST BS-CDMA system was conducted to investigate how the system performance is affected by the imperfect CSI. The detailed system parameters used in the exercise is first laid out, followed by discussion of the results obtained from the simulation studies.

In order to illustrate the performance degradation caused by ICI, LST BS-CDMA system with transmitter diversity of 2 ($N_T = 2$) and 4 ($N_T = 4$) is simulated. The number of antennas employed at the receiver is assumed to be equal to that at the transmitter. The length of spread codes used is 32 ($G = 32$) with one of them assigned for pilot. The number of users is fixed as 16, which corresponds to half-load condition. The assumption of channel stationarity, which is the underlying principle behind LST BS-CDMA is not valid in a mobile environment. The degree of non-stationarity depends on the Doppler frequency and the block size used. In the simulation, the bandwidth of 4 MHz is divided into M , where M refers to the block size used. M is varied from 64, 128 and 256. The doppler frequency is varied from 40 and 200Hz. A guard interval is fixed as one-quarter of the block size ($L = \frac{1}{4}M$). The channel used is an L_0 paths Rayleigh fading channel simulated based on Jakes' model [132] ($L_0 = \frac{1}{4}L$). Total duration of each spreading module is 0.64, 1.28 and 2.56 msec respectively. BPSK modulation is used. For system with 2 and 4 transmitter diversity, the effective chip rate is 6.4 and 12.8 Mcps over a 4 MHz channel.

The quality of the CSI obtained using the two different estimation techniques is first investigated and the result is shown in Figure 4.6. The impact of block size on the performance of improved estimation technique is investigated next and the result is shown in Figure 4.7. Lastly, the BER performance obtained using perfect CSI and estimated CSI is investigated and the result is shown in Figure 4.8 and 4.9.

In Figure 4.6, the MSE of the channel estimates obtained using the two different estimation methods namely the tap-selection estimator (TSE) and improved estimator (IE), is plotted against SNR values. For this exercise, the block size is fixed as 128 ($M = 128$) and Doppler frequency is fixed as 40Hz. From the Figure, it can be seen that for a LST BS-CDMA system

with two transmitter and receiver diversity, IE performs slightly better than TSE when random pilots are used. The use of random pilots in TSE yields CSI estimation quality that is slightly inferior to that obtained using IE. This performance difference confirms earlier hypothesis that points out the shortcoming of the basic approach due to the presence of ICI. As shown in Figure 4.6, the performance of TSE is slightly superior than IE when the optimum pilot is used. On the other hand, the quality of the channel estimates obtained using the improved technique increases as the number of transmit antenna is increased. This is due to the higher amount of energy that is packed at those non-null tones, making the effective SNR value at non-null tones higher and hence better quality estimates.

Figure 4.6: MSE Comparison across different Methods

Figure 4.7 shows the MSE performance of IE when the block size is varied between 64, 128 and 256. From the Figure, it can be seen that channel estimation accuracy generally improves as the block size is increased. This is because for a given order of transmitter diversity, the increase in block size effectively means that there will be more number of tones that are involved to obtain the channel estimates. The averaging effect contributes to the increased accuracy. However, it is observed that the gradient of the MSE curve decreases when $M = 256$. This is attributed to two factors. Firstly, the MSE is approaching its lower bound and there will come to a point when any increase in block size yields no significant improvement. Secondly, the underlying principle behind LST BS-CDMA is the channel

stationarity during a single spreading module. However, as block size increases, so does the total duration of a single spreading module. Thus, higher degree of channel variation contributes to the degradation of the estimation accuracy.

Figure 4.7: Performance of Improved Estimator, $F_d = 40\text{Hz}$

Next, the BER performance of LST BS-CDMA obtained using perfect knowledge of CSI and that obtained using estimated (imperfect) knowledge of CSI is being compared. The comparison is made using two deployment configurations. In the first configuration, the Doppler frequency is set at 40Hz which corresponds to low vehicular mobility of about 21kmph. The block size used is 128 ($M = 128$). In the second configuration, the Doppler frequency is set at 200Hz which corresponds to high vehicular velocity of around 108kmph. Since the channel coherence time is expected to be much shorter for high Doppler frequency [132], the block size is chosen as 64 ($M = 64$) in order to minimize the amount of channel variation that is experienced during a single block transmission.

The BER performance obtained from the first configuration is shown in Figure 4.8, with Figure 4.8(a) and 4.8(b) showing the results for ZF and MMSE technique respectively. From Figure 4.8(a), it can be seen that for 2×2 system, the BER performance resulting from the use of imperfect CSI obtained using TSE and IE do not show significant difference. Their performance closely match the BER performance obtained when perfect knowledge of CSI

(a) Performance of ZF Equalizer

(b) Performance of MMSE Equalizer

Figure 4.8: Performance of LST BS-CDMA with Doppler Frequency, $F_d = 40\text{Hz}$ and Block Size, $M = 128$

is assumed. However, it can be seen from the Figure that for 4×4 system, the BER performance obtained using estimated CSI from the improved technique shows slight superiority compared to the performance obtained using tap-selection approach. From 4.8(b), for 2×2 system, the use of channel estimates obtained using TSE with optimum pilot results in the best performance that is closest to that obtained using the tap-selection method. At BER of 10^{-3} , the gap is about 2dB.

Figure 4.9 shows the BER comparison based on the second configuration, with Figure 4.9(a) and 4.9(b) showing the result for ZF and MMSE technique respectively. Similar to earlier observation, the BER performance for 2×2 system with ZF equalizer shows insignificant difference regardless of the technique used to obtain the CSI estimates. For 4×4 system, the higher estimation accuracy obtained using IE results in slightly better performance than that obtained using TSE. From Figure 4.9(a), it can be seen that when the estimated CSI is used, the BER performance for 2×2 system is about 1dB worse (at BER of 10^{-3}) than the performance obtained with perfect CSI. This gap is relatively constant regardless of the technique used to obtain the CSI estimates. For 4×4 system, the observation is in line with earlier results. The BER performance obtained using the improved estimator is better than that obtained using the tap-selection estimator.

Based on the results presented here, it can be concluded that the proposed estimation algorithms are capable of producing fairly accurate estimates of the CSI and the imperfect estimates do not cause significant degradation to the BER performance of LST BS-CDMA system.

4.5.3 SISO BS-CDMA vs LST BS-CDMA: Performance Comparison

In order to illustrate the performance enhancement that can be offered by LST BS-CDMA system, its performance is compared with SISO BS-CDMA system. For this comparison, both systems are configured to transmit with the same chip rate. This is achieved by the use of 16-QAM modulation in SISO BS-CDMA and QPSK modulation for 2×2 LST BS-CDMA system. The comparison is made based on the BER performance obtained by using MMSE equalizers.

Figure 4.10 shows the performance comparison between the two system under different loading condition, with Figure 4.10(a), 4.10(b) and 4.10(c) showing the case for single-user

(a) Performance of ZF Equalizer

(b) Performance of MMSE Equalizer

Figure 4.9: Performance of LST BS-CDMA with Doppler Frequency, $F_d = 200\text{Hz}$ and Block Size, $M = 64$

Figure 4.10: SISO BS-CDMA vs LST BS-CDMA - A Performance Comparison

(SU), half-load and full-load respectively. From the figures, the error-rate advantage offered by LST BS-CDMA is distinctly significant. During SU scenario, the SNR value required to achieve $\text{BER}=10^{-2}$ for LST BS-CDMA is more than 6dB lower than that required by SISO BS-CDMA system. As shown in Figure 4.10(b), such performance advantage is more significant under half-load scenario. Figure 4.10(c) shows that this advantage is consistently retained under full-load scenario. Based on the comparison presented here, it can be clearly stated that when the use of multiple antenna is allowed, the use of LST architecture combined with BS-CDMA can potentially offer significant performance enhancement.

4.6 Summary

In this Chapter, a MIMO BS-CDMA architecture adopting the LST approach has been presented and referred to as LST BS-CDMA. The fundamental principle behind LST BS-CDMA is to deliver higher data rate by transmitting different data from different antennae simultaneously. This means that a LST BS-CDMA system employing N_T antennae at the transmitter delivers data rate that is N_T times higher than a BS-CDMA system with similar configuration but employing only a single antenna instead. Various channel estimation algorithms that allow accurate estimates of the CSI to be obtained efficiently were also described. Based on the results obtained from the simulation studies, it can be seen that the performance degradation encountered in LST BS-CDMA system with high Doppler frequency can be quite severe. In this context, the use of adaptive modulation and coding (AMC) scheme to improve overall system performance can be considered as part of the future works. With AMC scheme, higher modulation and lower code rate (and hence higher transmission rate) can be assigned to those users who are expected to perform well, i.e. during low mobility condition. On the other hand, lower modulation and higher code rate (and hence lower transmission rate) can be used when the system is expected to perform poorly, i.e. during high mobility condition.

The performance of LST BS-CDMA with different antenna configuration is compared against that of SISO BS-CDMA in terms of the required SNR to achieve a certain level of BER. However, it is important to note that the per-tone equalizer used in the LST BS-CDMA can still be improved by using iterative algorithms. In addition, coding such as convolution code with different code rate can also be used to improve the performance. Fur-

ther works is required to obtain a more comprehensive understanding on the performance of the SISO and LST BS-CDMA systems.

The contribution in this chapter can be summarized by the following points:

- a layered space-time architecture for BS-CDMA system is presented and referred to as LST BS-CDMA;
- the problem of channel estimation and pilot design in LST BS-CDMA is studied;
- using overaly pilot transmission, two channel estimation algorithms are proposed;
- performance of LST BS-CDMA system with different antenna configuration and mobility pattern obtained using extensive simulation study is presented;
- the accuracy of the channel estimates obtained using the different estimation approach is presented and compared in terms of the mean square error (MSE); and
- the impact of imperfect CSI on the system performance is presented.

Chapter 5

Space-Time Coded Block Spread CDMA (STC BS-CDMA) Systems

5.1 Introduction

In Chapter 4, LST BS-CDMA architecture is presented. LST BS-CDMA system exploits the available additional antennae to send data independently and hence increasing the spectral efficiency. The second class of MIMO architecture, namely Space-Time Block Coding (STBC), attempts to achieve a similar impact using a different philosophy. STBC systems focus on improving the signal quality reception and hence translating it into an improved BER performance. This would then allow higher modulation scheme such as 64-QAM or higher to be used. This Chapter presents the implementation of BS-CDMA system in such a MIMO environment, which will be referred to as STC BS-CDMA.

Chapter 5 is organized as follows. Section 5.2 describes the motivation behind the proposal of STC BS-CDMA. Section 5.3 lays out the system architectural model of STC BS-CDMA. Section 5.4 discusses the results from simulation studies.

5.2 Motivation

As highlighted in Chapter 4, future mobile broadband communication system demands extremely high transmission rate capability. However, given the scarcity of frequency spectrum, the regulatory frameworks limit the amount of bandwidth that is allocated to a particular application or standards. This necessitates the use of a transmission scheme that is capable of achieving higher spectral efficiency.

In this context, a particular MIMO architecture, namely LST, has been considered in Chapter 4 to enhance system capacity of BS-CDMA system. A LST architecture attempts to increase spectral utilization by treating each additional antenna as a new transmission channel and hence additional information can be transmitted. For example, a transmission scheme with an efficiency of 1 bps/Hz when using a single antenna can be improved an efficiency of 2 bps/Hz when an additional antenna is available. But the signal recovery at the receiver for such cases will be of higher complexity as additional process of segregating data from different transmit antenna is now required on top of the channel compensation process.

Alternatively, transmission efficiency can be attained through the use of higher modulation schemes. For example, if a system with QPSK modulation achieves 1 bps/Hz efficiency, then the same system with 16-QAM or 64-QAM modulation could potentially achieve an efficiency of 4 bps/Hz or 6 bps/Hz respectively. However, the cost that comes with such higher modulation scheme is the degradation in BER performance.

In the past, technological limitation and the unavailability of transmitter diversity technique that allows simple decoding at the receiver, has led to the incorporation of multiple antenna being exclusively applied to the BS receiver. Since a BS receiver is shared by many users and has a fixed location, the cost and complexity constraint is less stringent than that of a mobile devices. It was only recently that *Alamouti* [101] proposed a simple space-time coding technique that allows simple recovery process at the receiver. At the same time, the method was able to exploit the full spatial diversity that comes with two antenna systems. With the significant improvement in the BER performance, such space-time architecture allows the use of higher modulation scheme as mentioned earlier.

The original work proposed in [101] caters only to communication systems experiencing

flat-fading channels. Due to its tremendous potential, space-time coded transmission has since then attracted enormous interest among the research community [140]. One focus area is to extend the Alamouti concept for frequency-selective channels. A low complexity implementation of Alamouti scheme for OFDM system was first proposed in [112]. Based on similar mathematical principles, Alamouti coding scheme can also be applied to SC-FDE system, which is the backbone of BS-CDMA system.

The advantages of BS-CDMA as a candidate for high-rate communication systems, particularly for uplink deployment, has been thoroughly addressed in the previous Chapters. In order to enhance the performance of BS-CDMA further, a MIMO architecture that incorporates space-time encoder into a BS-CDMA system was proposed in [141] and named STC BS-CDMA. Details of STC BS-CDMA architecture is presented in the following sections.

5.3 System Model

This section first describes the transmission structure of a STC BS-CDMA system when the order of transmitter diversity is two. The signal processing algorithm to recover the transmitted data is presented. The last part of this section describes the space-time coding algorithm for the arbitrary order of transmitter diversity.

5.3.1 Transmitter Model

Figure 5.1: A Generalized Baseband STC BS-CDMA Transmitter Architecture

Figure 5.1 illustrates the generalized baseband model of a STC BS-CDMA transmitter. As of today, it is well established that the optimum rate (Rate = 1) for a STBC system is only achievable with the use of two transmit antennae. When higher antenna diversity is used,

the rate drops to $\frac{3}{4}$, $\frac{1}{2}$ or even less. As such, this thesis focuses on the use of two antennae as shown in Figure 5.1. However, a procedure to construct a generalized STBC encoder for arbitrary number of antenna is given in Section 5.3.3.

First, let $d_k(n)$ denotes the n^{th} information symbol belonging to the k^{th} user. The single continuous stream of information symbols is fed into a STBC encoder. The task of the STBC encoder is to apply Alamouti encoding to the single stream of symbols and eventually output the encoded information as two parallel streams. In essence, the STBC encoder breaks the continuous stream of symbols into chunks of $2M$ symbols, $\mathbf{d}_k(i)$ as:

$$\begin{aligned}\mathbf{d}_k(i) &= \begin{bmatrix} d_k(i2M+1) & \cdots & d_k(i2M+M) & d_k(i2M+M+1) & \cdots & d_k(i2M+2M) \end{bmatrix}^T \\ &= \begin{bmatrix} \mathbf{d}_{k,1}^T & \mathbf{d}_{k,2}^T \end{bmatrix}^T\end{aligned}\quad (5.1)$$

where i is related to n by $n = i2M + m$. i , M and m refers to the time index, block size and the index entry in $\mathbf{d}_k(i)$ respectively. For brevity, i will be omitted in subsequent discussions. From the $2M \times 1$ symbols, the STBC encoder constructs a $2M \times 2$ matrix, $\mathbf{D}_k$ as:

$$\mathbf{D}_k = \begin{bmatrix} \mathbf{d}_{k,1} & \mathbf{d}_{k,2} \\ -\mathbf{d}_{k,2}^{(<m>)} & \mathbf{d}_{k,1}^{(<m>)} \end{bmatrix} \quad (5.2)$$

where $\mathbf{d}_{k,p}^{(<m>)} = \mathbf{P}_M^{(1)} \mathbf{d}_{k,p}^*$ and $\mathbf{P}_M^{(1)}$ is a $M \times M$ permutation matrix as defined in [142]. The significance of such mathematical operation is that a time-reversal operation is performed on $\mathbf{d}_{k,p}^*$. Such an operation is needed in order to facilitate a simple space-time decoding process at the receiver. This will become clearer as it is described in detail shortly. The $2M \times 2$ matrix will then be output as two substreams, where the first column will be the first substream and the second column as the second substream output. Each substream will eventually be transmitted through a single antenna. Block spreading is then performed on each substream. Instead of operating on the $2M \times 1$ encoded symbols directly, the block spreading module operates on $M \times 1$ symbols at any one time and outputs a $G \times M$ chip matrix, $\mathbf{X}_{k,p,t}$, each time. As a result, the 4 chip matrix produced during two time epoch are given as:

$$\begin{aligned}\mathbf{X}_{k,1,1} &= \mathbf{c}_k \otimes \mathbf{d}_{k,1} & \mathbf{X}_{k,2,1} &= \mathbf{c}_k \otimes \mathbf{d}_{k,2} \\ \mathbf{X}_{k,1,2} &= \mathbf{c}_k \otimes -\mathbf{d}_{k,2}^{(<m>)} & \mathbf{X}_{k,2,2} &= \mathbf{c}_k \otimes \mathbf{d}_{k,1}^{(<m>)}\end{aligned}\quad (5.3)$$

As has been described in Chapter 3, the chips are then read out row-wise. The g^{th} chip-block to be transmitted from the p^{th} antenna during the t^{th} time epoch is represented by $\mathbf{x}_{k,p,t}(g)$,

which is essentially the g^{th} row of $\mathbf{X}_{k,p,t}$. For downlink transmitter, encoded symbols from all the K users are summed together. Before transmission, CP insertion is first performed and the g^{th} CP-appended chip-block, $\mathbf{x}_{k,p,t}^{(CP)}(g)$, is given as:

$$\mathbf{x}_{k,p,t}^{(CP)}(g) = \begin{bmatrix} \mathbf{I}_M^{(CP)T} & \mathbf{I}_M \end{bmatrix}^T \mathbf{x}_{k,p,t}(g) \quad (5.4)$$

where $\mathbf{I}_M^{(CP)}$ consists of the last L_{CP} rows of $\mathbf{I}_M$. To better illustrate this, a pictorial representation of the transmission structure is shown in Figure 5.2.

Figure 5.2: Block Transmission Structure of STC BS-CDMA System

5.3.2 Receiver Model

Figure 5.3: A Generalized Baseband STC BS-CDMA Receiver Architecture

The generalized baseband STC BS-CDMA receiver architecture is illustrated in Figure 5.3.

In the Figure, a system with two receive antennae is depicted for simplicity. However, if uplink deployment is intended, additional antenna could always be employed at the receiver to further improve the performance of the system. On the other hand, if only a single antenna is desired at the receiver, space-time coded system does not have the restriction of a LST system where $N_R \geq N_T$.

After CP removal, the received signal at the q^{th} receive antenna during the t^{th} interval ($t = 1, 2$) of the g^{th} chip-block time epochs is given as:

$$\begin{aligned} \mathbf{r}_{q,t}(g) &= \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{h}_{k,qp,t}(g) \otimes \mathbf{x}_{k,p,t}(g) + \mathbf{n}_{q,t}(g) \\ &= \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{H}_{k,qp,t}(g) \mathbf{x}_{k,p,t}(g) + \mathbf{n}_{q,t}(g) \end{aligned} \quad (5.5)$$

where $\mathbf{h}_{k,qp,t}(g) = [h_{k,qp,t}(g, 0) \ \cdots \ h_{k,qp,t}(g, M-1)]^T$ is k^{th} user's channel pair between the q^{th} receive and p^{th} transmit antenna during the corresponding period interval. $\mathbf{H}_{k,qp,t}(g)$ is a circular Toeplitz matrix with $\mathbf{h}_{k,qp,t}(g)$ as the first column entry. $\mathbf{n}_{q,t}(g)$ represents the AWGN component during the respective period interval. By substituting the respective transmitted vectors into Equation (5.5), the complete set of signals received during the g^{th} chip-block period can be expressed as:

$$\mathbf{r}_{q,1}(g) = \sum_{k=1}^K (\mathbf{H}_{k,q1,1}(g) c_{k,g} \mathbf{d}_{k,1} + \mathbf{H}_{k,q2,1}(g) c_{k,g} \mathbf{d}_{k,2}) + \mathbf{n}_{q,1}(g) \quad (5.6)$$

$$\mathbf{r}_{q,2}(g) = \sum_{k=1}^K \left(-\mathbf{H}_{k,q1,2}(g) c_{k,g} \mathbf{d}_{k,2}^{<m>} + \mathbf{H}_{k,q2,2}(g) c_{k,g} \mathbf{d}_{k,1}^{<m>} \right) + \mathbf{n}_{q,2}(g) \quad (5.7)$$

If the channel stationarity assumption is true, $\mathbf{H}_{k,qp,t}(1) \approx \cdots \mathbf{H}_{k,qp,t}(G) = \mathbf{H}_{k,qp,t}$ and hence, $\mathbf{r}_{q,t}(g)$ can be simplified accordingly.

Block despreading operation is then performed by first constructing a $M \times G$ matrix, $\mathbf{R}_{q,t}$, as:

$$\begin{aligned} \mathbf{R}_{q,t} &= [\mathbf{r}_{q,t}(1) \ \cdots \ \mathbf{r}_{q,t}(G)] \\ &= \begin{cases} \sum_{k=1}^K (\mathbf{H}_{k,q1,1} \mathbf{d}_{k,1} + \mathbf{H}_{k,q2,1} \mathbf{d}_{k,2}) \mathbf{c}_k^T + \mathbf{N}_{q,1} & \text{for } t = 1 \\ \sum_{k=1}^K \left(-\mathbf{H}_{k,q1,2} \mathbf{d}_{k,2}^{<m>} + \mathbf{H}_{k,q2,2} \mathbf{d}_{k,1}^{<m>} \right) \mathbf{c}_k^T + \mathbf{N}_{q,2} & \text{for } t = 2 \end{cases} \end{aligned} \quad (5.8)$$

where $\mathbf{N}_{q,t} = [\mathbf{n}_{q,t}(1) \ \dots \ \mathbf{n}_{q,t}(G)]$. Now, let k_0 denote the desired user. The block despreading is then carried out by multiplying $\mathbf{R}_{q,t}$ with $\mathbf{c}_{k_0}$ as follows:

$$\begin{aligned} \mathbf{y}_{q,1|k_0} &= \mathbf{R}_{q,1} \mathbf{c}_{k_0} \\ &= \sum_{k=1}^K \mathbf{H}_{k_0,q1,1} \mathbf{d}_{k,1}(\mathbf{c}_k \mathbf{c}_{k_0}) + \mathbf{H}_{k_0,q2,1} \mathbf{d}_{k,2}(\mathbf{c}_k \mathbf{c}_{k_0}) + \sum_{g=1}^G c_{k_0,g} \mathbf{v}_{q,1}(g) \\ &= \mathbf{H}_{k_0,q1,1} \mathbf{d}_{k_0,1} + \mathbf{H}_{k_0,q2,1} \mathbf{d}_{k_0,2} + \mathbf{v}_{q,1|k_0} \end{aligned} \quad (5.9)$$

$$\begin{aligned} \mathbf{y}_{q,2|k_0} &= \mathbf{R}_{q,2} \mathbf{c}_{k_0} \\ &= \sum_{k=1}^K \left(-\mathbf{H}_{k_0,q1,2} \mathbf{d}_{k,2}^{<m>}(\mathbf{c}_k \mathbf{c}_{k_0}) + \mathbf{H}_{k_0,q2,2} \mathbf{d}_{k,1}^{<m>}(\mathbf{c}_k \mathbf{c}_{k_0}) \right) + \sum_{g=1}^G c_{k_0,g} \mathbf{v}_{q,2}(g) \\ &= -\mathbf{H}_{k_0,q1,2} \mathbf{d}_{k_0,2}^{<m>} + \mathbf{H}_{k_0,q2,2} \mathbf{d}_{k_0,1}^{<m>} + \mathbf{v}_{q,2|k_0} \end{aligned} \quad (5.10)$$

where $\mathbf{v}_{q,t|k_0} = \sum_{g=1}^G c_{k_0,g} \mathbf{n}_{q,t}(g)$. The results in Equation (5.9) and (5.10) are obtained based on the assumption that perfectly orthogonal codes (such as Walsh Hadamard code) are used and hence the following property is true:

$$\mathbf{c}_k^T \mathbf{c}_{k_0} = \begin{cases} 1 & \text{for } k = k_0 \\ 0 & \text{otherwise} \end{cases} \quad (5.11)$$

Now, as shown in Figure 5.3, FFT operation needs to be performed on both received signals and hence, the two FFT blocks are depicted in the Figure. However, since the two signals do not arrive simultaneously, only a single FFT module is practically required for single antenna implementation at the receiver. Applying FFT operation on the two received signals would then result in:

$$\begin{aligned} \bar{\mathbf{y}}_{q,1|k_0} &= [\bar{y}_{q,1|k_0}(0) \ \dots \ \bar{y}_{q,1|k_0}(M-1)]^T \\ &= \mathbf{F}_M \mathbf{H}_{k_0,q1,1} \mathbf{F}_M^H \mathbf{F}_M \mathbf{d}_{k_0,1} + \mathbf{F}_M \mathbf{H}_{k_0,q2,1} \mathbf{F}_M^H \mathbf{F}_M \mathbf{d}_{k_0,2} + \mathbf{F}_M \mathbf{v}_{q,1|k_0} \\ &= \bar{\mathbf{H}}_{k_0,q1,1} \bar{\mathbf{d}}_{k_0,1} + \bar{\mathbf{H}}_{k_0,q2,1} \bar{\mathbf{d}}_{k_0,2} + \bar{\mathbf{v}}_{q,1|k_0} \end{aligned} \quad (5.12)$$

$$\begin{aligned} \bar{\mathbf{y}}_{q,2|k_0} &= [\bar{y}_{q,2|k_0}(0) \ \dots \ \bar{y}_{q,2|k_0}(M-1)]^T \\ &= \mathbf{F}_M \mathbf{H}_{k_0,q2,2} \mathbf{F}_M^H \mathbf{F}_M \mathbf{d}_{k_0,1}^{<m>} - \mathbf{F}_M \mathbf{H}_{k_0,q1,2} \mathbf{F}_M^H \mathbf{F}_M \mathbf{d}_{k_0,2}^{<m>} + \mathbf{F}_M \mathbf{v}_{q,2|k_0} \\ &= \bar{\mathbf{H}}_{k_0,q2,2} \mathbf{d}_{k_0,1}^* - \bar{\mathbf{H}}_{k_0,q1,2} \mathbf{d}_{k_0,2}^* + \bar{\mathbf{v}}_{q,2|k_0} \end{aligned} \quad (5.13)$$

where $\bar{\mathbf{H}}_{k,qp,t} = \text{diag}(\bar{h}_{k,qp,t}(0) \cdots \bar{h}_{k,qp,t}(M-1))$ is a $M \times M$ matrix with the l^{th} diagonal entry as the l^{th} tone of $\mathbf{h}_{k,qp,t}$, $\bar{\mathbf{d}}_{k,p} = \mathbf{F}_M \mathbf{d}_{k,p} = [\bar{d}_{k,p}(0) \cdots \bar{d}_{k,p}(M-1)]^T$ and $\bar{\mathbf{v}}_{q,t|k} = \mathbf{F}_M \mathbf{v}_{q,t|k} [\bar{v}_{q,t|k}(0) \cdots \bar{v}_{q,t|k}(M-1)]^T$ are the frequency-domain representation of $\mathbf{d}_{k,p}$ and $\mathbf{v}_{q,t|k}$ respectively. In Equation (5.13), the multiplication result $\mathbf{F}_M \mathbf{d}_{k,p}^{<m>} = \mathbf{d}_{k,p}^*$ makes use of the symmetry property of FFT which states that [143]:

$$\underbrace{\mathbf{x}^* \langle -\mathbf{n} \rangle_{\mathbf{N}}}_{\text{time-reversal conjugate}} \xrightarrow{FFT} \underbrace{\mathbf{X}^* [k]}_{\text{simple conjugate}} \quad (5.14)$$

From Equation (5.12) and (5.13), it can be deduced that the encoded symbols that were transmitted in time-domain, possess the property of an Alamouti code, but in frequency-domain instead. This property allows simple decoding process to be carried out on individual tones. The l^{th} tones from all N_R antennae are arranged as:

$$\begin{aligned} \bar{\mathbf{y}}_{k_0}(l) &= [\bar{y}_{1,1|k_0}(l) \cdots \bar{y}_{N_R,1|k_0}(l) \bar{y}_{1,2|k_0}(l) \cdots \bar{y}_{N_R,2|k_0}(l)] \\ &= \begin{bmatrix} \bar{h}_{k_0,11,1}(l) & \bar{h}_{k_0,12,1}(l) \\ \vdots & \vdots \\ \bar{h}_{k_0,N_R,1}(l) & \bar{h}_{k_0,N_R,2,1}(l) \\ \bar{h}_{k_0,12,2}^*(l) & -\bar{h}_{k_0,11,2}^*(l) \\ \vdots & \vdots \\ \bar{h}_{k_0,N_R,2,2}^*(l) & -\bar{h}_{k_0,N_R,1,2}^*(l) \end{bmatrix} \begin{bmatrix} \bar{d}_{k_0,1}(l) \\ \bar{d}_{k_0,2}(l) \end{bmatrix} + \begin{bmatrix} \bar{v}_{1,1|k_0}(l) \\ \vdots \\ \bar{v}_{N_R,1|k_0}(l) \\ \bar{v}_{1,2|k_0}(l) \\ \vdots \\ \bar{v}_{N_R,2|k_0}(l) \end{bmatrix} \\ &= \bar{\mathbf{H}}_{k_0}(l) \mathbf{d}_{k_0}(l) + \bar{\mathbf{w}}_{k_0}(l) \end{aligned} \quad (5.15)$$

Equalization can then be performed on each individual sub-carrier using such techniques as Orthogonal Restore Combining (ORC) or Minimum Mean Square Error (MMSE). The ORC and MMSE equalizer matrix is given respectively as:

$$\bar{\mathbf{G}}_{k_0,ORC}(l) = \begin{bmatrix} \alpha_{k_0,1}(l) & 0 \\ 0 & \alpha_{k_0,2}(l) \end{bmatrix} \begin{bmatrix} \bar{h}_{k_0,11,1}^*(l) & \cdots & \bar{h}_{k_0,N_R,1,1}^*(l) & \bar{h}_{k_0,12,2}(l) & \cdots & \bar{h}_{k_0,N_R,2,2}(l) \\ \bar{h}_{k_0,12,1}(l)^* & \cdots & \bar{h}_{k_0,N_R,2,1}^*(l) & \bar{h}_{k_0,11,2}(l) & \cdots & \bar{h}_{k_0,N_R,1,2}(l) \end{bmatrix} \quad (5.16)$$

$$\bar{\mathbf{G}}_{k_0,MMSE}(l) = (\bar{\mathbf{H}}_{k_0}^H(l) \bar{\mathbf{H}}_{k_0}(l) + \sigma_n^2 \mathbf{I}_{2 \times 2})^{-1} \bar{\mathbf{H}}_{k_0}^H(l) \quad (5.17)$$

where $\alpha_{k_0,1}(l)$ and $\alpha_{k_0,2}(l)$ is defined respectively as:

$$\begin{aligned}\alpha_{k_0,1}(l) &= \frac{1}{\sum_{q=1}^{N_R} |\bar{h}_{k_0,q1,1}(l)|^2 + \sum_{q=1}^{N_R} |\bar{h}_{k_0,q2,2}(l)|^2} \\ \alpha_{k_0,2}(l) &= \frac{1}{\sum_{q=1}^{N_R} |\bar{h}_{k_0,q2,1}(l)|^2 + \sum_{q=1}^{N_R} |\bar{h}_{k_0,q1,2}(l)|^2}\end{aligned}\tag{5.18}$$

From Equation (5.16), it can be seen that the performance of ORC technique largely depends on how close the value of $\bar{h}_{k_0,qp,1}(l)$ is to that of $\bar{h}_{k_0,qp,2}(l)$. When $\bar{h}_{k_0,qp,1}(l) \approx \bar{h}_{k_0,qp,2}(l)$, the channel matrix is an orthogonal matrix and hence the ORC technique can be applied effectively to separate the information signals transmitted from antenna 1 and 2. On the other hand, the channel matrix is no longer orthogonal during fast fading environment where $\bar{h}_{k_0,qp,1}(l) \neq \bar{h}_{k_0,qp,2}(l)$. As a result, the effectiveness of ORC technique is expected to be poor under such condition. Meanwhile, MMSE technique does not rely on the channel stationarity condition and can be considered as an alternative to ORC technique. After equalization, the equalized tones are transformed back into time-domain through IFFT operation. Symbol detection will then follow subsequently.

5.3.3 Transceiver Model with Arbitrary Antenna Diversity ($N_T \geq 2$)

SC CDMA system employing conventional spreading approach such as in [54] requires chip-wise processing. For such a system to exploit transmit diversity ($N_T \geq 2$), ST encoding would have to be performed at chip-level or after symbols are spread into chips. The proposed system, however, employing block spreading, allows the design in [144] to be easily adopted to the proposed STC BS-CDMA since encoding is performed on symbols instead of chips. At the end of the section, it is shown that the earlier STC BS-CDMA transmitter with is a special case of the STC encoder that is described in this section. To begin with, generalized real orthogonal design (GROD), $\mathcal{G}_r$ and generalized complex orthogonal design (GCOD), $\mathcal{G}_c$ are given in [144] as:

Definition 1 : Define $\mathbf{x} = [x_1, \dots, x_{N_S}]^T$ and let $\mathcal{G}_r(\mathbf{x})$ be an $N_D \times N_T$ matrix with entries $0, \pm x_1, \dots, \pm x_{N_S}$. If $\mathcal{G}_r^T(\mathbf{x})\mathcal{G}_r(\mathbf{x}) = \alpha(x_1^2 + \dots + x_{N_S}^2)\mathbf{I}_{N_T}$ with α positive, then $\mathcal{G}_r(\mathbf{x})$ is termed a GROD in variables $x_1, \dots, x_{N_S}$ of size $N_D \times N_T$ and rate $R = N_S/N_D$.

Definition 2 : Define $\mathbf{x} = [x_1, \dots, x_{N_S}]^T$ and let $\mathcal{G}_c(\mathbf{x})$ be an $N_D \times N_T$ matrix with entries $0, \pm x_1, \pm x_1^*, \dots, \pm x_{N_S}, \pm x_{N_S}^*$. If $\mathcal{G}_c^H(\mathbf{x})\mathcal{G}_c(\mathbf{x}) = \alpha(|x_1|^2 + \dots + |x_{N_S}|^2)\mathbf{I}_{N_T}$ with α positive, then $\mathcal{G}_c(\mathbf{x})$ is termed as GCOD in variables $x_1, \dots, x_{N_S}$ of size $N_D \times N_T$ and rate $R = N_S/N_D$.

When $N_T = 2$, a complex orthogonal matrix with rate $R = 1$ ($N_S = N_D = N_T$) can be easily constructed by applying simple manipulation to $\mathcal{G}_r(\mathbf{x})$. However, such a matrix does not exist for $N_T > 2$. General GCOD with rate $R = \frac{1}{2}$ for $N_T > 2$ can be constructed as follows:

1. Construct GROD $\mathcal{G}_r(\mathbf{x})$ of size $N_S \times N_T$ with $R=1$
2. Replace the symbols $x_1, \dots, x_{N_S}$ by their conjugates $x_1^*, \dots, x_{N_S}^*$ to arrive at $\mathcal{G}_r(\mathbf{x}^*)$
3. Form $\mathcal{G}_c(\mathbf{x}) = [\mathcal{G}_r^T(\mathbf{x}), \mathcal{G}_r^T(\mathbf{x}^*)]^T$.

The ST encoder takes in $N_S M \times 1$ symbols $\mathbf{d}_k = [d_k(1) \dots d_k(N_S M - 1)]^T$ and divides them into N_S blocks of size $M \times 1$ each, with the p^{th} block represented as $\mathbf{d}_{k,p} = [d_k((p-1)M + 1) \dots d_k(pM)]^T, p \in [1, N_S]$. The output of the ST encoder is:

$$\bar{\mathbf{D}} = \begin{bmatrix} \bar{\mathbf{d}}_{k,1}(1) & \dots & \bar{\mathbf{d}}_{k,1}(N_D) \\ \vdots & \ddots & \vdots \\ \bar{\mathbf{d}}_{k,N_T}(1) & \dots & \bar{\mathbf{d}}_{k,N_T}(N_D) \end{bmatrix} \quad (5.19)$$

where $\bar{\mathbf{D}}$ has been constructed by replacing $x_1, \dots, x_{N_S}$ in $\mathcal{G}_c(\mathbf{x})$ that was constructed using step one to three of the above algorithm by $\frac{1}{\sqrt{2}}\mathbf{d}_{k,1}, \dots, \frac{1}{\sqrt{2}}\mathbf{d}_{k,N_S}$ and $x_1^*, \dots, x_{N_S}^*$ by $\mathbf{P}_M^1 \mathbf{d}_{k,1}^*, \dots, \mathbf{P}_M^1 \mathbf{d}_{k,N_S}^*$. Here, $\mathbf{P}_M^1$ is a $M \times M$ permutation matrix [144] which effectively performs time-reversal operation on $\mathbf{d}_{k,p}^*$. For discussion, assume the case when $N_T = 2$. The encoder output is simply:

$$\bar{\mathbf{D}} = \begin{bmatrix} \bar{\mathbf{d}}_{k,1}(1) & \bar{\mathbf{d}}_{k,1}(2) \\ \bar{\mathbf{d}}_{k,2}(1) & \bar{\mathbf{d}}_{k,2}(2) \end{bmatrix} = \begin{bmatrix} \mathbf{d}_{k,1} & -\mathbf{P}_M^1 \mathbf{d}_{k,2}^* \\ \mathbf{d}_{k,2} & \mathbf{P}_M^1 \mathbf{d}_{k,1}^* \end{bmatrix} \quad (5.20)$$

where $\bar{\mathbf{d}}_{k,p}(i)$ refers to the block of symbols to be transmitted from the p^{th} antenna during i^{th} interval. Block spreading on each $M \times 1$ block results in $G \times M$ spread matrix, $\mathbf{X}_{k,p}(i) = \mathbf{c}_k \bar{\mathbf{d}}_{k,p}^T(i)$. Chips from the spread matrix are then read row-wise and G rows are transmitted one-by-one after CP insertion. At the receiver, CP is first removed and received chips are

then stacked row-by-row as:

$$\mathbf{R}(i) = \sum_{k=1}^K \left[\sum_{p=1}^2 \mathbf{h}_{k,1p}(i, 1) \otimes \mathbf{x}_{k,p}(i, 1) \quad \cdots \quad \sum_{p=1}^2 \mathbf{h}_{k,1p}(i, G) \otimes \mathbf{x}_{k,p}(i, G) \right]^T + \mathbf{n}(i) \quad (5.21)$$

where $\mathbf{x}_{k,p}(i, g) = c_{k,g} \bar{\mathbf{d}}_{k,p}(i)$. Performing block despreading operation (for user 1) results in $\mathbf{y}(i) = \sum_{p=1}^2 \mathbf{H}_{1,1p}(i) \bar{\mathbf{d}}_{1,p}(i) + \nu(i)$. Performing FFT during the $(2i)^{th}$ and $(2i+1)^{th}$ block period results in:

$$\begin{aligned} \bar{\mathbf{y}}(2i) &= \mathbf{F}_M \mathbf{y}(2i) = \bar{\mathbf{H}}_{1,11}(2i) \check{\mathbf{d}}_{k,1}(2i) + \bar{\mathbf{H}}_{1,12}(2i) \check{\mathbf{d}}_{k,2}(2i) + \bar{\nu}_1(2i) \\ \bar{\mathbf{y}}(2i+1) &= \mathbf{F}_M \mathbf{y}(2i+1) = -\bar{\mathbf{H}}_{1,11}(2i+1) \check{\mathbf{d}}_{k,2}^*(2i) + \bar{\mathbf{H}}_{1,12}(2i+1) \check{\mathbf{d}}_{k,1}^*(2i) + \bar{\nu}_1(2i+1) \end{aligned} \quad (5.22)$$

where the results in Equation (5.22) are obtained by using the time-reversal property of DFT which states that a conjugate time-reversal operation corresponds to the conjugate of individual frequency tones, i.e. $\mathbf{F}_M \mathbf{P}_M^1 \mathbf{d}_{k,p}^*(2i) \equiv \check{\mathbf{d}}_{k,p}^*(2i)$. Similarly, Orthogonal Restore Combining (ORC) [141] or MMSE techniques can then be performed on individual tones, followed by IFFT operation to recover the time-domain symbols for detection.

5.4 Numerical Results

This section presents the results from simulation study of STC BS-CDMA under different practically realistic deployment scenarios. After that, the performance of STC BS-CDMA is compared with that of LST BS-CDMA system.

5.4.1 BER Performance of STC BS-CDMA

The STC BS-CDMA architecture was modeled and its performance was investigated using monte-carlo simulation for both uplink and downlink deployment. The simulation studies target high data rate and high-mobility communication systems that delivers system performance improvement compared to the basic single antenna BS-CDMA system. Unlike LST architecture, STC architecture aims to enhance the system performance by improving the bit error rate. With the significant improvement expected, STC system allows the use of higher modulation schemes. For STC BS-CDMA simulation, 16-QAM and 64-QAM mod-

ulation is considered in addition to the basic BPSK and QPSK. The transmission carrier frequency is assumed to be 5GHz, with a chip-rate of 40Mcps. The transmitter diversity is chosen to be two ($N_T = 2$). Unlike LST processing, STC system has the flexibility in choosing the receiver diversity order. In this investigation, receiver diversity of 1, 2 and 3 are considered. Although it is possible to increase the receiver diversity further (especially for uplink), there is a limit for the performance gain that can be achieved and hence increasing the receiver diversity beyond certain value will not yield significant improvement. Two different block sizes are being considered, namely $M = 256$ and $M = 512$. The length of CP employed is $L_{CP} = \frac{1}{4}M$. This effectively means that a single chip-block is of size 320 and 640 respectively. For $M = 256$, the duration of each chip-block, T , is $8\mu\text{secs}$, where $6.4\mu\text{secs}$ is the duration of information chips and $1.6\mu\text{secs}$ is the duration of guard interval. At the receiver, the system bandwidth of 40MHz is divided into M subchannels and equalized individually. The processing gain used are 8, 16 and 32. For the processing gain of 32, each user is transmitting at bandwidth efficiency of 0.125bps/Hz and 0.1875bps/Hz when 16-QAM and 64-QAM modulation is used respectively. The number of actively transmitting users is varied according to different loading conditions in the system. The channel model assumes *Rayleigh* distribution that is generated based on Jakes' fading model [132] and the number of path is assumed to be 24. It is assumed that the maximum delay spread is less than the guard duration and hence no IBI occurs. Different vehicular mobility conditions are also used to simulate the practical deployment of typical mobile communication system, namely 0Hz (stationary), 463Hz (medium-high) and 926Hz (high). The complete set of simulation parameters is tabulated in Table 5.1.

In Figure 5.4, the performance of a 2×2 STC BS-CDMA system using QPSK modulation is shown. In Figure 5.4(a), the BER performance obtained using ORC equalizer is depicted. From the Figure, it can be seen that the BER performance is virtually free from any MUI and remains identical even under different loading conditions when the channel is stationary ($F_D = 0\text{Hz}$). However, an interesting observation from the results indicates that the single-user performance is affected by the presence of Doppler spread in the system. For example, the single-user performance at a BER of 1% is about 1dB worse when $F_D = 463\text{Hz}$ compared to when $F_D = 0\text{Hz}$. The performance degradation observed is about 3dB when $F_D = 926\text{Hz}$. This is attributed to the fact that under high mobility environment, the rapid channel variation now means that deeply fading channels would be encountered more frequently during a given period of time. The rapid frequency selective fading encountered

Table 5.1: STC BS-CDMA Systems: Simulation Parameters

	System I	System II
Carrier Frequency, F	5 GHz	5GHz
Chip rate, F_C	40.96 Mcps	40.96 Mcps
Number of Tx Antenna, N_T	2	4
Number of Rx Antenna, N_R	1 2	2 3
Modulation	BPSK QPSK	BPSK QPSK 16-QAM 64-QAM
OFDM / Chip-Block Size, M	256 (512)	256 (512)
FFT / IFFT Size, M	256 (512)	256 (512)
OFDM Block / Chip-Block Duration, T	8 μ secs (16 μ secs)	8 μ secs (16 μ secs)
Data Duration per OFDM Block, T_D	6.4 μ secs (12.8 μ secs)	6.4 μ secs (12.8 μ secs)
Guard Duration per OFDM Block, T_{CP}	1.6 μ secs (3.2 μ secs)	1.6 μ secs (3.2 μ secs)
Super-Block Size, M_{SB}	PG OFDM Block	PG OFDM Blocks
Super-Block Duration, T_{SB}	PG $\times$ 8 μ secs (16 μ secs)	PG $\times$ 8 μ secs (PG $\times$ 16 μ secs)
Processing Gain, PG	8 16 32	8 16 32
Symbols per Super-Block, N_{SB}	256 (512)	256 (512)
Load Condition	1 user Half-Load Full-Load	1 user Half-Load Full-Load
Doppler Spread, F_D	0 Hz 463 Hz 926 Hz	0 Hz 463 Hz 926 Hz
Vehicular Mobility, V	0 kph 100 kph 200 kph	0 kph 100 kph 200 kph
Channel Coherence Time, T_{CO}	∞ 387 μ secs 193 μ secs	∞ 387 μ secs 193 μ secs

during high mobility environment degrades the system performance. Another observation is that the impact of loading condition on the system performance depends on the amount of Doppler spread that is present in the system. Under moderately high mobility ($F_D = 463\text{Hz}$), performance degradation of less than 0.5dB is observed with half-load condition at BER of 1%. The performance degradation with full-load condition is less than 2dB. On the other hand, the performance degradation is more severe under very high mobility ($F_D = 926\text{Hz}$). At half-load condition and BER of 1%, the BER performance deviates by about 4dB from the single-user performance. Under full-load condition, the system performance exhibits an error-floor at around 8%.

The BER performance obtained using MMSE equalizer is plotted in Figure 5.4(b). From the Figure, it can be inferred that the performance of MMSE equalizer is superior than ORC. To compare the system performance of the two techniques, consider the case of a single-user with $F_D = 0\text{Hz}$. With ORC equalizer, BER of 1% is achieved at SNR value of around 9dB whereas an MMSE equalizer achieves the same BER at SNR value of 5dB. Another observation from the Figure is the smaller impact of load and Doppler conditions on the performance of MMSE equalizer. At BER of 10% and moderately high mobility ($F_D = 463\text{Hz}$), the half-load performance is close to the single-user performance while the full-load performance is only about 0.5dB worse than the single-user performance. Meanwhile, under very high mobility ($F_D = 926\text{Hz}$), the half-load performance shows about 0.5dB degradation from the single-user performance and the degradation grows to about 3.5dB at full-load condition. However, the full-load system performance exhibits an error-floor that is less than 0.1% when $F_D = 926\text{Hz}$.

With the significant performance improvement attained in STC BS-CDMA, higher modulation scheme, namely 16-QAM and 64-QAM had been studied and their performance is plotted in Figure 5.5 and 5.6 respectively. For 16-QAM modulation, the number of antenna employed at the receiver is 2. As shown in Figure 5.5, the BER performance for single-user scenario degrades by less than 1dB under severely high mobility condition (for BER of 1%). Under moderate mobility conditions, the system performance is sufficiently robust against the impact due to MUI. The MUI performance degradation at half-load and full-load condition is observed to be less than 0.5 and 3dB respectively at a BER of 1%. Meanwhile, considerable degradation due to MUI is observed during high mobility condition. At full system load, an error floor of about 8% is observed.

(a) Performance of ORC Equalizer

(b) Performance of MMSE Equalizer

Figure 5.4: BER Performance of STC BS-CDMA System (2×2 , QPSK)
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Figure 5.5: BER Performance of STC BS-CDMA System (2×2 , 16-QAM)

Figure 5.6 depicts the BER performance of a STC BS-CDMA system with 64-QAM. Since the higher modulation scheme chosen is expected to have significant impact on the error rate performance, the diversity order at the receiver is selected as 3 in order to make up for the performance loss. The single-user performance loss due to mobility condition is 0.2dB and $< 1\text{dB}$ when the Doppler spread is 463Hz and 926Hz respectively. Under moderately high mobility, performance loss at half-load is found to be approximately 1dB at BER of 1%. Meanwhile, an error floor of close to 1% is observed when the system is fully loaded. During severely high mobility, the system performance suffers considerably and an error floor of about 10 – 11% is observed when the system load is full.

Compared to the SISO case, derivation of theoretical SNR for MIMO case is not that straightforward. In SISO case, a much simplified SNR expression for each symbol can eventually be obtained. However, the per-tone equalization techniques in MIMO architectures cause symbols from the different antennas to mix together, causing inter-antenna interference. This makes obtaining SNR expression for each recovered symbol a much more challenging task.

Figure 5.6: BER Performance of STC BS-CDMA System (2×3 , 64-QAM)

5.4.2 LST BS-CDMA vs STC BS-CDMA: Performance Comparison

Now that STC BS-CDMA has been proposed as an alternative MIMO architecture to enhance the performance of BS-CDMA system, the error-rate performance of the two systems will be compared next. For this comparison, both systems are configured to transmit with the same chip rate. This is achieved by the use of QPSK modulation in 3×3 LST BS-CDMA and 64-QAM modulation for 2×2 STC BS-CDMA system. The comparison is made based on the BER performance obtained by using MMSE equalizers. The comparison is presented in Figure 5.7 with the specific results under different loading condition plotted in Figure 5.7(a), 5.7(b) and 5.7(c) respectively.

Figure 5.7(a) shows the performance comparison for $K = 1$. From Figure 5.7(a), it can be seen that the performance of STC BS-CDMA system generally outperforms LST BS-CDMA under all three Doppler spread conditions when there is only one user. However, an interesting observation from Figure 5.7(a) points out to the fact that the performance of STC BS-CDMA degrades much more severely compared to that of LST BS-CDMA. For $F_D = 0\text{Hz}$, the SNR required by STC BS-CDMA to achieve $\text{BER} = 10^{-2}$ is 0.9dB lower

Figure 5.7: LST BS-CDMA vs STC BS-CDMA - A Performance Comparison

than that required in LST BS-CDMA. For $F_D = 926\text{Hz}$, such superiority drops to 0.3dB.

Figure 5.7(b) shows the performance comparison for $K = 16$ or half-load condition. From Figure 5.7(b), STC BS-CDMA is seen to outperform LST BS-CDMA by the same margin for $F_D = 0\text{Hz}$. This is expected as both systems are free from MUI and hence both systems do not suffer any degradation due to the increase in the number of users. At $F_D = 463\text{Hz}$, the performance gap reduces to about 0.6dB while at $F_D = 926\text{Hz}$ both systems encountered error-floor that are higher than 10^{-2} . Nonetheless, the performance of STC BS-CDMA is still noticeably better than LST BS-CDMA for $F_D = 926\text{Hz}$.

Figure 5.7(c) shows the performance comparison for $K = 32$ or full-load condition. From Figure 5.7(c), both systems perform almost identical at $F_D = 463$ and 926Hz . Based on all the observation, it can be concluded that STC BS-CDMA consistently performs better than LST BS-CDMA system during low mobility condition. During high mobility condition, the performance advantage diminishes, in particular under full-loading condition. Other than the error-rate performance, it is interesting to compare the potential hardware complexity that will be associated with both systems. For LST BS-CDMA, 3 antennae are required at both the transmitter and receiver units, while only 2 are required for STC BS-CDMA. On the other hand, STC BS-CDMA uses 64-QAM, a much higher modulation compared to LST BS-CDMA. This usually results in reduction of the power efficiency in the amplifier. Based on this analysis, it is important to consider factors such as the size constraint of the mobile device and the cost of amplifier with highly linear operating range before deciding which of the two schemes provides the most suitable option for a particular system needs.

5.5 Summary

In this Chapter, a BS-CDMA system with MIMO antenna capability is presented. The multiple antennae at the transmitter are used to transmit space-time coded symbols and hence referred to as STC BS-CDMA. With minimal complexity, the use of such code delivers significant performance gain. The error rate improvement with 2 receiving antennae allows the use of 16-QAM modulation, while the improvement with 3 receiving antennae allows the use of 64-QAM modulation. In comparison to when QPSK modulation is used, 16-QAM and 64-QAM offers twice and four times effective symbol rate enhancement respectively.

Based on the extensive performance investigation that has been carried out so far, it can be concluded that the improved error-rate performance of STC BS-CDMA system allows higher modulation scheme to be used that will in turn enhance the transmission rate capability. One interesting issue that has not been addressed is given a desired data rate, which combination of STC code rate and modulation scheme would be the best option. This issue would be addressed as part of the future work of this thesis.

The contribution in this chapter can be summarized by the following points:

- a space-time block coded architecture for BS-CDMA system is presented and referred to as STC BS-CDMA;
- construction of orthogonal block codes with arbitrary order of transmit diversity is presented;
- performance of STC BS-CDMA system with different antenna configuration and mobility pattern obtained using extensive simulation study is presented; and
- the performance gain in STC BS-CDMA is substituted with the use of higher modulation scheme in order to deliver higher transmission rate.

Chapter 6

Space-Frequency Coded Block Spread CDMA (SFC BS-CDMA) Systems

6.1 Introduction

In Chapter 5, space-time coded MIMO architecture for BS-CDMA system was presented. While STC BS-CDMA offers significant BER performance, it faces two challenges. Firstly, the computational delay in processing a block of symbols is relatively high. Secondly, the long duration required to transmit an encoded block means that channel coherence time would need to be equally long. This makes the performance of STC BS-CDMA to be inferior among high mobility users and less suitable for applications where processing delay is critical. Space-Frequency Coded BS-CDMA (SFC BS-CDMA) attempts to mitigate the impact caused by the two issues and has been proposed as an alternative to STC BS-CDMA. Details of SFC BS-CDMA is presented in this Chapter.

Chapter 6 is organized as follows. Section 6.2 discusses the challenges encountered in STC BS-CDMA and how it is being addressed in SFC BS-CDMA system. Section 6.3 presents the system model of SFC BS-CDMA. Section 6.5 provides details of simulation studies and discusses the results of these studies.

6.2 Motivation

It is well-established that the formation of MIMO system with the use of multiple antennae at both the transmitter and receiver, enhances overall system capacity considerably. However, designing a transceiver architecture that is capable of realizing the promised potential remains a challenging research issue.

On one hand, layered space-time (LST) system attempts to deliver direct rate enhancement by exploiting each additional antenna to transmit additional information. However, the error rate performance of such system deteriorates as the number of antenna increases. Such degradation can be mitigated by the use of iterative detector with appropriate channel rank ordering [79]. However, such algorithms incur high processing complexity at the receiver. Without such advanced algorithms, the error rate of LST system sets a limit to the highest modulation scheme that can be used. In addition, the minimum number of antennae required at the receiver needs to be equal to that of the transmitter. In downlink transmission, the number of antenna employed at the BS transmitter is constrained by the limitation of the MS.

On the other hand, space-time coded (STC) system attempts to deliver remarkable error rate improvement which allows the use of higher modulation schemes. Unlike LST, STC system does not have the minimum antenna requirement. This means that regardless of the transmit diversity order, a receiver is able to recover the encoded symbols using a single antenna. Although higher diversity order is always desirable, the effective transmission rate drops when diversity order is higher than two.

While STC BS-CDMA has been shown to offer significant performance improvement to LST BS-CDMA system, it has two shortcomings. Firstly, STC BS-CDMA processing incurs long processing delay before symbols from a particular block are available for detection. This restricts the suitability of STC BS-CDMA in delay-critical applications. Secondly, the long duration of transmitting all the chips belonging to a particular block means the channel coherence time would need to be longer due to the channel stationarity assumption. This reduces the effectiveness of STC BS-CDMA system in supporting high mobility users.

It was in this context that SFC BS-CDMA was proposed in [145]. SFC BS-CDMA attempts to reduce the overall duration of chip transmission by transmitting encoded information

using the spatial and frequency dimension instead of spatial and temporal dimension. In STC BS-CDMA, the number of symbols in each block is fixed. As a result, the Alamouti encoded symbols with two antenna diversity would now take twice longer to transmit. The duration grows proportionally when higher diversity is employed. In SFC BS-CDMA system, the number of symbols in each block is made flexible, depending on the antenna diversity used. For example, the number of symbols in one block now reduces to $\frac{M}{2}$ when two transmit antennae are used. By allowing the number of symbols to vary, the total duration required to transmit the encoded information belonging to a single block is now fixed, regardless of the diversity order. This would address the major issues plaguing STC BS-CDMA systems.

6.3 System Model

This section presents the transmitter architecture of SFC BS-CDMA with transmit diversity order of two. The signal processing at the receiver is presented, followed by generalization of the space-frequency coding for arbitrary order of transmitter diversity.

6.3.1 Transmitter Model

Figure 6.1: A Generalized Baseband SFC BS-CDMA Transmitter Architecture

Figure 6.1 shows the general transceiver structure of SFC BS-CDMA, where only the k^{th} user is shown out of all K users. This work focuses on the uplink communication and the number of transmit antennas is assumed to be 2 ($N_T = 2$). The information bits

$d_k(n)$ belonging to the k^{th} user are first grouped into blocks of size $M \times 1$ with $\mathbf{d}_k(i) = [d_k(iM) \cdots d_k(iM + M - 1)]^T$, where n is related to i by $n = iM + m, m \in [0, M - 1]$. Without loss of generality, i will be omitted in subsequent discussions. Each block is then fed into SFBC encoder. The encoder first divides each block of M symbols into 2 smaller sub-blocks, $\mathbf{d}_{k,p} = [d_k((p-1)\frac{M}{2} + 1) \cdots d_k(p\frac{M}{2})]^T$, each of size $\frac{M}{2}$ and p is used to index these blocks. The space-frequency encoding procedure is given as follows:

1. Repeat or append $\mathbf{d}_{k,p}$ to itself to produce $\dot{\mathbf{d}}_{k,p}$ of size $M \times 1$

$$\dot{\mathbf{d}}_{k,p} = \left((\mathbf{d}_{k,p})^T (\mathbf{d}_{k,p})^T \right)^T$$
2. Perform cyclic conjugate on $\dot{\mathbf{d}}_{k,p}$ to produce $\ddot{\mathbf{d}}_{k,p}$

$$\ddot{\mathbf{d}}_{k,p} = \mathbf{P}_M^1 \dot{\mathbf{d}}_{k,p}$$

where $\mathbf{P}_M^1$ is the $M \times M$ permutation matrix with $(M+1)$ st entry of $[\mathbf{P}_M^1 \mathbf{a}]_m$ as

$$[\mathbf{P}_M^1 \mathbf{a}]_m = a((M-m) \bmod M)$$
3. Phase-rotate $\ddot{\mathbf{d}}_{k,p}$ by multiplying each bit with $e^{\frac{j2\pi m}{M}}$

$$[\ddot{\mathbf{d}}_{k,p}]_m = [\ddot{\mathbf{d}}_{k,p}]_m e^{\frac{j2\pi m}{M}} \quad \text{for } m = 0 \text{ to } M-1$$
4. The encoded streams to be transmitted from antenna 1 and 2 are given respectively as:

$$\mathbf{x}_{k,1} = \dot{\mathbf{d}}_{k,1} - \ddot{\mathbf{d}}_{k,2} \quad \text{and} \quad \mathbf{x}_{k,2} = \dot{\mathbf{d}}_{k,2} + \ddot{\mathbf{d}}_{k,1}$$

In step 1, $\dot{\mathbf{d}}_{k,p}$ is a repeated finite-length sequence with a special frequency domain property. In frequency domain, non-null tones of $\dot{\mathbf{d}}_{k,p}$ only exist at odd indices ($l \in \Omega_1 = [1, 3, \dots]$). In step 2, cyclic conjugate operation is performed on $\dot{\mathbf{d}}_{k,p}$ which mathematically can be represented by pre-multiplication with permutation matrix $\mathbf{P}_M^1$. In step 3, the cyclic conjugated repeated finite-length sequence $\ddot{\mathbf{d}}_{k,p}$ is phase-rotated, which results in its non-null tones to exist only at even indices ($l \in \Omega_2 = [2, 4, \dots]$).

Each output stream is fed into block spreader. The mathematical representation of the block spreading operation is given as:

$$\begin{aligned} \mathbf{X}_{k,p} &= \mathbf{x}_{k,p} \otimes \mathbf{c}_k^T \\ &= [\mathbf{x}_{k,p} c_{k,1} \cdots \mathbf{x}_{k,p} c_{k,G}] \\ &= [\mathbf{x}_{k,p}(1) \cdots \mathbf{x}_{k,p}(G)] \end{aligned} \tag{6.1}$$

where $\mathbf{c}_k$ is k^{th} user's spread code and G is the processing gain. The block spreader outputs $M \times G$ matrix, where $\mathbf{x}_{k,p}(g)$ is the g^{th} column entry. After cyclic prefix (CP) insertion, $\mathbf{x}_{k,p}(g)$ is transmitted from the p^{th} antenna.

6.3.2 Receiver Model

After CP removal, the received signals during the g^{th} interval is given as:

$$\begin{aligned} \mathbf{r}_q(g) &= \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{H}_{k,qp}(g) \mathbf{x}_{k,p}(g) + \mathbf{n}_q(g) \\ &= \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{H}_{k,qp}(g) \mathbf{x}_{k,p} c_{k,g} + \mathbf{n}_q(g) \end{aligned} \quad (6.2)$$

where $\mathbf{H}_{k,qp}(g)$ is k^{th} user's circular toeplitz channel matrix during g^{th} interval with its first column as $\mathbf{h}_{k,qp}(g) = [h_{k,qp}(0) \cdots h_{k,qp}(L-1)]^T$. $h_{k,qp}(l)$ is the l^{th} path of the channel between the q^{th} receive and p^{th} transmit antenna of the k^{th} user and $\mathbf{n}_q(g)$ is Additive White Gaussian Noise (AWGN). Assuming that the total transmission duration is within the channel coherence time, the channel parameters can be assumed to be little or relatively invariant, i.e. $\mathbf{H}_{k,qp}(1) \approx \cdots \approx \mathbf{H}_{k,qp}(G) = \mathbf{H}_{k,qp}$. Before other signal processing modules, block despreading operation is carried out to convert the received chips back into symbols. The block despreader takes in the received chips and arranges them column-by-column to construct $M \times G$ chip matrix. Assuming the signal detection is carried out for user k_0 , block despreading is performed by the following multiplication:

$$\begin{aligned} \mathbf{r}_q &= [\mathbf{r}_q(1) \cdots \mathbf{r}_q(G)] \mathbf{c}_{k_0} \\ &= \sum_{k=1}^K \sum_{p=1}^{N_T} \mathbf{H}_{k,qp} \mathbf{d}_{k,p} \left(\sum_{g=1}^G c_{k,g} c_{k_0,g} \right) + \sum_{g=1}^G \mathbf{n}_q(g) c_{k_0,g} \\ &= \sum_{p=1}^{N_T} \mathbf{H}_{k_0,qp} \mathbf{d}_{k_0,p} + \mathbf{v}_{k_0,q} \end{aligned} \quad (6.3)$$

where $\mathbf{v}_{k_0,q} = \sum_{g=1}^G \mathbf{n}_q(g) c_{k_0,g}$ can be viewed as the result of performing averaging operation on the AWGN component. Note that the above result is obtained based on the assumption

of orthogonality among the spreading codes and the following is true:

$$\sum_{g=1}^G c_{k,g} c_{k_0,g} = \begin{cases} 1 & \text{for } k = k_0 \\ 0 & \text{otherwise} \end{cases} \quad (6.4)$$

After despreading, FFT operation is performed on the despread symbols which results in the following:

$$\begin{aligned} \bar{\mathbf{y}}_{q,k_0} &= \sum_{p=1}^{N_T} \mathbf{F}_M \mathbf{H}_{k_0,qp} \mathbf{F}_M^H \mathbf{F}_M \mathbf{d}_{k_0,p} + \mathbf{F}_M \mathbf{v}_{k_0,q} \\ &= \sum_{p=1}^{N_T} \bar{\mathbf{H}}_{k_0,qp} \bar{\mathbf{d}}_{k_0,p} + \bar{\mathbf{v}}_q \end{aligned} \quad (6.5)$$

where $\bar{\mathbf{H}}_{k_0,qp}$ is the channel frequency response experienced by user k_0 between its p^{th} transmit and the q^{th} receive antenna, $\bar{\mathbf{d}}_{k_0,p}$ is the frequency tones of the signals transmitted by user k_0 on the p^{th} antenna and $\bar{\mathbf{v}}_q$ is the frequency tones of the 'averaged' AWGN component. $\bar{\mathbf{H}}_{k_0,qp}$, $\bar{\mathbf{d}}_{k_0,p}$ and $\bar{\mathbf{v}}_q$ are defined as follows:

$$\begin{aligned} \bar{\mathbf{H}}_{k_0,qp} &= \mathbf{F}_M \mathbf{H}_{k_0,qp} \mathbf{F}_M^H = \begin{bmatrix} \bar{h}_{k_0,qp}(0) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \bar{h}_{k_0,qp}(M-1) \end{bmatrix} \\ \bar{\mathbf{d}}_{k_0,p} &= \mathbf{F}_M \mathbf{d}_{k_0,p} = \begin{bmatrix} \bar{d}_{k_0,p}(0) & \cdots & \bar{d}_{k_0,p}(M-1) \end{bmatrix}^T \\ \bar{\mathbf{v}}_q &= \mathbf{F}_M \mathbf{v}_{k_0,q} = \begin{bmatrix} \bar{v}_q(0) & \cdots & \bar{v}_q(M-1) \end{bmatrix}^T \end{aligned} \quad (6.6)$$

The signals on adjacent subcarriers ($l_1 = 2l - 1$) and ($l_2 = 2l$) for $l = 1$ to $\frac{M}{2}$ can be expressed as:

$$\begin{aligned} \bar{y}_{q|k_0}(l_1) &= \bar{h}_{k_0,q1}(l_1) \bar{d}_{k_0,1}(l_1) + \bar{h}_{k_0,q2}(l_1) \bar{d}_{k_0,2}(l_1) + \bar{n}_q(l_1) \\ \bar{y}_{q|k_0}(l_2) &= \bar{h}_{k_0,q2}(l_2) \bar{d}_{k_0,1}^*(l_1) - \bar{h}_{k_0,q1}(l_2) \bar{d}_{k_0,2}^*(l_1) + \bar{n}_q(l_2) \end{aligned} \quad (6.7)$$

This can be arranged as

$$\begin{aligned}
 \bar{\mathbf{y}}_{k_0}(l) &= \begin{bmatrix} \bar{y}_{1|k_0}(l_1) & \cdots & \bar{y}_{N_R|k_0}(l_1) & \bar{y}_{1|k_0}^*(l_2) & \cdots & \bar{y}_{N_R|k_0}^*(l_2) \end{bmatrix}^T \\
 &= \begin{bmatrix} \bar{h}_{k_0,11}(l_1) & \bar{h}_{k_0,12}(l_1) \\ \vdots & \vdots \\ \bar{h}_{k_0,N_R1}(l_1) & \bar{h}_{k_0,N_R2}(l_1) \\ \bar{h}_{k_0,12}^*(l_2) & -\bar{h}_{k_0,11}^*(l_2) \\ \vdots & \vdots \\ \bar{h}_{k_0,N_R2}^*(l_2) & -\bar{h}_{k_0,N_R1}^*(l_2) \end{bmatrix} \begin{bmatrix} \bar{d}_{k_0,1}(l_1) \\ \bar{d}_{k_0,2}(l_1) \end{bmatrix} + \begin{bmatrix} \bar{v}_1(l_1) \\ \vdots \\ \bar{v}_{N_R}(l_1) \\ \bar{v}_1^*(l_2) \\ \vdots \\ \bar{v}_{N_R}^*(l_2) \end{bmatrix} \\
 &= \bar{\mathbf{H}}_{k_0}(l) \bar{\mathbf{d}}_{k_0}(l) + \bar{\mathbf{v}}_{k_0}(l)
 \end{aligned} \tag{6.8}$$

Equalization can then be performed on all subcarriers using techniques such as Orthogonal Restoring Combining (ORC) or Minimum Mean Square Error (MMSE). The ORC and MMSE equalizer matrix can be obtained respectively as:

$$\begin{aligned}
 \bar{\mathbf{G}}_{ORC,k_0}(l) &= \begin{bmatrix} \alpha_{k_0,1}(l) & 0 \\ 0 & \alpha_{k_0,2}(l) \end{bmatrix} \begin{bmatrix} \bar{h}_{k_0,11}^*(l_1) & \cdots & \bar{h}_{k_0,N_R1}^*(l_1) & \bar{h}_{k_0,12}(l_2) & \cdots & \bar{h}_{k_0,N_R2}(l_2) \\ \bar{h}_{k_0,12}^*(l_1) & \cdots & \bar{h}_{k_0,N_R2}^*(l_1) & \bar{h}_{k_0,11}(l_2) & \cdots & \bar{h}_{k_0,N_R1}(l_2) \end{bmatrix} \\
 &\tag{6.9} \\
 \bar{\mathbf{G}}_{MMSE,k_0}(l) &= \left(\bar{\mathbf{H}}_{k_0}^H(l) \bar{\mathbf{H}}_{k_0}(l) + \sigma_n^2 \mathbf{I}_{2 \times 2} \right)^{-1} \bar{\mathbf{H}}_{k_0}^H(l) \tag{6.10}
 \end{aligned}$$

where $\alpha_{k_0,1}$ and $\alpha_{k_0,2}$ is defined respectively as:

$$\begin{aligned}
 \alpha_{k_0,1} &= \frac{1}{\sum_{q=1}^{N_R} |\bar{h}_{q1,k_0}(l_1)|^2 + \sum_{q=1}^{N_R} |\bar{h}_{q2,k_0}(l_2)|^2} \\
 \alpha_{k_0,2} &= \frac{1}{\sum_{q=1}^{N_R} |\bar{h}_{q1,k_0}(l_2)|^2 + \sum_{q=1}^{N_R} |\bar{h}_{q2,k_0}(l_1)|^2}
 \end{aligned} \tag{6.11}$$

From Equation (6.9), it can be seen that the performance of ORC equalizer depends on how close the values of adjacent tones, $\bar{h}_{k_0,qp}(l_1)$ and $\bar{h}_{k_0,qp}(l_2)$ match each other. In moderately frequency-selective fading channels, such condition is still likely to occur. However, in severely frequency-selective channels, the performance of ORC is expected to be inferior as the orthogonality in the channel matrix would now be lost.

6.3.3 Transmitter Model with Arbitrary Antenna Diversity ($N_T > 2$)

The proposed SFC BS-CDMA system with 2 transmit antennae has been discussed in section 6.3.1. A general SF encoding process with multiple antennae at the transmitter ($N_T > 2$) is analyzed in this section. In the following, a review on the fundamental principles of orthogonal design is presented, followed by the signal construction procedures to encode the transmitted single-carrier signals in spatial- and frequency-domain.

To begin with, Generalized Real Orthogonal Design (GROD), $\mathcal{G}_r$, and Generalized Complex Orthogonal Design (GCOD), $\mathcal{G}_c$, are given as [146][144]:

Definition 1 : Define $\mathbf{x} = [x_1, \dots, x_{N_S}]^T$ as the $N_S \times 1$ information vector to be sent and let $\mathcal{G}_r(\mathbf{x})$ be an $N_D \times N_T$ matrix with entries $0, \pm x_1, \dots, \pm x_{N_S}$, i.e. N_D time slots are required to transmit N_S bits using N_T antennae. If $\mathcal{G}_r^T(\mathbf{x})\mathcal{G}_r(\mathbf{x}) = \alpha(x_1^2 + \dots + x_{N_S}^2)\mathbf{I}_{N_T}$ with α positive, then $\mathcal{G}_r(\mathbf{x})$ is termed a GROD in variables $x_1, \dots, x_{N_S}$ of size $N_D \times N_T$ and $R = N_S/N_D$.

Definition 2 : Define $\mathbf{x} = [x_1, \dots, x_{N_S}]^T$, and let $\mathcal{G}_c(\mathbf{x})$ be an $N_D \times N_T$ matrix with entries $0, \pm x_1, \pm x_1^*, \dots, \pm x_{N_S}, \pm x_{N_S}^*$. If $\mathcal{G}_c^H(\mathbf{x})\mathcal{G}_c(\mathbf{x}) = \alpha(|x_1|^2 + \dots + |x_{N_S}|^2)\mathbf{I}_{N_T}$ with α positive, then $\mathcal{G}_c(\mathbf{x})$ is termed as GCOD in variables $x_1, \dots, x_{N_S}$ of size $N_D \times N_T$ and rate $R = N_S/N_D$.

The space-frequency (SF) encoding for BS-CDMA system with N_T transmit antennae, $N_S = N_T$ and $R = \frac{1}{2}$ is given by the following:

1. SF Encoder takes in $\frac{M}{2}$ (or MR) symbols and divides into N_T smaller sub-blocks $\mathbf{d}_{k,p}$, each of size $\frac{M}{2N_T}$

$$\mathbf{d}_{k,p} = \left[d((p-1)\frac{M}{2N_T} + 1) \dots d(\frac{pM}{2N_T}) \right]^T \text{ for } p = 1 \text{ to } N_T$$
2. Repeat or append $\mathbf{d}_{k,p}$, $2N_T$ times to form $\dot{\mathbf{d}}_{k,p}$

$$\dot{\mathbf{d}}_{k,p} = \left[(\mathbf{d}_{k,p})^T \dots (\mathbf{d}_{k,p})^T \right]^T$$
3. Perform cyclic conjugate on $\dot{\mathbf{d}}_{k,p}$ to produce $\ddot{\mathbf{d}}_{k,p}$

$$\ddot{\mathbf{d}}_{k,p} = \mathbf{P}_M^1 \dot{\mathbf{d}}_{k,p}^*$$
4. $\mathcal{G}_c$ can be constructed by replacing $x_1 \dots x_{N_S}$ with $\langle \dot{\mathbf{d}}_{k,1} \rangle_{\theta_a} \dots \langle \dot{\mathbf{d}}_{k,N_S} \rangle_{\theta_a}$ and $x_1^* \dots x_{N_S}^*$ with $\langle \ddot{\mathbf{d}}_{k,1} \rangle_{\theta_a} \dots \langle \ddot{\mathbf{d}}_{k,N_S} \rangle_{\theta_a}$, where $\theta_a = \frac{2\pi(a-1)}{M}$ and a is the column index of the entry to be replaced within $\mathcal{G}_c(\mathbf{x})$

As a result of phase-rotation operation by θ_a , the non-null tones of the repeated finite-length sequence exist only at subcarriers $l \in \Omega_a = [a, 2N_T + a, \dots]$. For illustration purpose, consider the example of the space-frequency coded BS-CDMA bits when $N_T = 4$. Figure 6.2 shows the tone arrangement of the transmitted signals (at symbol-level).

Frequency Index	$l_8 \in \Omega_8$	$\bar{d}_{k,4}^*(l_1)$	$\bar{d}_{k,3}^*(l_1)$	$\bar{d}_{k,2}^*(l_1)$	$\bar{d}_{k,1}^*(l_1)$
	$l_7 \in \Omega_7$	$\bar{d}_{k,3}^*(l_1)$	$\bar{d}_{k,4}^*(l_1)$	$\bar{d}_{k,1}^*(l_1)$	$\bar{d}_{k,2}^*(l_1)$
	$l_6 \in \Omega_6$	$\bar{d}_{k,2}^*(l_1)$	$\bar{d}_{k,1}^*(l_1)$	$\bar{d}_{k,4}^*(l_1)$	$\bar{d}_{k,3}^*(l_1)$
	$l_5 \in \Omega_5$	$\bar{d}_{k,1}^*(l_1)$	$\bar{d}_{k,2}^*(l_1)$	$\bar{d}_{k,3}^*(l_1)$	$\bar{d}_{k,4}^*(l_1)$
	$l_4 \in \Omega_4$	$-\bar{d}_{k,4}(l_1)$	$-\bar{d}_{k,3}(l_1)$	$\bar{d}_{k,2}(l_1)$	$\bar{d}_{k,1}(l_1)$
	$l_3 \in \Omega_3$	$-\bar{d}_{k,3}(l_1)$	$\bar{d}_{k,4}(l_1)$	$\bar{d}_{k,1}(l_1)$	$-\bar{d}_{k,2}(l_1)$
	$l_2 \in \Omega_2$	$-\bar{d}_{k,2}(l_1)$	$\bar{d}_{k,1}(l_1)$	$-\bar{d}_{k,4}(l_1)$	$\bar{d}_{k,3}(l_1)$
	$l_1 \in \Omega_1$	$\bar{d}_{k,1}(l_1)$	$\bar{d}_{k,2}(l_1)$	$\bar{d}_{k,3}(l_1)$	$\bar{d}_{k,4}(l_1)$
		1	2	3	4
		Transmit Ant. Index			

Figure 6.2: Tone Arrangement

6.4 Discussion

This section presents comparisons between STC and SFC BS-CDMA. The impact of channel variation on the system performance is analyzed, followed by the computational delay incurred in the two systems. The complexities involved at both the transmitter and receiver are analyzed separately. Lastly, the impact of transmitter diversity order on the system design is presented.

6.4.1 Channel Variation

It was highlighted earlier that in high mobility communication systems where channel coherence time is short, suitability of STC BS-CDMA system is limited as the total transmission duration of a single block is large. This means that the assumption of channel being invariant is no longer true. This will severely affect the effectiveness of ORC equalizer. In

SFC BS-CDMA, both original and encoded bits are transmitted simultaneously by using different set of tone indices. This makes transmission duration of a single block in SFC BS-CDMA much shorter than that in STC BS-CDMA. However, though coherence time is no longer an issue in SFC BS-CDMA, its performance is affected by channel coherence bandwidth. Since two adjacent subcarriers now carry the original tones and their conjugate pair, effectiveness of ORC in SFC BS-CDMA depends much on frequency non-selectivity of the channel i.e. adjacent tones should be almost identical to each other. However, channel variation due to short coherence time and narrow coherence bandwidth do not affect the performance of MMSE equalizer in both STC and SFC BS-CDMA systems.

6.4.2 Decoding Delay

In STC BS-CDMA system, the number of information bits in a single block is fixed as M which is the block size. When transmitter diversity increases, so does the time taken to transmit a single block which is given by $(\frac{MGN_T}{R})$. This means that the decoding delay experienced by the system grows with transmitter diversity. Logically, this is understandable as the number of symbols processed simultaneously is N_TM or proportional to the number of transmit antennae used.

In SFC BS-CDMA, the number of information bits in a single block is $(\frac{MR}{N_T})$ or is varying according to the transmitter diversity. The total time taken to transmit a single block is fixed as (MG) . This means that the length of decoding delay does not grow with transmitter diversity. Therefore, irrespective of transmit diversity used, decoding delay in SFC BS-CDMA is fixed by allowing the flexibility of processing fewer number of information bits simultaneously. This gives SFC BS-CDMA systems an edge over STC BS-CDMA systems in delay-critical applications.

6.4.3 Transmitter Complexity

The transmitter structure of STC BS-CDMA is relatively simple and makes it favourable for uplink transmission, where the transmitting device is mobile unit. STC BS-CDMA transmitter is largely identical to conventional direct sequence CDMA (DS-CDMA) transmitter used in today's 3G systems, except for the additional processing unit that inserts CP and performs STBC encoding. In SFC BS-CDMA, however, additional processing is required

to generate the repeated finite-length sequences and to phase-rotate the sequences before transmission.

6.4.4 Receiver Complexity

In general, block processing in BS-CDMA systems requires larger buffer size to store the $M \times G$ chip matrix at the receiver as compared to conventional CDMA processing. In addition to this, STC BS-CDMA systems need additional buffer of size $M \times \frac{N_R}{R}$ to temporarily store the despread symbols before processing them altogether. This additional buffer is not required in SFC BS-CDMA as a single block now contains all the encoded information.

6.4.5 Transmitter Diversity Impact

As mentioned earlier, STC BS-CDMA systems with higher transmitter diversity is not very suitable for delay-critical application as decoding delay grows proportionally with the number of transmit antennae employed. Secondly, large buffer size is required in systems with high diversity order.

On the other hand, processing complexity in SFC BS-CDMA transmitter grows with the increase in transmit diversity order as more phase-rotating operations are required. Moreover, as transmit diversity increases, more tones are used to carry redundant or duplicate information. This effectively means that energy of each bit is now distributed over fewer number of subcarriers which reduces the frequency diversity.

In brief, SFC BS-CDMA offers an attractive alternative to STC BS-CDMA systems, in particular, those delivering delay-critical applications or services and systems requiring lower receiver storage complexity. This gain is traded off with higher transmitter complexity and some loss in overall frequency diversity.

6.5 Numerical Results

This section presents the results from simulation study of SFC BS-CDMA under different practically realistic deployment scenarios. After that, the performance of SFC BS-CDMA

is compared with that of STC BS-CDMA system.

6.5.1 BER Performance of SFC BS-CDMA

The SFC BS-CDMA architecture was modelled and its performance was investigated using monte-carlo simulation for both uplink and downlink deployment. The entire simulation studies target high data rate and high mobility communication systems that delivers system performance improvement compared to the basic single antenna BS-CDMA system. SFC is built on top of the same philosophy behind STC architecture, that is to improve the error rate performance and hence allowing the use of higher modulation scheme. Higher modulation scheme, namely 16-QAM and 64-QAM, are being considered in addition to the basic BPSK and QPSK. The transmission carrier frequency considered is 5GHz, with a chip-rate of 40Mcps. The transmitter diversity is chosen to be two ($N_T = 2$). For BPSK and QPSK schemes, the maximum receiver diversity order used is 2. For 16-QAM and 64-QAM, the maximum receiver diversity order is increased to 3. Two different block sizes are being considered, namely $M = 256$ and $M = 512$. The length of CP employed is $L_{CP} = \frac{1}{4}M$. This effectively means that a single chip-block is of size 320 and 640 respectively. For $M = 256$, the duration of each chip-block, T , is $8\mu\text{secs}$, where $6.4\mu\text{secs}$ is the duration of information chips and $1.6\mu\text{secs}$ is the duration of guard interval. At the receiver, the system bandwidth of 40MHz is divided into M subchannels and equalized individually. The processing gain used are 8, 16 and 32. For the processing gain of 32, each user is transmitting at bandwidth efficiency of 0.125bps/Hz and 0.1875bps/Hz when 16-QAM and 64-QAM modulation is used respectively. The number of active users is varied according to different loading conditions in the system. The channel model assumes *Rayleigh* distribution based on Jakes' fading model [132] and the number of path is assumed to be 24. It is assumed that the maximum delay spread is less than the guard duration and hence no IBI occurs. Different vehicular mobility conditions are also used to simulate the practical deployment of typical mobile communication system, namely 0Hz (stationary), 463Hz (medium-high) and 926Hz (high). The complete set of simulation parameters is tabulated in Table 6.1.

Figure 6.3 shows the BER performance of a SFC BS-CDMA system with 2×2 antenna configuration using QPSK modulation and MMSE technique as the equalization method. From the Figure, it can be observed that the system performs generally well and sufficiently robust against MUI impairment. At moderately high mobility ($F_D = 463\text{Hz}$), the perfor-

Table 6.1: SFC BS-CDMA Systems: Simulation Parameters

Carrier Frequency, F	5 GHz
Chip rate, F_C	40.96 Mcps
Number of Tx Antenna, N_T	2
Number of Rx Antenna, N_R	1 2 3
Modulation	BPSK QPSK 16-QAM 64-QAM
OFDM / Chip-Block Size, M	256 (512)
FFT / IFFT Size, M	256 (512)
OFDM Block / Chip-Block Duration, T	8 μ secs (16 μ secs)
Data Duration per OFDM Block, T_D	6.4 μ secs (12.8 μ secs)
Guard Duration per OFDM Block, T_{CP}	1.6 μ secs (3.2 μ secs)
Super-Block Size, M_{SB}	PG OFDM Block
Super-Block Duration, T_{SB}	PG $\times$ 8 μ secs (PG $\times$ 16 μ secs)
Processing Gain, PG	8 16 32
Symbols per Super-Block, N_{SB}	256 (512)
Load Condition	1 user Half-Load Full-Load
Doppler Spread, F_D	0 Hz 463 Hz 926 Hz
Vehicular Mobility, V	0 kph 100 kph 200 kph
Channel Coherence Time, T_{CO}	∞ 387 μ secs 193 μ secs

mance degradation due to the increase in system loading condition is not substantial. At BER level of 1%, the system performance with half-load condition is close to that with single-user, while the full-load system performs only less than 0.5dB worse. The impact due to MUI becomes more significant for high mobility ($F_D = 926\text{Hz}$) conditions. Nonetheless, the maximum degradation impact due to full-load condition is still limited to less than 1.5dB at BER level of 1% and to less than 3.5dB at BER level of 0.1%.

Since a 2×2 QPSK-modulated SFC BS-CDMA system offers sufficiently robust performance during severely high mobility condition, a higher modulation scheme, namely 16-QAM, is investigated next. The BER performance of a 2×2 SFC BS-CDMA system employing 16-QAM modulation scheme and MMSE equalization technique is shown in Figure 6.4. From the Figure, it can be seen that the performance degradation due to MUI now starts to feature during moderately high mobility condition. Compared to the single-user performance with $F_D = 463\text{Hz}$, the half-load performance is about 0.5dB worse at BER level of 1%, while the system with full-load condition performs about 2.5dB worse. For extremely high mobility condition, the degradation impact due to MUI becomes more severe. With $F_D = 926\text{Hz}$, the half-load system performance is slightly less than 2dB inferior to the single-user performance at BER level of 1%. Meanwhile, the system performance under full-load condition exhibit an error floor of about 5%.

To enhance the spectral efficiency further, the modulation scheme is increased to 64-QAM. To compensate the performance deterioration due to higher constellation used, the number of antenna used at the receiver is increased to 3 and the BER performance is shown in Figure 6.5. From the Figure, it can be seen that the system performance is now more significantly affected by the presence of Doppler spread in the system. When $F_D = 463\text{Hz}$, the full-load performance exhibits an error-floor of about 1% and when $F_D = 926\text{Hz}$, the error-floor falls just slightly below the 10% level.

6.5.2 STC BS-CDMA vs SFC BS-CDMA: Performance Comparison

After the performance of SFC BS-CDMA is presented, a comparison between the performance of STC and SFC BS-CDMA is presented next. Figure 6.6 shows the comparison between the two systems with specific results under different loading conditions plotted in Figure 6.6(a), 6.6(b) and 6.6(c).

Figure 6.3: BER Performance of SFC BS-CDMA System (2×2 , QPSK)

Figure 6.4: BER Performance of SFC BS-CDMA System (2×2 , 16-QAM)

From Figure 6.6(a), it can be seen that at $F_D = 0\text{Hz}$, the performance of STC BS-CDMA is about 0.6dB better than that of SFC BS-CDMA. However, such advantage disappear at $F_D = 463\text{Hz}$. For $F_D = 926\text{Hz}$, the performance of STC BS-CDMA becomes 0.6dB worse than that of SFC BS-CDMA. As mentioned earlier, this behaviour is due to the fact that STC BS-CDMA requires the CCT to be twice as long as that required by SFC BS-CDMA system.

From Figure 6.6(b), it can be seen that at $F_D = 463\text{Hz}$, both systems perform almost identical. However, SFC BS-CDMA outperformn STC BS-CDMA at $F_D = 926\text{Hz}$. At $\text{BER} = 10^{-2}$, the SNR required by SFC BS-CDMA is 1.3dB lower than the value required by STC BS-CDMA. During full-load condition as shown in Figure 6.6(c), similar pattern or behaviour is observed between the two systems. From this comparison study, it is clear that SFC BS-CDMA offers an alternative to STC BS-CDMA, especially when the mobility condition is high. While both systems employ the same modulation and antenna configuration to achieve the same data rate, other factors such as the additional complexity required in SFC BS-CDMA needs to be taken into consideration before deciding which of the two schemes provide the most suitable option for a particular application.

Figure 6.5: BER Performance of SFC BS-CDMA System (2×3 , 64-QAM)

Figure 6.6: STC BS-CDMA vs SFC BS-CDMA - A Performance Comparison

6.6 Summary

In this Chapter, a MIMO BS-CDMA system employing space-frequency coding has been described and has been referred to as SFC BS-CDMA. SFC BS-CDMA was initially proposed to address some of the issues facing STC BS-CDMA system. It was proven that SFC BS-CDMA system is capable of shortening the long processing delay encountered in STC BS-CDMA system and the degree in which the system performance is affected by the mobility environment. This is at the expense of slight additional processing introduced at the transmitter. Similar to STC systems, SFC systems promise good error rate capability and hence allowing the use of higher and more efficient modulation schemes. With 2 and 3 receiver diversity order, 16-QAM and 64-QAM can be used and its performance is slightly superior than that delivered by STC BS-CDMA system. However, theoretical analysis of the proposed system is yet to be addressed and will form the key area as part of the future work. In addition, a thorough system-to-system comparison between SFC BS-CDMA system with multicarrier-based SFC system would be essential in determining the suitability of the two systems under various different deployment scenarios.

The contribution in this chapter can be summarized by the following points:

- a space-frequency block coded architecture for BS-CDMA system is presented and referred to as SFC BS-CDMA;
- construction of orthogonal block codes with arbitrary order of transmit diversity is presented;
- performance of SFC BS-CDMA system with different antenna configuration and mobility pattern obtained using extensive simulation study is presented;
- the performance gain in SFC BS-CDMA is substituted with the use of higher modulation scheme in order to deliver higher transmission rate; and
- comparison of various aspects of system design between STC and SFC BS-CDMA is presented.

Chapter 7

Improving the Performance of High Mobility BS-CDMA Communication Systems

7.1 Introduction

The fundamental principle of BS-CDMA originates from the fact that in future mobile communication systems, the chip rate is expected to be high enough such that duration of each chip is only a small fraction of the channel coherence time (CCT). While such assumption is perfectly valid when vehicular mobility is low, it might not be the case when BS-CDMA is expected to be deployed in highly mobile environment. Any violation to the assumption would lead to imperfect removal of MUI, giving rise to the term *residual MUI*. This Chapter looks at this issue and describes a solution to mitigate the degradation impact caused by such impairment.

Chapter 7 is organized as follows. Section 7.3 discusses the performance degradation encountered in high mobility transmission and introduces the weighting factor to improve the performance of MMSE equalizer. Section 7.4 discusses a parameter called Channel Coherence Ratio (CCR) which serves as the performance indicator of a BS-CDMA system. Section 7.5 discusses the possible interference cancellation techniques to eliminate the MUI components.

7.2 Motivation

The ideal condition for BS-CDMA to operation in is one where the channel fading condition remains largely stationary during the whole transmission of a single block of information symbols. In this case, the potential MUI when multiple users are actively transmitting can be perfectly removed, resulting in MUI-free performance even in the presence of high number of transmitting users. However, residual MUI arises in non-ideal condition and its impact on performance deterioration increases in severity as mobility of transmission increases. The residual MUI manifests itself as additive interfering component. The MMSE equalizer matrices presented in Chapter 3-6 do not consider this into account and is not effective in suppressing weak-subcarriers' noise during high SNR value. As a result, the performance of MMSE technique shows an unusual increasing error-rate during high SNR. The following section proposes the use of weighting factor to improve the performance of MMSE equalizer during high SNR condition. The subsequent section proposes interference cancellation schemes in order to improve the overall system performance.

7.3 Improved MMSE Equalizer with Weighting Factor

The MMSE equalizers proposed so far are derived based on the assumption of perfect channel stationarity condition. This section proposes improved MMSE equalizers that consider the residual MUI components that arise when the channel is not totally stationary. The MMSE equalizers are proposed for SISO BS-CDMA and LST BS-CDMA.

7.3.1 Problem Description

In BS-CDMA reception where the channel is stationary or slowly fading, AWGN becomes the only additive component as all signals belonging to different transmitting users are almost perfectly eliminated. However, when the channel is rapidly fading, the block despreading operation is no longer fully effective in removing the MUI. The residual MUI that remains after the despreading operation becomes a new additive component. When mobility condition is moderate, the residual MUI component is relatively insignificant and thus its exclusion from the MMSE equalizer design does not cause significant performance degra-

dation. When mobility condition is high, the residual MUI component becomes large and cannot be ignored. This can be seen from the performance shown in Figure 3.6(b), 4.4(b) and 4.5(b) where the performance starts to exhibit increasing BER behavior for large SNR. This is because for large SNR values, the coefficient used in the MMSE equalizer becomes too small. This would lead to amplification of the residual MUI component at those weak subcarriers (the same as in ZF case). The following sections study this problem in the context of SISO BS-CDMA and LST BS-CDMA. It can be shown that the BER performance for high SNR value can be considerably improved with the use of appropriate weighting factor in the MMSE equalizer.

7.3.2 SISO BS-CDMA

The ideal environment for BS-CDMA deployment is where the channel happens to be stationary. Under such scenario, the block despreading operation is given in Equation (3.9) as:

$$\mathbf{y}_{k_0} = \bar{\mathbf{H}}\mathbf{d}_{k_0} + \mathbf{v}_{k_0} \quad (7.1)$$

which is essentially free from any MUI. After applying FFT transformation to Equation (7.1), the MMSE equalizer can be derived as:

$$\bar{\mathbf{G}}_{MMSE,k_0} = (\bar{\mathbf{H}}_{k_0}^H \bar{\mathbf{H}}_{k_0} + \sigma_n^2 \mathbf{I}_{M \times M})^{-1} \bar{\mathbf{H}}_{k_0}^H \quad (7.2)$$

However, the result of the block despreading operation is no longer 'clean' when the channel happens to be varying (non-stationary) during the transmission of G chip-blocks. Under fast-fading environment, the result of the block despreading operation is given in Equation (3.24) as:

$$\mathbf{y}_{k_0} = \tilde{\mathbf{H}}_{k_0} \mathbf{s}_{k_0} + \text{Residual MUI} + \text{AWGN} \quad (7.3)$$

Previously, the MMSE equalizer was constructed by simply replacing $\bar{\mathbf{H}}_{k_0}$ in Equation (7.2) with the channel mean $\tilde{\mathbf{H}}_{k_0}$. However, the resulting MMSE equalizer does not take into consideration the *Residual MUI* and hence sub-optimum performance is expected when channel variation exists. The performance of BS-CDMA's MMSE equalizer can be improved by modelling the residual MUI as white Gaussian component with different variance for different mobility conditions.

Let v be the weighting factor that corresponds to an estimate of the variance. The improved MMSE equalizer is then given by:

$$\mathbf{G}_{MMSE,k_0} = (\tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0} + (\sigma_n^2 + v)\mathbf{I}_{M \times M})^{-1} \tilde{\mathbf{H}}_{k_0}^H \quad (7.4)$$

In Figure 7.1, the BER performance of the improved MMSE equalizer under different mobility conditions with different value of weighting factor is plotted. The system parameter used for simulation is identical to that described in Chapter 3. The performance shown in Figure 7.1 is based on SNR value of 30dB. From Figure 7.1, it can be seen that the most significant performance improvement is achieved for $F_D = 116\text{Hz}$ (50kph) at which the optimum weighting factor is found to be 0.05. For $F_D = 232\text{Hz}$, the BER improves by about a factor of 10 and the optimum weighting factor can be either 0.05 or 0.09. As the mobility condition becomes more severe, smaller performance improvement is attained. This is because the *Residual MUI* would now appear as AWGN with high power and therefore reducing the overall SNR of the system.

Figure 7.1: BER Performance of BS-CDMA with Improved MMSE Equalizer (SNR=30dB)

7.3.3 LST BS-CDMA

Similarly, the l^{th} tone of the FFT-transformed despread symbols for LST BS-CDMA system is now given as:

$$\bar{\mathbf{y}}_{k_0}(l) = \bar{\tilde{\mathbf{H}}}_{k_0}(l)\bar{\mathbf{d}}_{k_0}(l) + \text{Residual MUI} + \text{AWGN} \quad (7.5)$$

where $[\bar{\tilde{\mathbf{H}}}_{k_0}(l)]_{q,p} = \bar{\tilde{h}}_{k_0,qp}(l)$ and $\bar{\tilde{h}}_{k_0,qp}(l)$ is the mean of the l^{th} tone of the channel between the q^{th} receive antenna and k_0 user's p^{th} transmit antenna. Based on Equation (7.5), the improved MMSE equalizer for the l^{th} tone can be derived as:

$$\bar{\mathbf{G}}_{MMSE,k_0}(l) = (\bar{\tilde{\mathbf{H}}}_{k_0}\bar{\tilde{\mathbf{H}}}_{k_0}^H + (\sigma_n^2 + \nu)\mathbf{I}_{N_R \times N_R})^{-1}\bar{\tilde{\mathbf{H}}}_{k_0}^H \quad (7.6)$$

The BER performance of LST BS-CDMA deployed under different mobility condition with different value of weighting factor is shown in Figure 7.2. For $F_D = 116Hz$ and $232Hz$, the BER performance improves by a factor of 10 or more. The weighting values that can be used to obtain optimum BER performance at $F_D = 116Hz$ are 0.03, 0.07 and 0.11. Meanwhile, the value for $F_D = 232Hz$ is found to be 0.09. Similar to the SISO BS-CDMA case, the performance improvement decreases as the mobility condition becomes more severe.

7.4 Performance Criteria for BS-CDMA - Channel Coherence Ratio (CCR)

7.4.1 Problem Description

It has been shown that the performance of BS-CDMA deteriorates during high mobility condition when the impact due to residual MUI starts to be more significant. However, it is so far not clear at which point the impact due to residual MUI starts to take into effect. The following section describes Channel Coherence Ratio (CCR) as the criteria that can be used to determine the performance of a BS-CDMA system.

7.4.2 Channel Coherence Ratio (CCR)

In [147], it was shown that as the performance of a BS-CDMA system degrades in the presence of Doppler spread, the multi-user performance depends on the Channel Coherence Ratio (CCR) value instead of the block size used. CCR is defined as the ratio between duration of a single spreading module and channel coherence time. The CCR value was then proposed as a key criteria to determine the performance of a BS-CDMA system.

The variation of BER performance for four BS-CDMA (SISO, LST, STC and SFC BS-CDMA) systems under different CCR values is shown in Figure 7.3, 7.4, 7.5 and 7.6 respectively. The CCR values used for SISO and SFC BS-CDMA systems are 0.0404, 0.0808, 0.1616, 0.3232, 0.4849, 0.6465, 0.9697, 1.2929 and 2.5859. Meanwhile, the values used for LST BS-CDMA system are 0.058, 0.116, 0.466, 0.698, 1.117, 1.400 and 2.794. Lastly, the values used for STC BS-CDMA system are 0.058, 0.116, 0.233, 0.698, 1.117 and 2.235. Based on the results, two parameters determine the system performance: (1) Single-user threshold CCR value ($TCCR_{single}$); and (2) MUI threshold CCR value ($TCCR_{MUI}$). $TCCR_{single}$ indicates the point where single-user performance starts to degrade, while $TCCR_{MUI}$ indicates the point where MUI effect starts to occur.

Table 7.1: CCR Threshold Values for Different BS-CDMA Systems

System	Single-User	Multi-User
SISO BS-CDMA	1.9000	0.1616
LST BS-CDMA	1.4000	0.5500
STC BS-CDMA (ORC)	0.2500	0.5500
STC BS-CDMA (MMSE)	1.1000	0.5500
SFC BS-CDMA	0.6465	0.0808

7.5 Interference Cancellation for Uplink BS-CDMA in Broad-band Mobile Communication Systems

7.5.1 Problem Description

This thesis work focuses on devising BS-CDMA architectures that are suitable for high rate transmission with MUI suppression or avoidance capability. In an ideal scenario where the channel remains stationary for a long period of time, it is possible to achieve virtually MUI-

Figure 7.2: BER Performance of LST BS-CDMA with Improved MMSE Equalizer (SNR=30dB)

Figure 7.3: Impact of CCR on System Performance of BS-CDMA

Figure 7.4: Impact of CCR on System Performance of LST BS-CDMA

Figure 7.5: Impact of CCR on System Performance of STC BS-CDMA

free performance for the different SISO and MIMO BS-CDMA architectures. However, channel coherence time is inversely related to Doppler Frequency [132]. In other words, the degree in which wireless channel link condition varies increases along with the mobility condition. This means that the basic requirement for BS-CDMA to operate ideally is not met. This in turn results in the presence of residual MUI that contributes to significant performance degradation during high mobility transmission. The present work proposes two interference cancellation methods to address this issue. Leveraging on the fact that the mobility condition of transmitting stations usually differ in the uplink environment, the residual MUI caused by each user on other users also differ. This makes the mobility condition a suitable criteria in determining the detection ordering.

7.5.2 Ordering Criteria

In conventional CDMA such as DS-CDMA scheme, one dominant factor that gives rise to multi-user interference is the near/far phenomena. In the near/far scenario, users that are near to the receiving Base Station are likely to drown out the signal from those that are located further away. The priority for interference canceller in conventional CDMA scheme is hence to process those users with strong signal (equivalent to HIU in BS-CDMA) and then remove them before decoding those users with signal. In BS-CDMA however, the multi-user interference that arises from such near/far scenario can still be minimal. This is provided that those strong users are not moving. The multi-user interference in BS-CDMA originates from those users who are moving at high vehicular mobility. The ordering priority is hence different between an interference canceller for conventional CDMA system and that for BS-CDMA system. Secondly, it is highly likely that those users (in conventional CDMA system) with strong signal would be experiencing better channel quality than those weak users and hence better decoding quality. This property does not apply for BS-CDMA. Instead, a HIU in BS-CDMA experiences more rapidly fading channel conditions and hence more likely of producing worse decoding quality than a LIU. Due to these factors, two MUI cancellation schemes for BS-CDMA are proposed and will be described in the following sections.

According to [133], the channel coherence time (CCT) is inversely proportional to the Doppler spread and the two are related by the equation $T_{co} = \frac{9}{16\pi F_D}$ where T_{co} and F_D refers to CCT and Doppler frequency respectively. This means that transmitting devices

with high mobility will experience more rapidly fading channel as compared to those with low mobility. Based on this and Equation (3.24), it can be concluded that users with the highest mobility will cause the highest amount of interference to other users, but will suffer the lowest level of interference from other users and the reverse is true for users with the lowest mobility. As such, this thesis proposes the use of mobility condition as the ordering criteria for detection. Since users with high mobility are high interferer users (HIU) and users with low mobility are low interfere users (LIU), the detection performance of HIU is not affected much by LIU while the detection performance of LIU is severely affected by HIU. Hence, detection will start in the descending order of mobility of users.

7.5.3 Method I: Successive MUI Cancellation (SMC)

The SMC process is depicted in Figure 7.7 where user 1 has the highest mobility and user K has the lowest mobility such that $F_D(1) > \dots > F_D(K)$. As shown in the Figure, processing starts with user 1 during iteration 1. After block despreading, FFT, equalization and IFFT operations are completed, the estimate of the transmitted symbol vector $\hat{\mathbf{d}}_1$ obtained is then used to reconstruct the interference that user 1 is contributing to other users. In general, the recursive detection algorithm is given as:

1. $\hat{\mathbf{d}}_k = \mathbf{0}$ for $k = 1$ to K
2. for $p = 1$ to P ,
3. for $k_0 = 1$ to K ,
4. $\mathbf{r}_{k_0}(g) = \mathbf{r}(g) - \sum_{k \neq k_0} \mathbf{h}_k(g) \otimes \hat{\mathbf{d}}_k c_{k,g}$
5. $\mathbf{y}_{k_0} = \sum_{g=1}^G \mathbf{r}_{k_0}(g) c_{k_0,g}$
6. $\bar{\mathbf{y}}_{k_0} = \mathbf{F}_M \mathbf{y}_{k_0}$
7. $\bar{z}_{k_0}(l) = G_{EQ|k_0}(l) \bar{y}_{k_0}(l)$ for $l = 0$ to $M - 1$
8. $\mathbf{z}_{k_0} = \mathbf{F}_M^H \bar{\mathbf{z}}_{k_0}$
9. $\hat{\mathbf{d}}_{k_0} = \text{demod}(\mathbf{z}_{k_0})$
10. end
11. end

where $G_{EQ|k_0}(l)$ is the l^{th} tone equalizer for k_0^{th} user's signal, p corresponds to the current iteration index and P is the number of iteration to be carried out. $\text{demod}(\cdot)$ is the demodulation function that depends on whether the modulation scheme used. From the

Figure 7.6: Impact of CCR on System Performance of SFC BS-CDMA

Figure 7.7: Iterative Successive Interference Cancellation for BS-CDMA

above algorithm, it can be seen that during the first iteration, only MUI component from users with higher or equal mobility is being removed. During subsequent iterations, MUI component from both higher and lower mobility users is removed.

7.5.4 Method II: Multi-Stage SMC (MSMC)

The idea behind MSMC originates from the fact that the detection performance of LIU is severely affected by interference from HIU but not vice-versa. For illustration purpose, consider a system with 2 transmitting users (User 1 and 2). Assume that user 1 is moving at high mobility while user 2 is close to stationary ($F_D(1) \gg F_D(2)$). The channel coherence time of user 1 is much smaller than that of user 2. Since user 1's channel is rapidly fading, $\mathbf{h}_1(1) \neq \dots \neq \mathbf{h}_1(G)$ while the channel for user 2 remains largely stationary, $\mathbf{h}_2(1) \approx \dots \approx \mathbf{h}_2(G)$. When block despreading is performed on user 1, potential MUI from user 2 is completely eliminated. However, performing block despreading for user 2 leaves behind residual MUI contribution from user 1. This analysis can be simply extended to a system with more number of users. Based on this, MSMC technique assigns different priority to users with different mobility and users with similar mobility conditions are assigned the same priority. Users with higher priority will be detected first by using higher number of iterations. In this way, a better estimate for HIU is first obtained through a few iterations of detection before it is used for interference reconstruction and cancellation from the signal that is to be used for detecting low mobility users.

Figure 7.8 illustrates the MSMC processing for an uplink BS-CDMA receiver. For illustration purpose, the number of users is assumed to be 6 and their mobility speed can be divided into 3 groups. Group 1 and 3 consists of HIU with the highest mobility and LIU with lowest mobility respectively. As shown in Figure 7.8, detection is performed only on users belonging to Group 1 during stage 1. (Note that the detection process within a group is identical to the SMC technique). In stage 2, MUI component from Group 1 users is reconstructed using the estimates obtained in stage 1 and then eliminated from the originally received signal $\mathbf{r}(g)$. SMC detection is then performed on users belonging to Group 2. After detection for Group 2 is completed, MUI component from Group 2 users is reconstructed and then eliminated from the originally received signal $\mathbf{r}(g)$. Detection is then carried out for Group 3 users. The general recursive algorithm for MSMC procedure is given as follows.

1. $\hat{\mathbf{d}}_k = \mathbf{0}$;
2. for $p = 1$ to P
3. for $q = p$ to 1,
4. for $k_0 \in \Omega_q$
5. $\mathbf{r}_{k_0}(g) = \mathbf{r}(g) - \sum_{k \neq k_0} \mathbf{h}_k(g) \circledast \hat{\mathbf{d}}_k c_{k,g}$
6. $\mathbf{y}_{k_0} = \sum_{g=1}^G \mathbf{r}_{k_0}(g) c_{k_0,g}$
7. $\bar{\mathbf{y}}_{k_0} = \mathbf{F}_M \mathbf{y}_{k_0}$
8. $\bar{\mathbf{z}}_{k_0} = G_{EQ|k_0}(l) \bar{\mathbf{y}}_{k_0}(l)$
9. $\mathbf{z}_{k_0} = \mathbf{F}_M^H \bar{\mathbf{z}}_{k_0}$
10. $\hat{\mathbf{d}}_{k_0} = \text{demod} \langle \mathbf{z}_{k_0} \rangle$
11. end
12. end
13. end

where p corresponds to the current stage, P is the number of stages (which is assumed to be equivalent to the number of groups in this paper), q corresponds to the current group in the current stage and Ω_q is the vector containing the list of users belonging to the q^{th} group.

7.5.5 Numerical Results: SISO BS-CDMA

A BS-CDMA system with uplink transmission is simulated with carrier frequency of 5GHz and a total bandwidth of 40MHz. The block size used is 256. The length of CP is assumed to be one quarter of the block size and the maximum channel delay spread is assumed to be less than the length of this guard duration. The spreading factor used is 64. QPSK modulation is used in the simulation. Jakes' Rayleigh model [132] is used to simulate the mobile communication channel and the number of paths is set as 24.

7.5.5.1 Performance Comparison of Normal BS-CDMA vs BS-CDMA with Interference Cancellation

Figure 7.9(a) shows the BER performance of a BS-CDMA system where 50% of the active users are stationary while the other 50% of the active users are moving at 60kph (278Hz)

and 180kph (556Hz). In such scenario, the MSMC technique requires only 2 stages and 2 mobility groupings with 32 users in each group. For analysis purpose, let one detection process for one user be one processing unit (PU). This means that a total of 96 PUs are required when MSMC technique is employed. On the other hand, SMC technique with 2 iterations would require a total of 128 PUs. As can be seen in Figure 7.9(a), residual MUI causes performance degradation when 50% of the users are moving at a Doppler frequency of 278Hz. When 50% of the users are moving at Doppler frequency of 833Hz, BS-CDMA is no longer effective and error floor of about 10% exists. Meanwhile, both SMC and MSMC techniques effectively removes all residual MUI and performs almost identical to a system with perfect removal of MUI (100% – 0Hz). However, SMC performs better than MSMC when 50% of the users are moving with Doppler frequency of 833Hz. This is because in SMC, detection for all users are performed twice. This means that during the second iteration in SMC, interference reconstruction of HIU is based on the estimates obtained with 2 iterations and thus resulting in a 'cleaner' signal for detection of LIU and thus the superior performance.

7.5.5.2 Performance Comparison of SMC vs MSMC Techniques

Figure 7.9(b) depicts the BER performance of a BS-CDMA system with the following users' mobility configuration: 25% – 0Hz, 25% – 278Hz, 12.5% – 556Hz, 12.5% – 833Hz, 12.5% – 1.11kHz and 12.5% – 1.39kHz. This means that 6 stages are required for MSMC and it can be easily calculated that the total number of PUs required is 192. In the Figure, it can be seen that with SMC method, the second iteration offers significant performance improvement compared to the first iteration. However, the improvement margin decreases as the number of iteration increases and only about 0.2dB gain is achieved during the 4th iteration. On the other hand, the average error performance obtained using the MSMC technique improves consistently with each stage and the performance gap with a system with perfect MUI removal is only about less than 2dB at BER of 10^{-3} at the 6th stage. In comparison to SMC with 4 iterations where 256 PUs are required, MSMC is capable of delivering a better performance with lower processing complexity. With larger number of groups, MSMC will perform better than SMC because more detection or refinement is carried out on HIU before they are subtracted and detection for LIU is performed and the degree of error propagation from LIU will also be lower.

7.5.6 Numerical Results: LST BS-CDMA

A LST BS-CDMA system with uplink transmission is simulated with carrier frequency of 5GHz using a total transmission bandwidth of 40MHz. For this exercise, the block size used is 256. Two different antenna configuration is used, namely 2×2 and 4×4 . The length of CP is fixed as one-quarter of the block size ($L_{CP} = 64$) and the maximum channel delay spread is assumed to be less than the length of the guard duration. The spreading factor used is 64 and QPSK modulation is used for the simulation. The mobile communication channel is generated based on Jakes' Rayleigh fading model [132] and the number of paths is set at 24.

The LST BS-CDMA system is simulated with three different mobility condition. The first configuration has 50% of the transmitting users being stationary while the remaining 50% are moving at 60kph (278Hz). The second configuration has 50% of the transmitting users being stationary while the remaining 50% are moving at 180kph (833Hz). The third configuration is as follows: 25% – 0kph (0Hz), 25% – 60kph (278Hz), 12.5% – 120kph (556Hz), 12.5% – 180kph (833Hz), 12.5% – 240kph (1.11kHz) and 12.5% – 300kph (1.39kHz). The three systems are referred to System I, II and III respectively.

The MSMC interference cancellation is used to improve the performance of the LST BS-CDMA system. This means that for System I and II, the detection consists of 2 stages, where the first stage attempts to detect high mobility users, i.e. users' with mobility of 60kph and 180kph respectively. For System III, the users are split into 6 groups of different mobility and hence the whole detection process now consists of 6 stages.

As described earlier, the performance of MMSE equalizer of a BS-CDMA system with high mobility users can be improved substantially by introducing the weighting factor. However, the optimum weighting factor depends on the degree of the overall mobility condition and the antenna configuration that is being used.

In Figures 7.10 and 7.11, the performance of the 3 systems described earlier is shown. In Figure 7.10, the performance of 2×2 system is plotted while Figure 7.11 shows the performance of 4×4 system. The performance shown in both Figure 7.10 and 7.11 is obtained based on the SNR value of 20dB. From Figure 7.10, the optimum weighting factor with 2×2 antenna is found to be 0.075, 0.225 and 0.225 for System I, II and III respectively. From Figure 7.11, the optimum weighting factor with 4×4 antenna is found to be 0.15, 0.3

and 0.4 respectively.

After obtaining the optimum factor, the BER performance of each system is investigated. The system performance with 2×2 and 4×4 antenna is depicted in Figure 7.12 and 7.13 respectively. From both Figures, it is shown clearly that considerably good error-rate performance can be obtained with sufficient number of stages for different mobility conditions.

Figure 7.8: Multi-stage Parallel-Successive Interference Cancellation for BS-CDMA

(a) Performance Comparison of Normal BS-CDMA vs BS-CDMA with Interference Cancellation

(b) Performance Comparison of SMC vs MSMC Technique

Figure 7.9: BER Performance of BS-CDMA

Figure 7.10: BER vs Weight Factor for 2 × 2 LST BS-CDMA System

Figure 7.11: BER vs Weight Factor for 4 × 4 LST BS-CDMA System

Figure 7.12: BER Performance of a 2 x 2 LST BS-CDMA System

Figure 7.13: BER Performance of a 4 x 4 LST BS-CDMA System

7.6 Summary

In this Chapter, the performance degradation encountered in BS-CDMA system with the presence of high mobility users is being addressed. A key parameter referred to as Channel Coherence Ratio (CCR) has been defined as the performance indicator of any BS-CDMA system. The performance of the MMSE equalizer has also been shown to improve when the appropriate weighting factor is used to incorporate the MUI component. Lastly, two MUI interference cancellation methods have been proposed by exploiting the mobility conditions found in practical deployment. In practical scenarios where users' mobility conditions tend to vary, this mobility information can be used as the ordering criteria for detection.

The techniques proposed has so far been proven using results from extensive simulation studies. A framework that provides theoretical analysis to these techniques is non-trivial, but would be critical in order to complete the work. This issue will be addressed as part of the future works.

The contribution in this chapter can be summarized by the following points:

- a number of physical challenges that limit the performance of BS-CDMA systems are being addressed;
- improved MMSE equalizer that takes into consideration the presence of *residual MUI* is presented;
- a performance criteria to determine the performance of BS-CDMA systems is presented; and
- interference cancellation schemes to overcome the performance loss due to *residual MUI* are proposed using mobility condition as the ordering criteria.

Chapter 8

Conclusions and Future Works

8.1 Conclusions

The core of this thesis is based on a new paradigm in processing CDMA chips. Instead of the conventional method of processing each symbol individually, the block approach takes a block of symbols and process them collectively. The resulting chip transmission structure leads to MUI-free transmission when the channel is experiencing slow fading. This is true even during asynchronous arrival of signals from different users as code orthogonality is preserved in BS-CDMA. In addition, it allows rearrangement of the processing sequence at the receiver. By performing block despreading operation before the other symbol recovery processing operations, the subsequent operations will now operate on symbol-level instead of chip-level. This translates to a reduction of those operations by a factor of the processing gain. The performance of BS-CDMA is enhanced further by incorporating multiple antennae at the transmitter and receiver that results in MIMO architectures. The use of MIMO architectures is proven to enhance spectral efficiency without requiring additional bandwidth. Higher spectral efficiency can be achieved through a number of ways. Three MIMO architectures are studied and proposed in this thesis. The first architecture is LST BS-CDMA. In LST BS-CDMA, spectral efficiency is directly enhanced by exploiting the additional antennae to transmit independent data. In STC and SFC BS-CDMA, encoding is applied and the additional antennae are used to transmit the encoded information. While STC and SFC BS-CDMA do not directly enhance the transmission rate, the signal gain in error performance allow the use of higher modulation schemes and hence higher

spectral utilization. STC BS-CDMA applies encoding in spatial and temporal domain. SFC BS-CDMA applies encoding in spatial and frequency domain. STC BS-CDMA offers transmitter architecture with lower processing complexity than SFC BS-CDMA. However, SFC BS-CDMA offers a reduced computational delay than STC BS-CDMA, making it a more suitable choice for delay critical applications. The performance of MIMO architectures largely depends on the quality of CSI that is available at the receiver. In this context, design of pilot sequence and estimation algorithms to obtain CSI are addressed, in particular for the LST BS-CDMA architecture.

One drawback that comes with BS-CDMA is the performance susceptibility to the presence of Doppler shift. Extensive simulation studies in this direction show that the error rate degradation remains within an acceptable range during low to moderately high mobility environment. In a typical practical scenario, users transmitting in the uplink channels are likely to be moving at different vehicular mobility. This means that the degree of interference introduced by a user to the system is likely to be different from that introduced by another user moving at different mobility speed. Based on such typical deployment scenario, iterative interference can be performed by using the mobility condition as the detection ordering criteria.

A new parameter referred to as Channel Coherence Ratio (CCR) has been defined in this work. The CCR value in a BS-CDMA system serves as the key parameter in determining the overall system performance. The CCR value reflects the degree of channel variation that occurs within the total duration of transmitting a single block.

8.2 Future Works

8.2.1 Imperfect Chip Recovery due to Clock-Drift

As different crystal oscillators are used in the transmitter and receiver, a clock drift will always exist in reality. This means that taking the receiver clock as the reference, the clock at the transmitter can either be slower or faster. In uplink transmission, the situation becomes more complex as there are now multiple transmitting devices coming into the picture.

So far throughout this work, such drift is assumed to be absent. It means that it is assumed that a chip always arrive at the appropriate slot. In the presence of clock drift, such assumption is not always guaranteed. A thorough analysis is required on how the presence of such drift would impact the performance of a BS-CDMA system.

8.2.2 Variable Data Rate Transmission

One definite feature in any future communication system would be the asymmetric traffic between the upstream and downstream communications and the ability to dynamically adjust the data rate depending on the demand and channel conditions. Such variable rate capability can be attained using three different mechanisms. The first mechanism is the use of variable orthogonal spreading factor (VOSF) that allows the length of the spreading factor to be adjustable depending on the propagation channel conditions. The second mechanism is the multi-code assignment to different users depending on the data rate requirement. The third mechanism is to allow the use of different level of modulation schemes, depending on the channel conditions.

It is interesting to find out how BS-CDMA would perform under the different mechanism. The use of VOSF might have an impact on the chip transmission sequence and need to be addressed carefully. For the multi-code assignment, it could be a straight-forward extension to the current BS-CDMA system. However, the use of different level of modulation scheme need to be carefully analysed. The proper indicator for the the channel condition needs to be established. This is because a user with good channel condition might be moving with high mobility speed and hence might be a high source of interference to the system. Nonetheless, designing the proper parameter and determining the suitability of BS-CDMA for such deployment would be a challenging research issue.

8.2.3 Estimation of Mobility Condition

It is now clear that the performance of BS-CDMA is influenced by the mobility of the transmitting device and the degree of mobility of other transmitting devices. The performance degradation due to the *residual* MUI that arises during high mobility can be mitigated with the use of iterative interference cancellation. The interference cancellation techniques that have been proposed so far use the mobility condition as the criteria in determining the de-

tection ordering. However, such information is assumed to be available and no method has been devised to estimate the information when it is unavailable. Hence, a good estimation algorithm is critical in enabling the practical viability of the proposed interference cancellation methods. Another area of interest is the study on how the error in the estimation affects the performance of the iterative cancellation algorithms. In addition, such information will also be required to support the dynamic rate capability as it needs to determine the channel conditions beforehand. Based on initial analysis, the degree of mobility of a BS-CDMA transmitting user can be determined from the estimates of the CSI that have been obtained. The task of estimating this parameter would be a critical and challenging research area.

8.2.4 Throughput Performance Index

In this work, a basic BS-CDMA system was initially introduced. This is followed by its MIMO variants, namely LST BS-CDMA, STC BS-CDMA and SFC BS-CDMA. The motivation behind such MIMO architectures is to enhance the transmission rate capability or in short, to increase the overall system throughput. So far, it has been mentioned that all the three architectures are capable of delivering higher rate. However, it is unclear which one delivers the best performance enhancement. For example, while LST BS-CDMA promises higher data rate enhancement, a higher SNR value is required to obtain a particular level of BER as compared to as STC or SFC BS-CDMA system. Meanwhile, with the significantly lower error rate, STC and SFC BS-CDMA allow the use of higher modulation schemes. On the other hand, STC and SFC BS-CDMA system churns out data at a lower rate compared to LST BS-CDMA. It is thus important to determine what constitutes a good throughput. The task of defining the performance index would be useful and remains a challenging research area that need to be addressed.

8.2.5 Interference Analysis

In this work, various BS-CDMA architectures have been proposed. The performance of each architecture has been extensively studied under different channel conditions when the transmission mobility ranges from low to high. The impact of residual MUI on system performance has also been addressed thoroughly. The practicality of the study can be fur-

ther extended to consider interference analysis. The scarcity of air spectrum means that any wireless communication system is most likely to be sharing the transmission medium with other systems. One interesting research issues is the intercell interference that arises when neighboring cells' Base Stations are transmitting with the same frequency. Various design considerations such as cell-size, transmission power and frequency reuse factor would determine the severity of intercell interference that will arise. Another interesting research issue is the interference robustness due to the presence of other systems transmitting on adjacent frequency channels. In some cases, the spectral sidelobes coming from adjacent frequency channels might cause significant level of interference that might lead to considerable performance loss. An in-depth study on interference analysis is essential in order to further justify the practicality and co-existence capability of BS-CDMA systems.

8.2.6 Performance Analysis of BS-CDMA Systems in Nakagmi-m Channels

The channel model used throughout the simulation studies of this thesis work is based on the Rayleigh fading model. A recent study shows that in many urban environments, Nakagami distribution is the best-fit distribution model for the wireless propagation channel parameters obtained from field measurements. In MIMO systems, the actual diversity gain is largely influenced by the degree of correlation among the multiple antennas. A low correlation or independently fading characteristics between the antennas translates to higher diversity gain. However, the physical constraints such as insufficient spatial spacing between the antennas often results in some degree of correlation between them. In such cases, the diversity gain is reduced. The mathematical framework of Nakagami distribution model allows such channels to be generated and the degree of correlation can be easily determined by the m -parameter. Investigating the performance of BS-CDMA systems under Nakagami fading channels would be part of the future work of this thesis.

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Appendix A

Basic Matrices

A.1 Toeplitz Matrix

A Toeplitz matrix, $\mathbf{A}$, is an $n \times n$ matrix when the entries are given as follows:

$$\mathbf{A} = \begin{bmatrix} a_0 & a_{-1} & a_{-2} & \cdots & a_{-(n-1)} \\ a_1 & a_0 & a_{-1} & \ddots & \vdots \\ a_2 & a_1 & a_0 & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ a_{n-1} & \cdots & \cdots & \cdots & a_0 \end{bmatrix} \quad (\text{A.1})$$

A.2 Circular Toeplitz Matrix

A special form of Toeplitz matrix is the circular Toeplitz matrix where the columns are circular-shifted versions of the first column. The entries of a circular Toeplitz matrix are given as:

$$\mathbf{A} = \begin{bmatrix} a_0 & a_{n-1} & a_{n-2} & \cdots & a_1 \\ a_1 & a_0 & a_{n-1} & \ddots & \vdots \\ a_2 & a_1 & a_0 & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ a_{n-1} & \cdots & \cdots & \cdots & a_0 \end{bmatrix} \quad (\text{A.2})$$

One nice property of a circular Toeplitz matrix is that it is diagonalizable by pre- and post-multiplying it with IFFT and FFT matrix. This property can be expressed as:

$$\mathbf{F}_n \mathbf{A} \mathbf{F}_n^H = \begin{bmatrix} \bar{a}_0 & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \bar{a}_{n-1} \end{bmatrix} \quad (\text{A.3})$$

where $\mathbf{F}_n$ is the $n \times n$ FFT matrix for a n-point FFT operation. The l^{th} diagonal entry, $\bar{a}(l)$ is simply the l^{th} frequency tone of the first column vector of $\mathbf{A}$, given as $\begin{bmatrix} a_0 & a_1 & \cdots & a_{n-1} \end{bmatrix}^T$.

Appendix B

Polynomial Functions

B.1 Laguerre Polynomials

The associated *Laguerre* polynomials are given by the formula

$$L_n^k(x) = \frac{1}{n!} \sum_{i=0}^n \frac{n!}{i!} \binom{k+n}{n-i} (-x)^i \quad (\text{B.1})$$

where $\binom{n}{k}$ is the binomial coefficient. The first few associated *Laguerre* polynomials are:

$$\begin{aligned} L_0^k(x) &= 1 \\ L_1^k(x) &= -x + k + 1 \\ L_2^k(x) &= \frac{1}{2}[x^2 - 2(k+2)x + (k+1)(k+2)] \\ L_3^k(x) &= \frac{1}{6}[-x^3 + 3(k+3)x^2 - 3(k+2)(k+3)x + (k+1)(k+2)(k+3)] \end{aligned} \quad (\text{B.2})$$

With regards to *Laguerre* polynomial, the following definite integral property is true:

$$\int_0^\infty t^{\alpha-1} e^{-pt} L_a^\beta(pt) L_b^\gamma(pt) dt = \frac{p^{-\alpha} \Gamma(\alpha) \Gamma(b - \alpha + \gamma + 1) \Gamma(a + \beta + 1)}{a! b! \Gamma(1 - \alpha + \gamma) \Gamma(\beta + 1)} {}_3F_2(-a, \alpha, \alpha - \gamma; -b + \alpha - \gamma, \beta + 1; 1) \quad (\text{B.3})$$

where $\Gamma(n) = (n - 1)!$ is the gamma function. ${}_pF_q(a_1, \dots, a_p; b_1, \dots, b_q; x)$ refers to the generalized hypergeometric function which can be defined in the form of a hypergeometric series, i.e. a series for which the ratio of successive terms can be written as:

$$\frac{c_{k+1}}{c_k} = \frac{(k + a_1)(k + a_2) \dots (k + a_p)}{(k + b_1)(k + b_2) \dots (k + b_q)} x \quad (\text{B.4})$$

Therefore, a generalized hypergeometric function ${}_pF_q(a_1, \dots, a_p; b_1, \dots, b_q; x)$ has p parameters of type 1 and q parameters of type 2.

(B.5)

Appendix C

Derivation of Theoretical Symbol-wise SNR for BS-CDMA system

C.1 Zero-Forcing Technique

Recall from Equation (3.24) that the result of the block despreading operation is given by:

$$\begin{aligned}
 \mathbf{y}_{k_0} &= \sum_{g=1}^G \sum_{k=1}^K (\tilde{\mathbf{H}}_k + \dot{\mathbf{H}}_k(g)) \mathbf{s}_k c_{k,g} c_{k_0,g} + \mathbf{v} \\
 &= \underbrace{\tilde{\mathbf{H}}_{k_0} \mathbf{s}_{k_0} \left(\sum_{g=1}^G c_{k_0,g}^2 \right)}_1 + \underbrace{\sum_{k \neq k_0} \tilde{\mathbf{H}}_k \mathbf{s}_k \left(\sum_{g=1}^G c_{k_0,g} c_{k,g} \right)}_0 + \sum_{k=1}^K \dot{\mathbf{H}}_k(g) \mathbf{s}_k \underbrace{\left(\sum_{g=1}^G c_{k_0,g} c_{k,g} \right)}_{\neq 0} + \mathbf{v} \\
 &= \tilde{\mathbf{H}}_{k_0} \mathbf{s}_{k_0} + \underbrace{\sum_{k=1}^K \dot{\mathbf{H}}_k(g) \mathbf{s}_k \left(\sum_{g=1}^G c_{k_0,g} c_{k,g} \right)}_{\text{Residual MUI}} + \mathbf{v}
 \end{aligned} \tag{C.1}$$

and the ZF multiplier matrix as:

$$\mathbf{G}_{k_0, ZF} = \gamma (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \tilde{\mathbf{H}}_{k_0} \tag{C.2}$$

After applying the ZF equalization, the recovered symbol vector is given as:

$$\begin{aligned}\hat{\mathbf{s}}_{k_0} &= \mathbf{G}_{k_0, ZF} \mathbf{Y}_{k_0} \\ &= \mathbf{s}_{k_0} + \sum_{k=1}^K \sum_{g=1}^G \gamma (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \dot{\mathbf{H}}_{k_0}(g) \mathbf{s}_k c_{k,g} c_{k_0,g} + \gamma (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \tilde{\mathbf{H}}_{k_0} \mathbf{v}\end{aligned}\quad (\text{C.3})$$

Now, the covariance matrix of $\hat{\mathbf{s}}_{k_0}$ can be derived as:

$$\begin{aligned}E[\hat{\mathbf{s}}_{k_0} \hat{\mathbf{s}}_{k_0}^H] &= E[\mathbf{s}_{k_0} \mathbf{s}_{k_0}^H] + \sum_k \sum_{g=1}^G E\left[\frac{\gamma^2}{G^2} (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \tilde{\mathbf{H}}_{k_0}^H \dot{\mathbf{H}}_k(g) \mathbf{s}_k \mathbf{s}_k^H \dot{\mathbf{H}}_k^H(g) \tilde{\mathbf{H}}_{k_0} (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1}\right] \\ &\quad + E\left[\gamma^2 (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \tilde{\mathbf{H}}_{k_0}^H \mathbf{v} \mathbf{v}^H \tilde{\mathbf{H}}_{k_0} (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1}\right] \\ &= \sigma_s^2 \mathbf{I}_M + \sum_{k=1}^K \sum_{g=1}^G \frac{1}{G^2} \sigma_s^2 (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \dot{\mathbf{H}}_k(g) \dot{\mathbf{H}}_k^H(g) \tilde{\mathbf{H}}_{k_0}^H (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \\ &\quad + \gamma^2 \sigma_n^2 (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0} (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1} \\ &= \underbrace{\sigma_s^2 \mathbf{I}_M}_{\text{Desired Component}} + \underbrace{\frac{\sigma_s^2}{G^2} \sum_{k=1}^K \sum_{g=1}^G \left\{ (\mathbf{G}_{k_0, ZF} \dot{\mathbf{H}}_k(g)) (\mathbf{G}_{k_0, ZF} \dot{\mathbf{H}}_k(g))^H \right\}}_{\text{Residual MUI}} \\ &\quad + \underbrace{\sigma_n^2 (\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1}}_{\text{AWGN}}\end{aligned}\quad (\text{C.4})$$

Now, let $\lambda_{kk_0,g,m} = [\mathbf{G}_{k_0, ZF} \dot{\mathbf{H}}_k(g)]_{m,m}$ and $\beta_{k_0,m} = [(\gamma \tilde{\mathbf{H}}_{k_0}^H \tilde{\mathbf{H}}_{k_0})^{-1}]_{m,m}$. The power of the m^{th} recovered symbol, $\hat{s}_{k_0}(m)$, can be expressed as:

$$\begin{aligned}SNR_m &= \frac{\sigma_s^2}{\sigma_s^2 \left\{ \beta_{k_0,m} + \frac{1}{G^2} \sum_{k=1}^K \sum_{g=1}^G |\lambda_{kk_0,g,m}|^2 \right\}} \\ &= \frac{1}{\beta_{k_0,m} + \frac{1}{G^2} \sum_{k=1}^K \sum_{g=1}^G |\lambda_{kk_0,g,m}|^2}\end{aligned}\quad (\text{C.5})$$

C.2 Minimum Mean Square Error (MMSE) Technique

Firstly, recall that the result of the block despreading operation is given in Equation (C.1). The MMSE multiplier matrix can then be worked out as:

$$\mathbf{G}_{k_0, MMSE} = \gamma \tilde{\mathbf{H}}_{k_0}^H (\gamma \tilde{\mathbf{H}}_{k_0} \tilde{\mathbf{H}}_{k_0}^H + \mathbf{I}_{M \times M})^{-1}\quad (\text{C.6})$$

After applying the MMSE equalization, the recovered symbol is given as:

$$\begin{aligned}\hat{\mathbf{s}}_{k_0} &= \mathbf{G}_{k_0, MMSE} \mathbf{y}_{k_0} \\ &= \gamma \tilde{\mathbf{H}}_{k_0}^H (\gamma \tilde{\mathbf{H}}_{k_0} \tilde{\mathbf{H}}_{k_0}^H + \mathbf{I}_{M \times M})^{-1} \tilde{\mathbf{H}}_{k_0} \mathbf{s}_{k_0} \\ &\quad + \sum_{k=1}^K \sum_{g=1}^G \gamma \tilde{\mathbf{H}}_{k_0}^H (\gamma \tilde{\mathbf{H}}_{k_0} \tilde{\mathbf{H}}_{k_0}^H + \mathbf{I}_{M \times M})^{-1} \dot{\mathbf{H}}_k(g) + \gamma \tilde{\mathbf{H}}_{k_0}^H (\gamma \tilde{\mathbf{H}}_{k_0} \tilde{\mathbf{H}}_{k_0}^H + \mathbf{I}_{M \times M})^{-1} \mathbf{v}_{k_0}\end{aligned}\quad (\text{C.7})$$

Now, the covariance matrix of $\hat{\mathbf{s}}_{k_0}$ can be derived as:

$$\begin{aligned}E[\hat{\mathbf{s}}_{k_0} \hat{\mathbf{s}}_{k_0}^H] &= \mathbf{G}_{k_0, MMSE} E[\mathbf{y}_{k_0} \mathbf{y}_{k_0}^H] \mathbf{G}_{k_0, MMSE}^H \\ &= \mathbf{G}_{k_0, MMSE} \tilde{\mathbf{H}}_{k_0} E[\mathbf{s}_{k_0} \mathbf{s}_{k_0}^H] \tilde{\mathbf{H}}_{k_0}^H \\ &\quad + \mathbf{G}_{k_0, MMSE} \left(\sum_{k=1}^K \sum_{g=1}^G \dot{\mathbf{H}}_k(g) E[\mathbf{s}_k \mathbf{s}_k^H] \dot{\mathbf{H}}_k^H(g) \mathbf{G}_{k_0, MMSE}^H + \mathbf{G}_{k_0, MMSE} E[\mathbf{v}_{k_0} \mathbf{v}_{k_0}^H] \right) \\ &= \mathbf{G}_{k_0, MMSE} \left(\sigma_n^2 \left(\frac{\sigma_s^2}{\sigma_n^2} \tilde{\mathbf{H}}_{k_0} \tilde{\mathbf{H}}_{k_0}^H + \mathbf{I}_{M \times M} \right) \right) \mathbf{G}_{k_0, MMSE}^H \\ &\quad + \frac{1}{G^2} \mathbf{G}_{k_0, MMSE} \left(\sum_{k=1}^K \sum_{g=1}^G \dot{\mathbf{H}}_k(g) \dot{\mathbf{H}}_k^H(g) \right) \mathbf{G}_{k_0, MMSE}^H \\ &= \underbrace{\sigma_s^2 \mathbf{G}_{k_0, MMSE} \tilde{\mathbf{H}}_{k_0}}_{\text{Desired Component}} + \underbrace{\frac{1}{G^2} \mathbf{G}_{k_0, MMSE} \left(\sum_{k=1}^K \sum_{g=1}^G \dot{\mathbf{H}}_k(g) \dot{\mathbf{H}}_k^H(g) \right) \mathbf{G}_{k_0, MMSE}^H}_{\text{Residual MUI}}\end{aligned}\quad (\text{C.8})$$

Now, let $a_{k_0, m}$ and $b_{k_0, m}$ be the following:

$$\begin{aligned}a_{k_0, m} &= [\mathbf{G}_{k_0, MMSE} \tilde{\mathbf{H}}_{k_0}]_{m, m} \\ b_{k_0, m} &= [\mathbf{G}_{k_0, MMSE} (\sum_{k=1}^K \sum_{g=1}^G \dot{\mathbf{H}}_k(g) \dot{\mathbf{H}}_k^H(g)) \mathbf{G}_{k_0, MMSE}^H]_{m, m}\end{aligned}\quad (\text{C.9})$$

This means that the m^{th} detected symbol has the following variance:

$$E[|\hat{s}_{k_0, m}|^2] = \sigma_s^2 a_{k_0, m} + \frac{1}{G^2} b_{k_0, m} \quad (\text{C.10})$$

However, the m^{th} detected symbol can be expressed as $\hat{s}_{k_0, m} = a_{k_0, m} s_{k_0, m} + v_{k_0, m}$, where $v_{k_0, m}$ corresponds to the irreducible noise component after equalization. The variance can then be expressed as:

$$E[|\hat{s}_{k_0, m}|^2] = a_{k_0, m}^2 \sigma_s^2 + E[|v_{k_0, m}|^2] \quad (\text{C.11})$$

By substituting Equation (C.10) into (C.11), the variance of the noise is found to be:

$$\begin{aligned}\sigma_s^2 a_{k_0,m} + \frac{1}{G^2} b_{k_0,m} &= a_{k_0,m}^2 \sigma_2^2 + E[|v_{k_0,m}|^2] \\ E[|v_{k_0,m}|^2] &= \sigma_s^2 a_{k_0,m}(1 - a_{k_0,m}) + \frac{1}{G^2} b_{k_0,m}\end{aligned}\tag{C.12}$$

and the SNR of the m^{th} symbol is given by:

$$SNR_m = \frac{\sigma_s^2 a_{k_0,m}^2}{\sigma_s^2 a_{k_0,m}(1 - a_{k_0,m}) + \frac{1}{G^2} b_{k_0,m}}\tag{C.13}$$

Appendix D

Spectral Characteristics of Pilot Sequence

From Equation (4.17), the l^{th} entry of the phase-rotated repeated finite-length sequence to be transmitted from the p^{th} antenna is given as $\tilde{t}_p(l) = [\tilde{\mathbf{t}}_p(l)] = \hat{t}_p(l)e^{j\theta_p l}$. The m^{th} tone of the transmitted pilot sequence is then given as:

$$\begin{aligned}
 \bar{t}_p(m) &= \sum_{l=0}^{M-1} \tilde{t}_p(l) e^{-\frac{j2\pi ml}{M}} \\
 &= \sum_{l=0}^{M-1} \hat{t}_p(l) e^{\frac{j2\pi(p-1)l}{M}} e^{-\frac{j2\pi ml}{M}} \\
 &= \sum_{l=0}^{\frac{M}{N_T}-1} t_p(l) \left(\sum_{p_1=1}^{N_T} e^{\frac{j2\pi(p-1)(l+\frac{(p_1-1)M}{N_T})}{M}} e^{-\frac{j2\pi m(l+\frac{(p_1-1)M}{N_T})}{M}} \right) \\
 &= \sum_{l=0}^{\frac{M}{N_T}-1} t_p(l) e^{-\frac{j2\pi(m-p-1)l}{M}} \sum_{p_1=1}^{N_T} e^{-\frac{j2\pi(m-p-1)(p_1-1)}{N_T}} \\
 &= \begin{cases} N_T \sum_{l=0}^{\frac{M}{N_T}-1} t_p(l) e^{-\frac{j2\pi(m-p-1)l}{M}} & \text{for } m = p-1, N_T + p-1, \dots \\ 0 & \text{Otherwise} \end{cases}
 \end{aligned} \tag{D.1}$$