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Multi-user detection for multi-rate DS/CDMA systems

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ABSTRACT

MULTI-USER DETECTION FOR MULTI-RATE DS/CDMA SYSTEMS

by
Jinwen Ma

Wireless cellular communication is witnessing a rapid growth in market, technology and range of services. Current and future demands for wireless communication services motivate the need for handling multi-media traffic types. In a multimedia communication system, users with different and even time-varying rates and quality of services (QoS) requirements, such as voice, image and data, must be accommodated. The use of Spread Spectrum modulation with Code Division Multiple Access (CDMA) technology is an attractive approach for economical spectrally efficient and high quality cellular and personal communication services. This dissertation explores the technologies of applying different interference cancellation techniques to multi-rate CDMA systems that serve users with different QoS.

Multiple Access Interference (MAI) and multipath propagation are the major issues in wireless communication systems. It is also true for multi-rate CDMA systems. Multiuser detection has been shown to be effective in combating the near-far problem and providing superior performance over conventional detection method. In this dissertation, we combine both linear and non-linear multiuser detection techniques, such as linear minimum mean squared error (LMMSE) detector, non-linear decision feedback detector, with other signal processing techniques, such as array processing and multipath combining, to create effective near-far resistant detectors for multi-rate CDMA systems.

Firstly, we propose MMSE receivers for synchronous multi-rate CDMA system and compare the performance with the corresponding multi-rate decorrelating detectors. The multi-rate decorrelating detector is optimally near-far resistant and easy to implement. The proposed linear MMSE multi-rate receiver can be adaptively implemented only with the knowledge of the desired user. Due to the fact that MMSE detector offers best trade-off between the MAI cancellation and noise variance enhancement , it is shown that multi-rate MMSE receiver can offer better

performance than the multi-rate decorrelator when the interfering users' Signal to Noise Ratio (SNR) is relatively low comparing to the desired user's SNR.

Secondly, for asynchronous multi-rate CDMA system, we propose multi-rate multi-shot decorrelating detectors and multi-rate multi-shot MMSE detectors. The performance of multi-shot detectors can be improved monotonically with increasing the number of stacked bits, but a great computational complexity is going to be introduced in order to get better performance. A debiasing method is introduced to multi-rate multi-shot linear detectors. Debiasing method optimizes multi-rate detectors based on the multi-rate multi-shot model. Debiasing multi-shot MMSE detector for multi-rate signals can offer better performance than the corresponding debiasing multi-shot decorrelating detector.

Thirdly, we propose linear space-time receivers for multi-rate CDMA systems. The minimum mean-squared error criteria is used. We perform a comparative study on the multi-rate receiver which uses either multipath (temporal) processing or array (spatial) processing, and the one which uses both array and multipath (space-time) processing. The space-time receiver for the multi-rate CDMA signals give us the potential of improving the capacity of multi-rate systems. The space-time processing combined with multiuser detection have the advantages of combating multipath fading through temporal processing, reducing MAI through MMSE method and provide antenna or diversity gain through spatial processing and increasing the capacity of the multi-rate CDMA systems.

Lastly, the group-wise interference cancellation methods are proposed for multi-rate CDMA signals. The non-linear decision feedback detection (DFD) schemes are used in the proposed receivers. The proposed interference cancellation schemes benefit from the nature of the unequal received amplitudes for multi-rate CDMA signals. Users with same data rate are grouped together. Users with the highest data-rate are detected first. Interference between the groups is cancelled in a successive order. The results show that the group-wise MMSE DFD yields better performance than multi-rate linear MMSE detector and multi-rate decorrelating detector, especially for highly loaded CDMA systems.

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MULTI-USER DETECTION FOR MULTI-RATE DS/CDMA SYSTEMS

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CHAPTER 1

INTRODUCTION

1.1 General Overview

Access to the developing information highway occurs through two avenues: fixed, wired systems (such as phone sockets in the wall or Ethernet connections) or wireless cellular systems (such as wireless local area networks, cellular telephone systems, or satellite communication systems). While we can theoretically increase, without bound, the data throughput in wired communication systems by laying more copper wire or fiber optic cable, we are unable to provide analogous gains in wireless communication systems because of the finite nature of the frequency spectrum. In addition, the presence of interfering users, channel fading, and multi-path propagation make the problem of wireless communication over these channels more difficult. Therefore, one purpose of our work, is to provide methods for effective and computationally efficient ways for communicating over such wireless channels.

The wireless communications systems are going to be required not just handling speech, but also handling high speed data and images. Users with different and even time-varying information rates must be accommodated in the future wireless communication systems. A mobile terminal may set up and modify sessions for voice, data and images through wireless connections to the base stations. So the objective of this work is to provide methods for integrating the wireless communication systems with existing communication networks in order to communicating multimedia type traffic.

In order to state our purpose more precisely and to provide more specific motivation, we discuss the architecture of a cellular phone system which we will use throughout this work as a representative wireless communication system. In this system, service area is partitioned into cells. A base station lies at the center of each cell and acts as an interface between the wireless user and the switched telephone network. If one were to make a cellular phone call, the signal would first be transmitted to the base station which corresponds to that user's cell; the base station receives the signal and transmits the cell over the switched network to its desti-

nation. The basic system model of the remote user and base station can be applied to various wireless systems with different channel characteristics and cell sizes.

The base station receives signals from multiple users and it must demodulate their signals prior to further transmission towards a final destination. As the popularity of wireless communication services grows and as multimedia data services are going to be offered, more sophisticated base station receiver will be needed to handle the increased demand. The motivation of our work is to explore efficient demodulation methods under the multimedia traffic type for base station receivers. We will explore the potentials of using linear and non-linear detection schemes, for multi-rate traffic type, in conjunction with other signal processing techniques as a basis for receivers design. In the next section, we will discuss the CDMA multiple access method which is emerging as a standard for current and future wireless communication systems.

1.2 CDMA Background

Previous cellular systems, such as the AMPS [1] cellular system, used analog technology. Many current and most future systems is using and will use digital technology to take advantage of advanced signal processing methods to increase the systems' efficiency. Time Division Multiple Access (TDMA) is the mulitple access scheme used in most digital cellular telephone systems, such as European GSM systems and United States IS-136. TDMA techniques allocate a channel in the cell to each user sequentially in non-overlapping, equal interval, short duration of time (termed a time slot, which last hundreds of microseconds). A user transmits information during an assigned time slot once per frame.

The CDMA technique is an alternative digital modulation scheme, which assigns code sequences to each active user. All users then transmit their code-modulated data simultaneously using one carrier and the whole bandwidth of the cell. One disadvantage of TDMA is that it allows for only a fixed number of users at any time. If all the time slots have been assigned, no more users can enter the systems unless someone else leaves. This appears to be an advantage for CDMA, since there is no such hard limitation exists (theoretically). The addition of more

users simply causes a degradation in system performance. Another advantage which CDMA gives over TDMA for bursty channels is that valuable bandwidth resources are not wasted when no data is transmitted.

Other advantages of CDMA come from the spectral characteristics of transmitted signals. Since the spreading sequences are long pseudo-random sequences, users' transmitted signals are spread over the bandwidth which is much wider than the meaage bandwidth. The bandwidth of spread spectrum signals is wide enough so that the channel affects the sub-bands differently. Such channels are termed frequency selective channels. In wireless communication channels, the effects of scattering make the received signals contain multi-path components which are delayed and distorted versions of the original transmitted signal. While the multi-path phenomenon are often viewed as a problem to many communication systems, the frequency selective nature of the CDMA channel allows the detector to take advantage of them by resolving their individual signal components and providing a form of frequency diversity.

More advantages of CDMA [2] are capacity increasing over conventional analog cellular systems by an order of magnitude, supporting about three times the number of users per cell as a TDMA system and more reliable soft-handoff possibility. All these attractive features of DS-CDMA over TDMA prompt the development of CDMA cellular systems.

1.3 Conventional Detection

The base station is going to perform bit detection for each user. That means it is going to detect the desired user's signal which is contaminated by multiple access interference (MAI) due to the interfering users and background noise. Let's consider a simple two user system with desired user 1 and interfering user 2. The vector version of the received signal in a bit synchronous channel is:

$$\mathbf{r} = \sqrt{w_1}\mathbf{s}_1 b_1 + \sqrt{w_2}\mathbf{s}_2 b_2 + \mathbf{n}, \tag{1.1}$$

where $w_k, \mathbf{s}_k, b_k$ are the corresponding kth user's received energy, signature waveform (normalized) and information bits. The vector $\mathbf{n}$ is additive white Gaussian noise with zero mean and covariance matrix $\sigma^2\mathbf{I}$. A filter matched to the desired user's

signature waveform followed by a hard limiter make the what so called conventional detector. Let $< \cdot >$ denotes the standard inner product in Euclidean space. For the desired user one, the output of the hard limiter is:

$$\hat{b}_1 = \text{sgn}\{< \mathbf{r}, \mathbf{s}_1 >\}. \qquad (1.2)$$

If the signature waveforms of two users are orthogonal, which means the cross correlation $\rho \equiv < \mathbf{s}_1, \mathbf{s}_2 >$ equals to zero, then the conventional detector is the optimum (maximum likelihood) detector. If $|\rho| > 0$, the estimated information bit of the desired user 1 becomes

$$\hat{b}_1 = \text{sgn}\{\sqrt{w_1}b_1 + \sqrt{w_2}\rho b_2 + < \mathbf{s}_1, \mathbf{n} >\}. \qquad (1.3)$$

When ρ is small ($|\rho| \ll 1$) and the received amplitudes of two users are same, the conventional detector can work well. However, the detector performance can be degraded as the received power w_2 of the interfering user increases. This happens when all the users are transmitting at the same power level, and an interfering user is much closer to the base station (so called near-far problem). So the conventional detector rely on power control to equalize the received amplitudes of the users' signals.

1.4 Multiuser Detection Background

In order to solve near-far problem, Verdù proposed the optimum multiuser receiver or maximum likelihood (ML) receiver [3][4] which can be shown that by considering the structures of interfering user's signals, using a matched filter bank with a Viterbi algorithm, the minimum bit error rate can be achieved. This method set up the basis for multiuser detection. Unlike conventional detection, multiuser detection utilizes the knowledge of the received interfering users' information to detect the desired user's bits. The major advantage of multiuser detection is that it can provide significant performance gains over the conventional detector when the received amplitudes of all active users are similar and even under perfect power control. Hence in systems without perfect power control or no power control, multiuser detection is necessary to prevent the near-far problem; in systems with perfect

power control, multiuser detection is desirable to provide performance gains over the conventional detector.

Even though the optimum multiuser detector can minimize the probability of error, however, the computational complexity is exponential in the number of users. As a result of the computational complexity of this receiver, a number of suboptimal receivers were developed [5] [6] [7] [8] [9] [10] [11] (See [12] for a survey). These suboptimal multiuser detectors can effectively combat the near-far problem and provide significant performance gains with less computational complexity which is linear in the number of users.

Similar to zero-forcing equalizer which is used to combat ISI [13], a decorrelating detector uses a linear transformation which is basically the inverse of the cross-correlation matrix to zero out the other users' interference after the matched filter bank. The decorrelating detector has the advantage of being easy to implement and no estimation is needed for the received signals' energies. The decorrelating detector is optimally near-far resistant, meaning that its capability to reject multiple access interference (in terms of the near-far resistance performance measure) is equivalent to the optimum multiuser detector.

Linear decorrelating detector has the drawback of noise variance enhancement which limits its performance in situation where noise level is dominant over or comparable to MAI even though this sub-optimal method has the computational simplicity (linear in the number of users). Among linear sub-optimal multiuser detection, linear minimum mean squared error (LMMSE) estimation based multiuser detection choose the linear transformation $\mathbf{T}$ which minimize the Mean Squared Error (MSE) between the output data and the transmitted information. The MSE is given by $E[\|\mathbf{T}\mathbf{S}^T\mathbf{r} - \mathbf{b}\|^2]$ in the context of model 1.1, where the $N \times 1$ vector $\mathbf{s}_k$ is the kth user's signature sequence, $\mathbf{S} = [\mathbf{s}_1\mathbf{s}_2]$, $\mathbf{b} = [b_1 b_2]^T$. N is corresponding to the spreading length of the signature sequence. This receiver also maximizes the Signal to Interference plus Noise Ratio (SINR) among all linear receivers. The LMMSE receiver offers best trade-off between MAI cancellation and noise variance enhancement, while maintaining computational simplicity (linear in the number of users) and near-far resistance. Reference [12] is a tutorial on multiuser detection and reference [14] is a survey of this area.

Implementation of the preceding front end could be cumbersome, since it requires knowledge of the spreading waveforms and the propagation channels of all users, even if only one particular user is of interest. Furthermore, the complexity, both of the implementation of the front end and of the processing of the output, grows with the number of users. These difficulties can be overcome by using adaptive implementations of multiuser detection based on an alternative front end. Reference [15] derives a coherent, LMMSE detector which can be adaptively implemented with a training sequence. Its disadvantage is that whenever a sudden and significant change in the channel occurs (due to a severe fade or the arrival or departure of a strong interferer), it requires the retransmission of training sequences to reinitialize the detector. To overcome this disadvantage, references [16] and [17] show how the LMMSE multiuser detector can be adaptively implemented without training sequences; this blind adaptive multiuser detector requires only knowledge of the desired user's signature waveform and timing information, the same requirement as the conventional single-user matched filter.

1.4.1 Multi-shot Detection

Decorrelating and LMMSE multiuser detectors have received the most interest due to their good performance and simple mathematical formulation. Under the asynchronous conditions, decorrelating and LMMSE detectors are ideally infinite memory-length (refereed to as IIR) detectors. In an ideal implementation, the memory length equals to the number of users times the data packet length which often can be assumed to approach infinity. However, only the detector with a finite observation window length can be implemented. In order to achieve the performance of ideal (infinite memory length) detectors, the adaptive, decentralized, one-shot multiuser detector was proposed by [18][15][19][20][21][16]. Inside one-shot decorrelators, a matched filter is matched to one of the user's signature waveform while others are matched to the left and right parts of other users' signature waveforms. After the multi-shot matched filter bank, there is a decorrelator followed by a soft decoder which combines the three parts (left, central, right) of information together and decode them. While achieving inherently sub-optimal characteristics, one-shot decorrelating detector increases the dimension of the matched filter output vector

and cross correlation matrix, also it may require long adaptation time, and the adaptation must possibly be repeated frequently. Also, there exists the correlation matrix singularity problem for some relative delays between the users' codes. The other, termed multi-shot approach, was proposed in [22]. In this receiver, the input signals are processed by each filter at a time corresponding to the bit timing of each user. After the matched filter bank, a decorrelator which based on the matrix inversion introduced in reference [22] was followed. The cross-correlation matrix of the filters' output is non-singular in the multi-shot receiver, while it may become singular in one-shot approach. The performance of multi-shot decorrelating detector can be improved if the MAI due to edge symbols which are introduced by the truncation can be removed. Reference [22] introduced two signal processing methods into the multi-shot decorrelating detector and made the final multi-shot detector approximately attains infinite-memory-length decorrelating detector, which means the receiver can achieve the optimality in the sense of minimizing the MAI introduced by the edge symbols.

1.4.2 Multiuser Detection and Space-time Processing

The analog of the matched filter for multipath signals is the well known Rake filter [23][13] which coherently combines the outputs of filters matched to the time translates of a user's transmitted signal. The optimum multiuser detector for multipath signals [24] uses a bank of Rake filters in order to provide sufficient statistics. The Rake decorrelating detector [25][26] is a bank of Rake filters followed by a linear transformation based on the correlations of the users' multipath signals. The potential gains in using an array of sensors and spatial signal processing in conjunction with multiuser detection was recognized in [27], but the complexity is exponential in the number of users. A linear multiuser detector for vector channels derived in [28] decorrelates the users' signals using a linear combiner. It was called spatial-temporal decorrelator and is optimally near-far resistant when the users' spatial parameters (direction of arrival) are known.

1.5 Multi-rate Access Methodologies

A large variety of services is expected in future wireless networks such as the third generation cellular mobile systems. Users with very different, and even time-varying, rates and quality of service (QoS) requirements must be accommodated. A mobile terminal may set up and modify sessions for voice, data, images, as well as video through wireless connections to the base station. In order to provide such services, the network must be able to statistically multiplex users with different rates and/or QoS requirements while maximizing the spectral efficiency. Significant efforts are being made to integrate the cellular network with fixed networks for communicating both data and voice. CDMA provides natural implementations for a wireless communication systems that can provide services for a variety of information sources such as voice, video and data, which inherently have different data rates. Multi-rate CDMA systems have been investigated from different perspectives. Access methodologies and their comparison were proposed in [29][30]. They are:

- Fixed chip rate, variable spreading length (VSL) access scheme [29][31].

- Multi-code (MC) access scheme [32][33].

- Fixed processing gain, variable chip rate (VCR) access scheme [30][29].

- Multi-modulation access scheme [31], [34].

For VSL access method, the chip rate is constant and different data rates are accommodated with the assignment of signature sequences of different lengths. The signature sequence design (e.g. [35]) in the past has relied on codes in a set having the same length. The development of more versatile signature sequence sets is necessary for this proposed scheme. The advantage of the VSL access scheme is that only simple hardware for building a multi-rate CDMA radio interface is needed. However, in a conventional multi-rate CDMA system [36], this scheme suffers from performance degradation as the bit rates increase.

The MC access method can be implemented in single rate CDMA systems and thus offers the system construction advantage of being able to use the existing single rate receiver. High rate users are accommodated by multiplexing their data

onto several signature sequences. As a result, high rate data from a user is sent in parallel. MC schemes have been examined in references [32][33].

In the VCR access method, the chip rates of the signature sequences for the high rate and low rate users are different, but spreading lengths for different rate users are same. If we sample the signature sequences at the high chip rate, the discrete-time model is an equivalent VSL system. The additional constraint on the low rate signature sequences is that every M (rate ratio) chips form a group which take on the same value.

The multi-modulation access method utilizes an M-ary QAM modulation scheme and varies the modulation-level M to accommodate multiple bit rates. In the other word, users of different bit rates use different modulation schemes. Under the same signal to noise ratio per bit, transmitter powers will be different for different users. High rate users transmit at very high powers and in a conventional multi-rate CDMA systems, this will cause a severe near-far problem for relatively low rate users.

1.6 Multi-rate Multiuser Detection

With the pursuit of an ubiquitous wireless communications system that can provide wireless transport for a variety of information sources, it will be desirable to develop multiple bit rate systems. It is clear that voice, data and video have inherently different data rates. Multi-rate CDMA systems have been investigated from different perspectives. Access methodologies and their comparison were presented in [29][30]. The appropriate choice of the chip rate, chip pulse shape, processing gain, number of codes and modulation format were considered in [36][37][33][31][38]. The relationship between quality of services and the transmitted power and processing gains was studied in [39][40]. Multi-rate receiver design, however, is still a relatively new research area. Recently, some receivers originally proposed for single-rate systems were investigated for use in multi-rate CDMA schemes. Reference [36] proposed the conventional multi-rate receiver. Reference [41][42] proposed the decorrelating based multi-rate receiver. The further analysis of decorrelating based multi-rate receiver was studied by [43]. The performance of jointly optimal detection was studied in

[44]. Reference [45] analyzed the optimum near-far resistance of dual-rate CDMA signals for random signature sequences. A comparison of maximum likelihood-based detection for MC and VSL access schemes was proposed by [45]. Reference [34] proposed a successive interference cancellation method by using multi-modulation access scheme and reference [32][33] proposed the multi-code multi-rate receiver. Reference [46] proposed LMMSE receiver for multi-rate CDMA systems.

1.7   Thesis Overview

Due to the different access schemes described in section 1.5, it is patent that the choice of multi-rate access method will necessarily bias the choice of receiver. Thus the objective of our work is firstly, to investigate the receiver design for multi-rate CDMA communication systems for different access strategies and secondly, examine if one multi-rate access scheme is better than the other for a specific multi-rate receiver.

- **Chapter 2:** In order to provide the benchmarks for the performance of other detection methods, we study the maximum likelihood multi-rate receiver in this chapter. As the methods under study are near optimal, they can fulfill this benchmarking role. As the probability of error is intractable, the comparison will be based on the determination of asymptotic multiuser performance measure. The asymptotic multiuser efficiency (AME, denoted by η)[12] is an important performance measure for multiuser communication systems. It measures the exponential rate of convergence of its error probability to zero as the noise variance $\sigma^2 \to 0$, relative to the rate in a single user setting. The worst-case AME over all possible interference amplitudes is termed the near-far resistance (NFR) [12] and is denoted as $\bar{\eta}$. The AME is given for VSL maximum likelihood detection in this chapter.

- **Chapter 3:** the LMMSE multi-rate receiver is proposed in this chapter. The MC and VSL access schemes are considered. In the VSL system, a repetition code is used for low-rate user. It is noted that the optimum near-far resistance can be achieved by optimum receiver [3] and decorrelator [5]. This is also true for multi-rate CDMA systems. The optimum multi-rate receiver was

proposed by [44][45]. The decorrelating-based multi-rate receiver was proposed by [41], [42]. The optimum multi-rate receiver suffers from the same problem as the single rate optimum receiver which is the computational complexity (exponential in the number of users). The multi-rate decorrelating detector is optimally near-far resistance and easy to implement, but this receiver requires the knowledge of all the active users' signature waveforms. The proposed LMMSE multi-rate receiver minimizes the MSE between the decision statistic and the transmitted information bits. This receiver also maximizes the SINR and achieves the optimum near-far resistance among all linear receivers. The adaptive implementation of the LMMSE receiver makes it more attractive than the decorrelating receiver for multi-rate CDMA signals. The proposed LMMSE multi-rate receiver can be adaptively implemented only with the knowledge of the desired user. Due to the fact that LMMSE detector offers best trade-off between the MAI cancellation and noise variance enhancement, it is shown in this chapter that the multi-rate LMMSE receiver can offer better performance than the multi-rate decorrelator. In addition, we investigate the near-far resistance of this receiver for the different access schemes. It is shown that for both low-rate and high-rate users, the performance in the VSL scheme is better than that in the MC scheme.

- **Chapter 4:** We propose linear multi-shot receivers for asynchronous multi-rate CDMA system in this chapter. First proposed is the multi-shot decorrelating receiver for a dual-rate CDMA system. In this receiver, the desired user's information bits are decoded by first employing a set of modified correlators whose signals are time shifted versions of different rate users' signature waveforms. After the multi-shot matched filter bank, the filtered data for a finite length is put into the memory, then a decorrelating type detector is followed. Another receiver proposed in this chapter is multi-shot LMMSE receiver for dual-rate CDMA system. Instead of using decorrelating method, the LMMSE criterion is used in this receiver to minimize the MSE between the output data and the transmitted information bits. It is well known that a longer observation window or a longer FIR linear processor results in better performance, but also

greater computational complexity. The idea of debiasing which was suggested in [22] is introduced to the above two receivers. The performance of dual-rate receivers without debiasing improves monotonically with increasing the number of stacked bits. Debiasing optimizes dual-rate detectors based on the dual-rate multi-shot model. Debiasing multi-shot LMMSE detector for dual-rate signals can offer better performance than the corresponding debiasing multi-shot decorrelating detector. Also presented in this chapter is the performance comparison of different access methods.

- **Chapter 5:** We propose linear space-time multiuser receivers for multi-rate CDMA signals for multipath slow Rayleigh fading channels. As we all know, the Rake receiver is optimal for combating multi-path fading in the absence of MAI. Multiuser detection is optimal for suppressing the MAI. The space division multiple access techniques can provide antenna/or diversity gain by using multiple antennas at base station. By combining the time domain techniques, such as Rake receiver and multiuser detection for multi-rate signals with space-domain techniques, the resulting space-time receiver show potential of improving the capacity and performance for multi-rate CDMA systems. Firstly, we propose LMMSE space-time receiver for multi-rate CDMA systems. In this receiver, there are multiple antennas at the front end. After that, the code array/channel matched filters which correspond to low-rate users and high-rate users are followed. LMMSE processor which is working on the low-rate users' information bit interval is followed. The explicit knowledge of the channel and array coefficients is needed in the code array/channel matched filter part of this receiver. This receiver also introduces M bits processing delay for high rate users and a computational complexity which increases with the rate ratio M. Another type 2 space-time MMSE receiver is proposed. In the type 2 space-time receiver, the linear multiuser detection, array/channel weighting combining are combined together into one filter. In the adaptive implementation of type 2 space-time receiver, joint channel estimation and information detection can be implemented. The estimated channel information can be used for coherent combining. Secondly, a two-stage space-time LMMSE

receiver is proposed for multi-rate CDMA systems. This receiver generates decisions for high rate users at every high-rate bit interval of duration T_h. After every $(M-1)T_h$, the low rate users' decision is made based on the information provided by the first stage. The two-stage space-time LMMSE receiver reduces the computational complexity of the linear space-time multi-rate receiver and suppress the processing delay for high rate users with sacrificing a minor performance. We also propose the two-stage space-time receiver in the form of type 2 structure. at last, we compare the performance of one-stage and two-stage space-time receiver. We compare the performance of Maximum Ratio Combining (MRC) with the MMSE combining. We also compare the performance for the different access methods.

- **Chapter 6:** We propose group-wise interference cancellation methods for multi-rate CDMA signals in this chapter.The non-linear decision feedback detection (DFD) schemes are used in the proposed receivers. The proposed interference cancellation schemes benefit from the nature of the unequal received amplitudes for multi-rate CDMA signals. Users with same data rate are grouped together. Users with the highest data-rate are detected first. Interference between the groups is cancelled in a successive order. The results show that the group-wise MMSE DFD yields better performance than multi-rate LMMSE detector and multi-rate decorrelating detector, especially for highly loaded CDMA systems.

Most of the results in this thesis have been presented previously. Chapter 3 was originally presented in the 1997 International Conference of Information, Communications and Signal Processing (ICICS) [46]. Part of the work of chapter 4 was presented at the 1998 ICC [47] and other parts were presented at the 1998 CISS [48]. The results of dual-rate MMSE multipath receivers were presented in 1998 PIMRC [49]. The results of space-time dual-rate MMSE receivers were presented in 1999 CISS [50]. The different access methods comparison for space-time dual-rate receivers is going to be presented in 1999 Globecom. The results of group-wise MMSE interference cancellation schemes are submitted to ICC'2000.

CHAPTER 2

MAXIMUM LIKELIHOOD-BASED DETECTION FOR MULTI-RATE CDMA SIGNALS

2.1 Introduction

In order to provide benchmarks for performance measure of other detectors of later chapters, we study the Maximum Likelihood (ML) multiuser receiver for dual-rate CDMA signals in this chapter. Since the multi-rate access method will bias the choice of the receiver, the comparison of maximum likelihood joint detection for two multi-rate access schemes is presented. Two access methods to be examined are Multi-Code (MC) access method and Variable Spreading Length (VSL) access method. As the probability of error is intractable, the comparison will be based on the determination of asymptotic multiuser performance measures. The AME measures the exponential rate of convergence of its error probability to zero as the noise variance $\sigma^2 \to 0$, relative to the rate in a single user setting. The AME is denoted by η. The worst-case AME over all possible interference amplitudes is termed the near-far resistance (NFR) [12] and is denoted by $\bar{\eta}$. The AME is given for VSL maximum likelihood detection in this chapter.

This chapter is organized as follows: In section 2.2, two access methods are discussed, different performance measures and the relevant features of maximum-likelihood detection for CDMA systems are reviewed. In section 2.3, we introduce the dual-rate CDMA signal model which will be used throughout the dissertation. Also, the asymptotic multiuser efficiency of ML detector for VSL CDMA is given. In section 2.4, comparison of the performance achievable for two access methods is provided. In section 2.5, we provide analysis of the performance of ML receivers under different access schemes.

2.2 Preliminaries

We shall present results for a dual-rate CDMA system where a user can transmit information at one of two data rates. It is noted that the results of a dual-rate system can be easily generalized to accommodate the offering of more than two data rates.

VSL CDMA
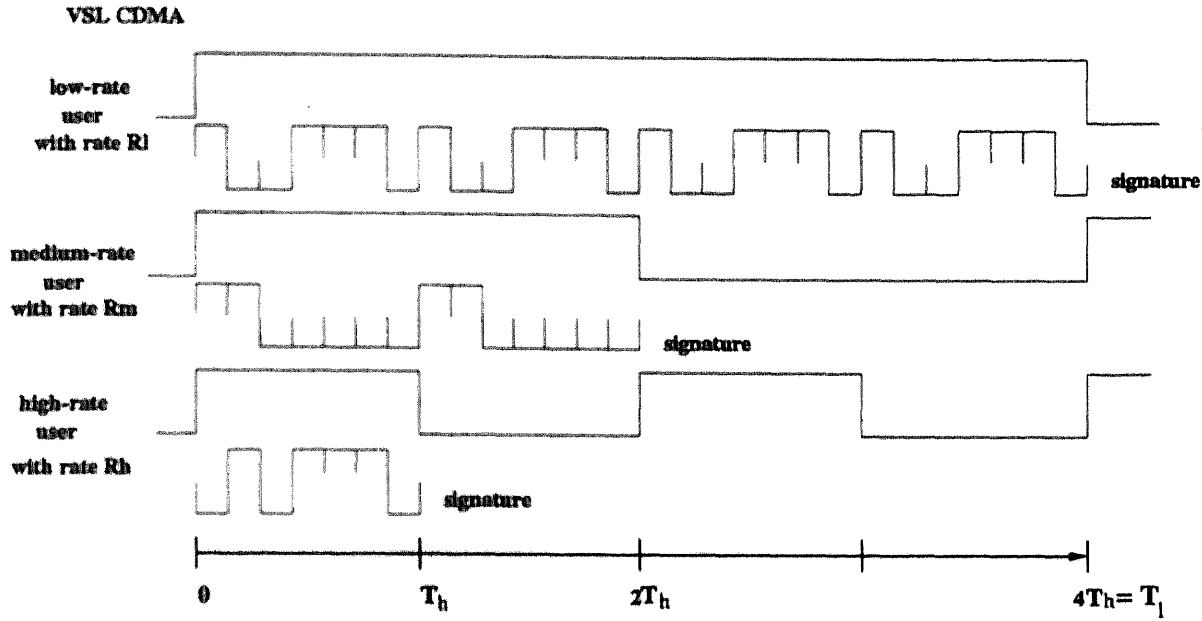
Figure 2.1 Variable spreading length multi-rate access technique

2.2.1 Multi-rate Access Methods

Two multi-rate access methods considered in this chapter are the multi-code access method and the variable spreading length access method. For variable spreading length access method, a constant chip rate is maintained and different data rates are accommodated with the assignment of signature sequences of different lengths. Figure 2.1 shows an example of the VSL access method with three users with different data rate ($R_h = 2R_m = 4R_l$). The multi-code access method can be implemented in single rate system and therefore offers the system construction advantage of being able to use existing single rate receivers. High rate users are accommodated by multiplexing their data onto several signature sequences. As a result, high rate data from a user is sent in parallel. Figure 2.2 shows an example of how three users with data rates $R_h = 2R_m = 4R_l$, would be supported via multi-code access. Essentially, the high rate user is converted into three virtual low rate users. Due to the assumption of constant chip rate for both access methods, the two access schemes can be constructed to have similar bandwidth requirements.
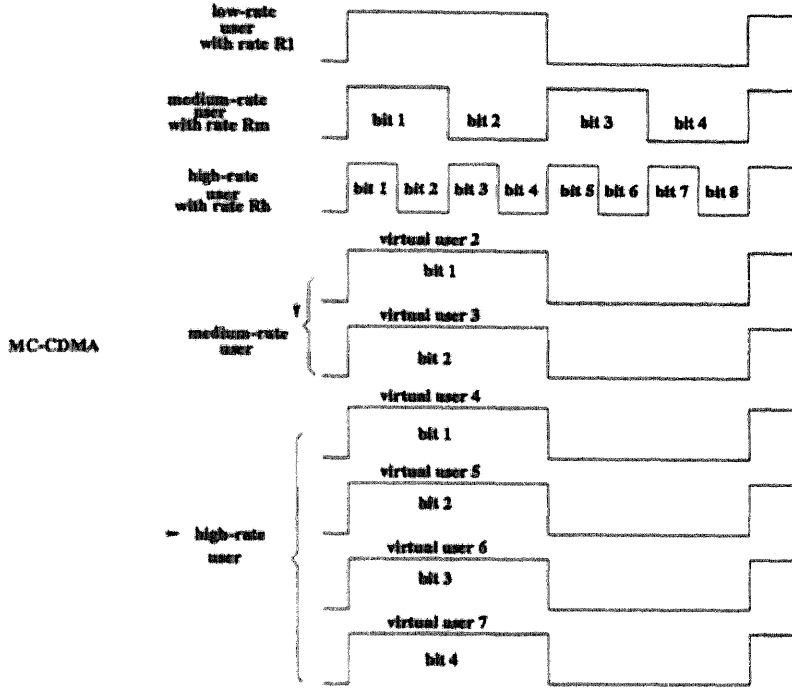
Figure 2.2 Multi-code multi-rate access technique

2.2.2 Performance Measure

In practice, the main performance measure of interest is the probability of bit error or bit error rate (BER). However, in some cases, the probability of bit error is intractable. In these cases the performance measure is going to be the asymptotic multi-user efficiency and near-far resistance (NFR). The AME describes the performance degradation of a receiver due to multiple-access interference (MAI) as the additive channel noise diminishes for a fixed set of received powers for all users. The worst-case AME over all possible received powers for the interfering users is the near-far resistance (NFR).

2.2.3 Review of Single Rate Maximum-likelihood Joint Detection

We begin with a single rate synchronous CDMA system. Assuming synchronous transmission channel with constant bit-rate and coherent detection. The received base-band information of K active users in a given symbol interval T can be described as:

$$r(t) = \sum_{k=1}^{K} \sqrt{w_k} b_k(t) s_k(t) + n(t) = s(t, b) + n(t), \tag{2.1}$$

where $w_k, b_k(t) \in \{-1, 1\}, s_k(t)$ are the received signal energy, transmitted information bits and normalized signature waveform of user k. The signal $n(t)$ is additive white Gaussian noise with zero mean and variance σ^2. The maximum likelihood joint detector optimizes the performance for all active users simultaneously while the minimum probability of error receiver provides optimal performance for a single active user. The optimum or maximum-likelihood decision on b which maximize the log-likelihood function [3] is $\hat{b}$. It can be written as:

$$\hat{b} = \arg \left\{ \max_{b \in \{-1,1\}^K} \left[2 \int_0^T r(t)s(t,b)dt - \int_0^T s(t,b)^2 dt \right] \right\}. \tag{2.2}$$

The vector form of estimated data bits $\hat{b} = [\hat{b}_1, \cdots, \hat{b}_K]^T$ which maximizes the likelihood function can be written as:

$$\hat{b} = \arg \max_{b_k \in \{\pm 1\}} 2b^T W y - b^T W R W b, \tag{2.3}$$

$$\begin{aligned} where \quad y \quad &= [y_1, \cdots, y_K]^T, \\ y_k \quad &= \int_0^T r(t)s_k(t)dt, \\ [R]_{i,j} \quad &= \int_0^T s_i(t)s_j(t)dt. \end{aligned}$$

Matrix W is a $K \times K$ diagonal matrix with $[W]_{k,k} = \sqrt{w_k}$, y_k is the matched filter output of kth user. The interested performance measure here is the asymptotic multiuser efficiency (AME). According to [51], the kth user's asymptotic efficiency of a detector is defined as

$$\eta_k = \sup \left\{ 0 \leq r \leq 1; \lim_{\sigma \to 0} \frac{P_k(\sigma)}{Q\left(\frac{\sqrt{w_k}}{\sigma}\right)} < +\infty \right\}, \tag{2.4}$$

where P_k is the bit error rate of kth user. In the absence of other users, the minimum bit error rate is equal to $Q(\sqrt{w_k}/\sigma)$. $Q(x)$ is the complementary cumulative distribution function of a zero-mean, unit-variance Gaussian random variable. The corresponding near-far resistance which is defined for each detector as its worst case asymptotic efficiency of user k over all possible energies for the interfering users is

$$\bar{\eta}_k = \inf_{w_j \geq 0, j \neq k} \eta_k. \tag{2.5}$$

According to the definition of the previous two equations, the asymptotic efficiency and near-far ratio of the maximum likelihood joint detection is [6][51]:

$$\eta_k^{ML} = \min_{\varepsilon \in \{0, \pm 1\}^K, \varepsilon_k = 1} \frac{1}{w_k} \varepsilon^T W R W \varepsilon, \tag{2.6}$$

$$\eta_k^{ML} = \frac{1}{(R^{-1})_{k,k}}. \tag{2.7}$$

After the reviewing of the maximum likelihood detection of single rate system, we are going to review the maximum likelihood detection of the multi-rate system.

2.3 Maximum Likelihood Detection for VSL Dual-rate CDMA Signals

We begin by making our discussion about maximum likelihood detection for VSL access due to the reason that VSL access merits the investigation of detection schemes which can take advantage of waveform structure.

2.3.1 Dual-rate Synchronous Signal Model

In a variable spreading length dual-rate CDMA system, we classify all the active users into two groups. One is called low-rate (LR) user group. Another is called high-rate (HR) user group. There are K_l low rate users and K_h high rate users. Bit interval of low-rate users is T_l. The bit interval of high-rate users is T_h. For each T_l, high-rate users transmit $M = T_l/T_h$ bits (M is supposed to be an integer). M is the rate ratio of two different signal rates R_l ($R_l = 1/T_l$) and R_h ($R_h = 1/T_h$). Chip rates for both high and low rate users in the VSL system are identical, and it equals to T_c. The processing gain for low-rate users equals to $N_l = T_l/T_c$ and the processing gain for high-rate users equals to $N_h = T_h/T_c$.

The baseband signal transmitted by low-rate user i over the time interval $[0, T_l)$ is:

$$\sqrt{w_i^{(l)}} b_i^{(l)} s_i^{(l)}(t).$$

The base-band equivalent signal transmitted by high-rate user j over the symbol interval $[(m-1)T_h, mT_h)$ is

$$\sqrt{w_j^{(h)}} b_{j,m}^{(h)} s_j^{(h)}(t - (m-1)T_h),$$

where

$w_i^{(l)} \equiv$ received energy of low-rate user i

$w_j^{(h)} \equiv$ received symbol energy of high-rate user j

$b_i^{(l)} \in \{-1, +1\}$ is the data symbol of low-rate user i

$b_{j,m}^{(h)} \in \{-1, +1\}$ is the data symbol of high-rate user j for mth subinterval of T_l

$s_i^{(l)}$ is the normalized signature waveform of low-rate user i, $\int_0^{T_l} s_i^{(l)^2}(t)dt = 1$

$s_j^{(h)}$ is the normalized signature waveform of high-rate user j

(where $(\cdot)^{(l)}$ and $(\cdot)^{(h)}$ denote low rate and high rate respectively and m is an integer which takes value from 1 to M).

In this chapter, we assume coherent detection with synchronous CDMA transmission channel with white Gaussian noise of zero mean and variance σ^2. The total received signal during the time interval $[0, T_l)$ form $K_v = K_l + MK_h$ virtual users and it can be written as:

$$r(t) = \sum_{i=1}^{K_l} \sqrt{w_i^{(l)}} b_i^{(l)} s_i^{(l)}(t) + \sum_{j=1}^{K_h} \left\{ \sum_{m=1}^{M} \sqrt{w_{j,m}^{(h)}} b_{j,m}^{(h)} s_j^{(h)}(t - (m-1)T_h) \right\} + n(t). \quad (2.8)$$

The noiseless chip rate sampled version of received information of K_l low-rate users during the mth subinterval of $[0, T_l)$ is the vector of size N_h:

$$\mathbf{r}_m^{(l)} = \mathbf{S}_m^{(l)} \mathbf{W}_m^{(l)} \mathbf{b}_m^{(l)}, \quad (2.9)$$

where

$$
\begin{aligned}
\mathbf{S}_m^{(l)} &= [\mathbf{s}_{1,m}^{(l)}, \cdots, \mathbf{s}_{K_l,m}^{(l)}], & N_h \times K_l \\
\mathbf{W}_m^{(l)} &= \operatorname{diag}\left\{ \sqrt{w_{1,m}^{(l)}}, \cdots, \sqrt{w_{K_l,m}^{(l)}} \right\}, & K_l \times K_l \\
\mathbf{b}_m^{(l)} &= [b_1^{(l)}, \cdots, b_{K_l}^{(l)}]^T, & K_l \times 1 \\
\mathbf{b}_1^{(l)} &= \cdots = \mathbf{b}_M^{(l)}.
\end{aligned}
$$

Suppose equal energy in each subinterval for low-rate users, then $\mathbf{w}_1^{(l)} = \cdots = \mathbf{w}_M^{(l)}$. Noting that when the repetition coding scheme is used for low-rate user i, which means the signature sequences repeat themselves M times during each low-rate bit interval, so the $N_l \times 1$ vector $\mathbf{s}_i^{(l)}$ equals to $[\mathbf{s}_{i,1}^{(l)^T}, \cdots, \mathbf{s}_{i,M}^{(l)^T}]^T$, where $s_{i,m}^{(l)}(n)$ corresponds to $s_i^{(l)}(n)$ for $(m-1)N_h \le n \le mN_h$, and $\mathbf{s}_{i,1}^{(l)} = \cdots = \mathbf{s}_{i,M}^{(l)}$, therefore, $\mathbf{S}_1^{(l)} = \cdots = \mathbf{S}_M^{(l)}$.

The noiseless chip rate sampled version of received information of K_h high-rate users during the mth subinterval of $[0, T_l)$ is a N_h vector

$$\mathbf{r}_m^{(h)} = \mathbf{S}_m^{(h)} \mathbf{W}_m^{(h)} \mathbf{b}_m^{(h)}, \quad (2.10)$$

where

$$
\begin{aligned}
\mathbf{S}_m^{(h)} &= [\mathbf{s}_{1,m}^{(h)}, \cdots, \mathbf{s}_{K_h,m}^{(h)}], & N_h \times K_h \\
\mathbf{W}_m^{(h)} &= \operatorname{diag}\left\{ \sqrt{w_1^{(h)}}, \cdots, \sqrt{w_{K_h}^{(h)}} \right\}, & K_h \times K_h \\
\mathbf{b}_m^{(h)} &= [b_{1,m}^{(h)}, \cdots, b_{K_h,m}^{(h)}]^T. & K_h \times 1
\end{aligned}
$$

$s_{j,m}^{(h)}$ is the vector form of $s_j^{(h)}(t - (m-1)T_h)$. The resulting noiseless vector form of received signal during the mth subinterval of $[0, T_l)$ is a N_h vector

$$\mathbf{r}_m = \mathbf{S}_m \mathbf{W}_m \mathbf{b}_m, \tag{2.11}$$

where

$$
\begin{aligned}
\mathbf{S}_m &= [\mathbf{S}_m^{(l)} \mathbf{S}_m^{(h)}], & N_h \times (K_l + K_h) \\
\mathbf{W}_m &= \mathrm{diag}\left\{\mathbf{W}_m^{(l)} \mathbf{W}_m^{(h)}\right\}, & (K_l + K_h) \times (K_l + K_h) \\
\mathbf{b}_m &= [\mathbf{b}_m^{(l)T} \mathbf{b}_m^{(h)T}]^T. & (K_l + K_h) \times 1
\end{aligned}
$$

The resulting vector form of received signal during $[0, T_l)$ is a N_l vector:

$$
\begin{aligned}
\mathbf{r} &= \mathbf{r}^{(l)} + \mathbf{r}^{(h)} + \mathbf{n} \tag{2.12} \\
&= [\mathbf{r}_1^{(l)T}, \cdots, \mathbf{r}_M^{(l)T}]^T + [\mathbf{r}_1^{(h)T}, \cdots, \mathbf{r}_M^{(h)T}]^T + \mathbf{n} \tag{2.13} \\
&= \mathbf{S}\mathbf{W}\mathbf{b} + \mathbf{n}, \tag{2.14}
\end{aligned}
$$

where

$$
\begin{aligned}
\mathbf{S} &\equiv [\mathbf{S}^{(l)} \mathbf{S}^{(h)}] & N_l \times (K_l + MK_h) \\
\mathbf{S}^{(l)} &= [\mathbf{s}_1^{(l)}, \cdots, \mathbf{s}_{K_l}^{(l)}] & N_l \times K_l \\
\mathbf{S}^{(h)} &= [\mathbf{s}_{1,1}^{(h)}, \cdots, \mathbf{s}_{K_h,1}^{(h)}, \cdots, \mathbf{s}_{1,M}^{(h)}, \cdots, \mathbf{s}_{K_h,M}^{(h)}] & N_l \times MK_h \\
\mathbf{W} &\equiv \mathrm{diag}\{\mathbf{W}^{(l)} \mathbf{W}^{(h)}\} & K_v \times K_v \\
\mathbf{W}^{(l)} &= \mathrm{diag}\left\{\sqrt{w_1^{(l)}}, \cdots, \sqrt{w_{K_l}^{(l)}}\right\} & K_l \times K_l \\
\mathbf{W}^{(h)} &= \mathrm{diag}\{\mathbf{W}_1^{(h)}, \cdots, \mathbf{W}_M^{(h)}\} & MK_h \times MK_h \\
\mathbf{b} &\equiv [\mathbf{b}^{(l)T} \mathbf{b}^{(h)T}]^T & K_v \times 1 \\
\mathbf{b}^{(l)} &= [b_1^{(l)}, \cdots, b_{K_l}^{(l)}]^T & K_l \times 1 \\
\mathbf{b}^{(h)} &= [\mathbf{b}_1^{(h)T}, \cdots, \mathbf{b}_M^{(h)T}]^T & MK_h \times 1 \\
\mathbf{n} &\sim \mathcal{N}(0, \sigma^2 \mathbf{I}_{N_l}).
\end{aligned}
$$

$\mathcal{N}(0, \sigma^2)$ stands for Gaussian normal distribution with zero mean and variance σ^2. $\mathbf{I}_{N_l}$ is the identity matrix of size N_l. $s_{j,m}^{(h)}$ is signature sequence ($N_l \times 1$) of the virtual high-rate user $(m-1)K_h + j, j = 1, \cdots, K_h, m = 1, \cdots, M$. In VSL system, $s_{j,m}^{(h)}$ only has non-zero support for the interval $[(m-1)N_h, mN_h)$, which means $\mathbf{S}^{(h)} = \mathrm{diag}\{\mathbf{s}_1^{(h)}, \cdots, \mathbf{s}_M^{(h)}\}$.

The vector $\mathbf{b}$ include the information bits which are transmitted by all $K_v = K_l + MK_h$ virtual users during the time interval $[0, T_l)$. The order of the vector $\mathbf{b}$ which is defined in the equation 2.14 is actually has the same form as:

$$\mathbf{b} = [b_1^{(l)}, \cdots, b_{K_l}^{(l)}, b_{1,1}^{(h)}, \cdots, b_{K_h,1}^{(h)}, \cdots, b_{1,M}^{(h)}, \cdots, b_{K_h,M}^{(h)}]^T. \tag{2.15}$$

This ordering defines an equivalent low rate system in which $K_l + MK_h$ bits are transmitted in $[0, T_l)$. In this equivalent virtual low-rate system, when the VSL access method is used, the data bits of high rate users in the mth subinterval is transmitted using a signature which is non-zero only in that interval.

2.3.2 Maximum Likelihood Detection

Based on the maximum likelihood detection of single rate CDMA system, we are going to review the maximum likelihood detection for dual-rate CDMA systems . The sampled output of a bank of correlators for optimal joint detection over a high rate bit interval $[(m-1)T_h, mT_h)$ is:

$$\mathbf{y}_m = \mathbf{S}_m^T \cdot (\mathbf{r}_m + \mathbf{n}_m) = \mathbf{R}_m \mathbf{W}_m \mathbf{b}_m + \mathbf{n}_{y,m}, \qquad (2.16)$$

where $\mathbf{n}_m \sim \mathcal{N}(0, \mathbf{I}_K)$. K is the number of active users and equals to $(K_l + K_h)$. The $K \times K$ cross-correlation matrix $\mathbf{R}_m$ can be written as:

$$
\begin{aligned}
\mathbf{R}_m &= \begin{bmatrix} \mathbf{S}_m^{(l)T} \\ \mathbf{S}_m^{(h)T} \end{bmatrix} \begin{bmatrix} \mathbf{S}_m^{(l)} & \mathbf{S}_m^{(h)} \end{bmatrix} \\
&= \begin{bmatrix} \mathbf{R}_{ll,m} & \mathbf{R}_{lh,m} \\ \mathbf{R}_{hl,m} & \mathbf{R}_{hh,m} \end{bmatrix}.
\end{aligned}
\qquad (2.17)
$$

The sampled output of a bank of correlators for optimal joint detection over low rate bit interval $[0, T_l)$ is:

$$\mathbf{y} = \mathbf{S}^T \cdot \mathbf{r} = \mathbf{R} \mathbf{W} \mathbf{b} + \mathbf{n}_y. \qquad (2.18)$$

Let's partition all the active users into two groups such that $\mathbf{S} = [\mathbf{S}_1 \mathbf{S}_2]$, then the cross-correlation matrix can be written as:

$$\mathbf{R} = \begin{bmatrix} \mathbf{R}_{11} & \mathbf{R}_{12} \\ \mathbf{R}_{21} & \mathbf{R}_{22} \end{bmatrix} \quad where \quad \mathbf{R}_{ij} = \mathbf{S}_i^T \mathbf{S}_j. \qquad (2.19)$$

If we consider the detection of users with non-zero support over $[(m-1)T_h, mT_h)$ then there will be a subset of users whose signals will be truncated or not present over this interval. Let $\mathbf{S}_2$ be the matrix whose columns are the spreading vectors for users with non-zero support on $[(m-1)T_h, mT_h)$. The signals actually employed to perform detection on this interval will be the result of projecting the codes in $\mathbf{S}_2$ onto the interval of interest. Thus the signature cross correlation matrix $\mathbf{R}_m$

of the active users with non-zero support on $[(m-1)T_h, mT_h)$ can be written as: $\mathbf{R}_m = \mathbf{S}_2^T \mathbf{P} \mathbf{S}_2$, where $\mathbf{P}$ is the appropriate projection. As $\mathbf{P}$ is a projection, this implies that $\mathbf{R}_{22} - \mathbf{R}_m \geq 0$.

Let's define $\eta_k^{(l)}$ to be the asymptotic multiuser efficiency (AME) of user k of the joint optimum detection over the low-rate bit interval and in this interval there are total $K_v = K_l + MK_h$ number of bits (virtual users) are transmitted. Similarly, let's define $\eta_k^{(h)}$ to be the AME of user k of the joint optimum detection over the high-rate interval and in this interval there are total $K = K_l + K_h$ bits transmitted.

$$
\begin{aligned}
if \quad D &= \begin{bmatrix} \mathbf{R}_{11} & \mathbf{R}_{12} \\ \mathbf{R}_{21} & \mathbf{R}_{22} - \mathbf{R}_m \end{bmatrix} \\
then \quad \eta_k^{(l)} &= \min_{\varepsilon \in \{0, \pm 1\}^{K_v}, \varepsilon_k = 1} \frac{1}{w_k^{(l)}} (\varepsilon^T \mathbf{W} \mathbf{R} \mathbf{W} \varepsilon) \\
&= \min_{\varepsilon \in \{0, \pm 1\}^{K_v}, \varepsilon_k = 1} \frac{1}{w_k^{(l)}} (\varepsilon^T \mathbf{W} \mathbf{D} \mathbf{W} \varepsilon + \varepsilon^{*T} \mathbf{w}_m \mathbf{R}_m \mathbf{w}_m \varepsilon^*) \\
(where \quad \varepsilon^* &= [\varepsilon(1), \cdots, \varepsilon(K)]^T) \\
&\geq \min_{\varepsilon \in \{0, \pm 1\}^{K_v}, \varepsilon_k = 1} \frac{1}{w_k^{(l)}} (\varepsilon^T \mathbf{W} \mathbf{D} \mathbf{W} \varepsilon) \\
&\quad + \min_{\varepsilon^* \in \{0, \pm 1\}^K, \varepsilon_k^* = 1} \frac{1}{w_k^{(l)}} (\varepsilon^{*T} \mathbf{w}_m \mathbf{R}_m \mathbf{w}_m \varepsilon^*) \\
&\geq \min_{\varepsilon^* \in \{0, \pm 1\}^K, \varepsilon_k^* = 1} \frac{1}{w_k^{(l)}} (\varepsilon^{*T} \mathbf{w}_m \mathbf{R}_m \mathbf{w}_m \varepsilon^*) \\
&= \eta_k^{(h)}
\end{aligned}
$$

Thus the conclusion we can draw is that for a multi-rate system using VSL access method, detection is preferably performed over the low-rate interval. If a detection of low-rate user is performed over high-rate symbol duration, then the AME is upper bounded by $\frac{1}{M}$ which means $\eta_k^{(h)} \leq \frac{1}{M}$. So the performance degradation for high-rate users detected over high-rate interval is not as severe as that for low-rate users.

2.4 Conclusion

Through the comparison of the asymptotic multiuser efficiency (AME) of maximum likelihood receiver for dual-rate CDMA signals, we can get that the AME on a high-rate bit interval, $[(m-1)T_h, mT_h)$, for those users with non-zero support on that interval is always worse than or equal to the AME on the low-rate interval. The AME of a low-rate user when the receiver is performed over the high-rate bit interval $[(m-1)T_h, mT_h)$ is upper-bounded by $\eta_k^{(h)} \leq \frac{1}{M}$.

CHAPTER 3

MINIMUM-MEAN SQUARED-ERROR (MMSE) ESTIMATION BASED DETECTION FOR MULTI-RATE CDMA SIGNALS

3.1 Introduction

In the previous chapter, the asymptotic multiuser efficiency (AME) of the maximum likelihood receiver for a dual-rate CDMA system is analyzed in order to provide benchmarks for the performance measure. In this chapter, we investigate the LMMSE receivers for multi-rate CDMA signals. The LMMSE receiver has the advantage of a simple linear structure. In comparison with dual-rate decorrelating detector, dual-rate LMMSE receiver can offer best trade-off between the MAI suppression and the noise variance enhancement. The other advantage of LMMSE receiver is that it can be implemented adaptively with only the knowledge of the desired user's signature sequence and timing information. The performance of two access schemes, VSL and MC, is compared in this chapter.

This chapter is organized as follows: In section 3.2, LMMSE receiver for dual-rate synchronous CDMA system is proposed. Section 3.3 proposes the high-rate mode LMMSE multi-rate receiver. Section 3.4 discusses the adaptive implementation issues for the LMMSE multi-rate receiver and section 3.5 provides the simulation results of the proposed receivers.

3.2 LMMSE Receiver for Dual-rate CDMA Signals

Let's consider a synchronous dual-rate CDMA system with rate-ratio between high-rate and low-rate is an integer M. The VSL access method is selected in this chapter for receiver design. Equation 2.8 of the previous chapter is a representation of the received signal in the dual rate system which is equivalent to a single rate system with $K_v = K_l + MK_h$ virtual low rate users each transmitting a single bit during $[0, T_l)$. Now we use this representation to describe the low rate LMMSE receiver. The chip rate sampled version of the received signal over the time interval $[0, T_l)$ has the same form as equation 2.14,

$$\mathbf{r} = \mathbf{SWb} + \mathbf{n}. \tag{3.1}$$

The received signals are first filtered with a bank of matched filters which matched to the different users' signature waveforms. The columnized properly ordered (K_l low-rate users followed by M-bits of K_h high-rate users) matrix notation of matched filter output over the low-rate interval has the same form of equation 2.18 and can be shown as:

$$\mathbf{y} = \mathbf{S}^T \cdot \mathbf{r} = \mathbf{RWb} + \mathbf{n}_y. \tag{3.2}$$

The matched filter output $\mathbf{y}$ is a K_v vector $\mathbf{y} = [\mathbf{y}^{(l)^T} \mathbf{y}_1^{(h)^T} \cdots \mathbf{y}_M^{(h)^T}]$ with

$$\mathbf{y}^{(l)} = \mathbf{S}^{(l)^T} \cdot \mathbf{r} \tag{3.3}$$

$$\mathbf{y}_m^{(h)} = \mathbf{S}_m^{(h)^T}(\mathbf{r}_m + \mathbf{n}_m), \tag{3.4}$$

where $\mathbf{n}_m \sim \mathcal{N}(0, \mathbf{I}_K)$. Figure 3.1 shows the system construction of the low-rate LMMSE receiver for dual-rate CDMA signals. The matched filter bank are supposed to have the order of K_l low-rate users' matched filter which are sampled at low-rate bit interval followed by K_h high-rate users' matched filters which are sampled at high-rate users' bit interval. The matrix $\mathbf{R}$ in equation 3.2 is the cross-correlation matrix between the different user's signature waveform. For the VSL access method, $\mathbf{R}$ has the size of $K_v \times K_v$ and can be written as

$$\mathbf{R} = \begin{bmatrix} \mathbf{R}_{ll} & \mathbf{R}_{lh,1} & \mathbf{R}_{lh,2} & \cdots & \mathbf{R}_{lh,M} \\ \mathbf{R}_{hl,1} & \mathbf{R}_{h,1} & 0 & \cdots & 0 \\ \mathbf{R}_{hl,2} & 0 & \mathbf{R}_{h,2} & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \mathbf{R}_{hl,M} & 0 & 0 & \cdots & \mathbf{R}_{h,M} \end{bmatrix},$$

where $\mathbf{R}_{ll}$ is the $K_l \times K_l$ cross-correlation matrix between K_l low rate users with $R_{ll}(i,j) = \int_0^{T_l} s_i^{(l)}(t) s_j^{(l)}(t)dt, (i, j = 1, \cdots, K_l)$; $\mathbf{R}_{lh,m}$ is the $K_l \times K_h$ cross-correlation matrix between K_l low rate users and K_h high rate users during the mth subinterval of T_l, $R_{lh,m}(i,j) = \int_{(m-1)T_h}^{mT_h} s_i^{(l)}(t) s_j^{(h)}(t - (m-1)T_h)dt, (i = 1, \cdots, K_l; j = 1, \cdots, K_h; m = 1, \cdots, M)$; $\mathbf{R}_{hl,m} = \mathbf{R}_{lh,m}^T$; $\mathbf{R}_{h,m}$ is a $K_h \times K_h$ cross-correlation matrix between high-rate users during the mth subinterval of T_l, $R_{h,m}(i,j) = \int_{(m-1)T_h}^{mT_h} s_i^{(h)}(t - (m-1)T_h)s_j^{(h)}(t - (m-1)T_h)dt, (i,j = 1, \cdots, K_h; m = 1, \cdots, M)$. Since for the VSL system, $s_{j,m}^{(h)}$, which represents the signature sequence of $((m-1)K_h + j)$th virtual high-rate user, has only non-zero support for $(m-1)N_h \leq N \leq mN_h$, this introduces

the orthogonality between the pairs of virtual high-rate users. Therefore, we have a sparse construction in the above cross-correlation matrix for VSL system. After the matched filter bank, the minimum mean squared error detection is used to cancel the MAI and minimize the mean-squared error between the output vector and the original information data vector. $\mathbf{b}$ is the vector of the transmitted information bits. $\mathbf{y}$ is the $(K_l + MK_h)$ vector which is the input to the linear transformation part. We need to find a $\mathbf{T}_{opt}$ such that

$$\mathbf{T}_{opt} = \arg\min_{T}\{E[\|\mathbf{Ty} - \mathbf{Wb}\|^2]\}. \tag{3.5}$$

Suppose $\Theta = \mathbf{Wb}$, we assume that all the components of the random multiuser information bits $\mathbf{b}$ are independent and identically distributed with zero mean and unit variance. The random vector Θ and noise vector $\mathbf{n}_y$ are statistically independent. The solution to equation 3.5 which can be found in [17][15] is:

$$\begin{aligned}
\hat{\Theta}_{MMSE} &= \mathbf{E}(\Theta) + \Sigma_{\Theta y}\Sigma_{yy}^{-1}(\mathbf{y} - \mathbf{E}(\mathbf{y})) \\
&= \Sigma_{\Theta y}\Sigma_{yy}^{-1}\mathbf{y} \\
&= (\mathbf{R} + \sigma^2\mathbf{W}^{-2})^{-1}\mathbf{y}, \tag{3.6}
\end{aligned}$$

where $\Sigma_{\Theta y} = \mathbf{W}^2\mathbf{R}, \Sigma_{yy} = \mathbf{RW}^2\mathbf{R} + \sigma^2\mathbf{R}$. So the solution of $\mathbf{T}_{opt}$ is going to be

$$\mathbf{T}_{opt} = (\mathbf{R} + \sigma^2\mathbf{W}^{-2})^{-1}. \tag{3.7}$$

The final decision is made by $\hat{\mathbf{b}} = \mathrm{sgn}(\hat{\Theta}_{MMSE})$ and the $\hat{\Theta}_{MMSE}$ can be get from

$$\hat{\Theta}_{MMSE} = \mathbf{T}_{opt}\mathbf{y} = \mathbf{T}_{opt}\mathbf{RWb} + \mathbf{n}_\Theta, \tag{3.8}$$

where $\mathbf{n}_\Theta$ is the Gaussian noise vector with zero mean and covariance matrix $\sigma^2\mathbf{T}_{opt}\mathbf{R}\mathbf{T}_{opt}^T$. In order to find the estimation error of $\hat{\Theta}_{MMSE}$, we can further decompose $\hat{\Theta}_{MMSE}$ as follows,

$$\begin{aligned}
\hat{\Theta}_{MMSE} &= \mathbf{T}_{opt}\mathbf{y} = (\mathbf{R} + \sigma^2\mathbf{W}^{-2})^{-1}\mathbf{y} \\
&= (\mathbf{R} + \sigma^2\mathbf{W}^{-2})^{-1} \cdot (\mathbf{RWb} + \mathbf{n}) \\
&= \Theta - \sigma^2\mathbf{T}_{opt}\mathbf{W}^{-1}\mathbf{b} + \mathbf{T}_{opt}\mathbf{n} \tag{3.9} \\
&= \Theta + \mathbf{e},
\end{aligned}$$

where the estimation error, $\mathbf{e} = -\sigma^2 \mathbf{T}_{opt} \mathbf{W}^{-1} \mathbf{b} + \mathbf{T}_{opt} \mathbf{n}$, contains both bias and noise components. The error covariance matrix of LMMSE can be found to be equal to

$$E\left\{(\Theta - \hat{\Theta})(\Theta - \hat{\Theta})^T\right\} = \text{cov}(\mathbf{e}) = \sigma^2 \mathbf{T}_{opt}. \tag{3.10}$$

The bit error rate for kth LR/HR user conditioned on the other users' bits is

$$P_k(\sigma) = Q\left(\frac{(\mathbf{T}_{opt} \mathbf{R} \mathbf{W} \mathbf{b})_k}{\sigma\sqrt{(\mathbf{T}_{opt} \mathbf{R} \mathbf{T}_{opt}^T)_{k,k}}}\right) \tag{3.11}$$

All the above results is built on the assumption that all the users' signature waveforms and the received amplitude are either known or estimated correctly. The advantage of the above multi-rate LMMSE method is that it can be implemented adaptively. The LMMSE adaptive implementation issue is going to be discussed in the later part. By using the data driven adaptive LMMSE detectors, only the knowledge of the desired user's signature is required in this receiver. Through the LMMSE method, one can also get the best trade-off between the MAI cancellation and noise variance enhancement.

3.3 High-rate Mode LMMSE Receiver for Dual-rate CDMA Signals

The objective of proposing another high-rate mode LMMSE receiver is that: the detector which is working over the maximum symbol interval has a large dimension and computational complexity and introduces $(M - 1)$ bits detection delay for high-rate users. In order to solve these problems, high-rate mode LMMSE receiver which is shown in Figure 3.2 is proposed in this part. In this high-rate mode LMMSE receiver, the matched filters are all set to work on the high-rate users' bit interval. This means that after one processing interval, low-rate users only get partial received information and high-rate users can get the whole received bit information. The delay which is introduced by the low-rate mode receiver for high-rate users is eliminated. In addition, the cross-correlation matrix has the smaller size of $(K_l + K_h) \times (K_l + K_h)$.

The matched filter output $\mathbf{y}_m$ (m is corresponding to the mth subinterval of the low-rate bit interval T_l) is a K (where $K = K_l + K_h$) vector and has the same form of equation 2.16 which is

$$\mathbf{y}_m = \mathbf{S}_m^T \cdot (\mathbf{r}_m + \mathbf{n}_m) = \mathbf{R}_m \mathbf{W}_m \mathbf{b}_m + \mathbf{n}_{y,m}, \tag{3.12}$$

and

$$\mathbf{R}_m = \begin{bmatrix} \mathbf{S}_m^{(l)^T} \\ \mathbf{S}_m^{(h)^T} \end{bmatrix} \begin{bmatrix} \mathbf{S}_m^{(l)} & \mathbf{S}_m^{(h)} \end{bmatrix}$$
$$= \begin{bmatrix} \mathbf{R}_{ll,m} & \mathbf{R}_{lh,m} \\ \mathbf{R}_{hl,m} & \mathbf{R}_{hh,m} \end{bmatrix}$$

where $\mathbf{R}_{ll,m}$ is a $K_l \times K_l$ cross-correlation matrix between K_l low rate users during mth subinterval of T_l, $R_{ll,m}(i,j) = \int_{(m-1)T_h}^{mT_h} s_i^{(l)}(t)s_j^{(l)}(t)dt, (i,j = 1, \cdots, K_l)$; $\mathbf{R}_{lh,m}$ is $K_l \times K_h$ cross-correlation matrix for mth subinterval of T_l of K_l low rate users with K_h high rate users , $R_{lh,m}(i,j) = \int_{(m-1)T_h}^{mT_h} s_i^{(l)}(t)s_j^{(h)}(t - (m-1)T_h)dt, (i = 1, \cdots, K_l; j = 1, \cdots, K_h; m = 1, \cdots, M)$; $\mathbf{R}_{hl,m} = \mathbf{R}_{lh,m}^T$; $\mathbf{R}_{hh,m}$ is the cross-correlation matrix between K_h high rate users for the mth subinterval of T_l, $R_{hh,m}(i,j) = \int_{(m-1)T_h}^{mT_h} s_i^{(h)}(t - (m-1)T_h)s_j^{(h)}(t - (m-1)T_h)dt, (i,j = 1, \cdots, K_h)$ and $\mathbf{n}_{y,m} \sim \mathcal{N}(0, \sigma^2 \mathbf{R}_m)$.

According to LMMSE criteria, the linear transformation for the mth subinterval can be found by:

$$\mathbf{T}_{opt,m} = \arg\min_{\mathbf{T}_m}\{E[\|\mathbf{T}_m\mathbf{y}_m - \mathbf{W}_m\mathbf{b}_m\|^2]\}$$
$$\mathbf{T}_{opt,m} = (\mathbf{R}_m + \sigma^2\mathbf{W}_m^{-2})^{-1}. \tag{3.13}$$

The decision statistic for the mth subinterval is:

$$\hat{\Theta}_{MMSE,m} = \mathbf{T}_{opt,m}\mathbf{y}_m = \mathbf{T}_{opt,m}\mathbf{R}_m\mathbf{W}_m\mathbf{b}_m + \mathbf{T}_{opt,m}\mathbf{n}_{y,m}. \tag{3.14}$$

For high-rate user, we can decode the information at every subinterval $[(m - 1)T_h, mT_h)$. The estimated information bit of kth high-rate user of mth subinterval is:

$$\hat{b}_{k,m}^{(h)} = \text{sgn}\{(\hat{\Theta}_{MMSE,m})_{K_l+k}\}. \tag{3.15}$$

The probability that high-rate user k is decoded incorrectly for the mth subinterval conditioning on the other user's information bits equals to:

$$P_{k,m}^{(h)}(\sigma) = Q\left(\frac{(\mathbf{T}_{opt,m}\mathbf{R}_m\mathbf{W}_m\mathbf{b}_m)_{K_l+k}}{\sigma\sqrt{(\mathbf{T}_{opt,m}\mathbf{R}_m\mathbf{T}_{opt,m}^T)_{K_l+k,K_l+k}}}\right) \tag{3.16}$$

For low-rate users, only the partial information can be obtained during every subinterval. A soft decoding rule is going to be applied for low-rate users. The soft

decoding rule which is used here is the maximum ratio combining. The estimated information of kth low-rate user is:

$$\hat{b}_k^{(l)} = \text{sgn}\left\{\left(\sum_{m=1}^{M}\frac{1}{(\mathbf{T}_{opt,m}\mathbf{R}_m\mathbf{T}_{opt,m}^T)_{k,k}}(\mathbf{T}_{opt,m}\mathbf{y}_m)_k\right)\right\}. \qquad (3.17)$$

The probability that the kth low-rate user is detected incorrectly conditioned on the other users' information bits is:

$$P_k^{(l)}(\sigma) = Q\left(\frac{\sum_{m=1}^{M}(\mathbf{T}_{opt,m}\mathbf{R}_m\mathbf{W}_m\mathbf{b}_m)_k}{\sigma\sqrt{\sum_{m=1}^{M}((\mathbf{T}_{opt,m}\mathbf{R}_m\mathbf{T}_{opt,m}^T)_{k,k})}}\right) \qquad (3.18)$$

3.4 Multi-rate LMMSE Receiver-Adaptive Implementation

Dual rate decorrelating detectors that working on the matched filter outputs $\mathbf{y}$ (see equation 2.18) is proposed by [41] and is called low-rate decorrelating detector. Reference [41] also proposed another high-rate decorrelating detector which is working on the matched filter output $\mathbf{y}_m$. The analysis of this two decorrelating detectors can be found in reference [41][52]. In order to implement the dual-rate decorrelating detectors, the receiver need the knowledge of all the active users' signature sequences. This obviously limits the application of dual-rate decorrelating detectors in many areas. Motivated by the fact that the data-driven adaptive LMMSE detector can successfully decode the desired user's information bit out of single rate CDMA data [17] with the knowledge of only the desired user's signature, we implement the dual-rate LMMSE receivers adaptively using the method in reference [17]. The dual-rate LMMSE detectors essentially apply the LMMSE idea right on the chip-rate sampled data $\mathbf{r}$ or $\mathbf{r}_m$, according to low-rate LMMSE receiver or high-rate LMMSE receiver. When we consider the low-rate LMMSE receiver, the processor is working on the low-rate users' bit interval T_l. The low-rate LMMSE receiver decodes a desired ith low-rate user's information or a desired virtual $((m-1)K_h + j)$th high-rate user's bit information as follow,

$$ith \qquad low-rate \quad user \quad \hat{b}_i^{(l)} = \text{sgn}\left\{\mathbf{s}_i^{(l)^T}\hat{\Sigma}_{rr}^{-1}\mathbf{r}\right\},$$

$$((m-1)K_h+j)th \quad high-rate \quad user \quad \hat{b}_{j,m}^{(h)} = \text{sgn}\left\{\mathbf{s}_{j,m}^{(h)^T}\hat{\Sigma}_{rr}^{-1}\mathbf{r}\right\},$$

where $\hat{\Sigma}_{rr}$ is the estimated covariance matrix of the received signal and $\hat{\Sigma}_{rr}^{-1}\mathbf{r}$ can be sequentially estimated and updated from data sequence $\{\mathbf{r}(n)\}$ [17]. Vector $\mathbf{s}_i^{(l)}$

is simply the chip-rate sampled signature of the ith low-rate user. Vector $\mathbf{s}_{j,m}^{(h)}$ is chip-rate sampled signature of virtual high-rate user $((m-1)K_h + j)$. In the VSL system, $\mathbf{s}_{j,m}^{(h)}$ is simply a vector with its non-zero components formed from the chip-rate sampled signature of jth high-rate user for $(m-1)N_h \leq N \leq mN_h$. The important features of the adaptive low-rate LMMSE receiver are

- Only the signature of the desired user is needed in decoding its information bits.

- LMMSE filters in this adaptive implementation scheme are simply generalized matched filters [17], one can output M-bits of desired high-rate users' information sequentially from same LMMSE filter, as long as $\hat{\Sigma}_{\mathbf{rr}}^{-1}$ and $\mathbf{r}$ are available.

The dimension of data for low-rate LMMSE detector is proportional to the processing interval T_l. The high-rate mode LMMSE receiver has the lower dimension corresponding to the processing interval of T_h which equals to T_l/M. The high-rate mode LMMSE receiver decodes a desired ith low-rate user's partial information or jth high-rate user's information from data $\mathbf{r}_m$ over the interval of $[(m-1)T_h, mT_h)$, $(m = 1, 2, \cdots, M)$.

$$ith \quad low-rate \quad user \quad \hat{b}_{i,m}^{(l)} = \text{sgn}\left\{\mathbf{s}_{i,m}^{(l)}{}^T \hat{\Sigma}_{\mathbf{r}_m\mathbf{r}_m}^{-1} \mathbf{r}_m\right\},$$

$$jth \quad high-rate \quad user \quad \hat{b}_j^{(h)} = \text{sgn}\left\{\mathbf{s}_j^{(h)T} \hat{\Sigma}_{\mathbf{r}_m\mathbf{r}_m}^{-1} \mathbf{r}_m\right\},$$

where $\hat{\Sigma}_{\mathbf{r}_m\mathbf{r}_m}^{-1}$ is of less dimension (proportional to T_h) and can be obtained from data sequence $\{\mathbf{r}_m(n)\}$. $\mathbf{s}_j^{(h)}$ is simply the chip-rate sampled signature of the jth high-rate user. $\mathbf{s}_{i,m}^{(l)}$ is only the mth segment of chip-rate sampled signature of ith low-rate user. Since low-rate users' information are embedded within a longer symbol interval $T_l = MT_h$, using partial data $\mathbf{r}_m$ within sub-interval to decode low-rate users' information causes performance degradation for low-rate users. In order to alleviate this problem, we propose to use signal to interference-plus-noise ratio (SINR) of desired low-rate user j with each subinterval to weight partial results to make a better decision on the jth low-rate user's information bit as follows,

$$\hat{b}_j^{(l)} = \text{sgn}\left\{\sum_{m=1}^{M} \text{SINR}_j \mathbf{s}_{j,m}^{(l)}{}^T \hat{\Sigma}_{\mathbf{r}_m\mathbf{r}_m}^{-1} \mathbf{r}_m\right\}, \tag{3.19}$$

where SINR$_j$ can be calculated from the formula derived in [17] by noticing that models in 3.12 should be used. When the repetition coding scheme is used for low-rate users, the signature sequences of all users in every subinterval is same, which means $\mathbf{S}_m$ and $\mathbf{R}_m$ are no longer functions of index m. In this case, maximum ratio combining reduces into a simple equal gain combining, and equation 3.19 can be simplified as

$$\hat{b}_j^{(l)} = \text{sgn}\left\{ \left(\sum_{m=1}^{M} \mathbf{s}_{j,m}^{(l)\,T} \hat{\Sigma}_{\mathbf{r}_m \mathbf{r}_m}^{-1} \cdot \mathbf{r}_m \right) \right\}. \tag{3.20}$$

Each virtual user's signature sequence has length $MN_h = N_l$ and is normalized over the interval of interest. In MC method, $\mathbf{s}_{i,m}^{(h)} \neq 0$ and $\mathbf{s}_j^{(l)} \neq 0$ for every $1 \leq N \leq MN_h$. In VSL method, $\mathbf{s}_j^{(l)} \neq 0 \forall N$, but $\mathbf{s}_{i,m}^{(h)} \neq 0$ only for $(m-1)N_h \leq N \leq mN_h$.

3.5 Simulation Results and Discussions

In chapter 2, we show that for maximum likelihood detectors, the AME over the high-rate bit interval, for those users with non-zero support on that interval in a low-rate system is always worse than or equal to the AME over the low-rate interval. It was proved by reference [41] for the decorrelating detector that the AME of the low-rate user employing repetition coding over the low-rate bit interval is same as the AME over the high-rate bit interval, and for high-rate users, the AME over the low-rate bit interval is always better than the AME over the high-rate bit interval.

In this section, the performance of LMMSE receiver working over the low-rate bit interval will be evaluated numerically and compared with the LMMSE receiver working over the high-rate bit interval. In the mean time, the performance of LMMSE receivers working over the different bit interval will be compared with the corresponding decorrelating receivers. The performance comparison of LMMSE receivers and decorrelating detectors between VSL system and MC system is also provided in this section.

A direct sequence CDMA system is considered. In this system, there are two low-rate users and one high-rate user. This means $K_l = 2, K_h = 1$. The rate-ratio is set to be equal to two which means $M = 2$. For the VSL system, the processing gain for high-rate users is set to be $L_h = 7$, the processing gain of low-rate users is

set to be $L_l = 2 \times 7 = 14$. The repetition coding scheme is used for low-rate users which means that the code repeat itself M times during one low-rate bit interval. The Gold sequences are chosen to be the spreading sequences. The SNR of the kth user will be $10 \log \frac{w_k}{N_0/2}$.

In Fig. 3.3, the Bit Error Rate (BER) of low-rate user 1 is plotted as a function of the increasing SNR of the other users. As expected, LMMSE dual-rate receiver can attain single user lower bound when the other users' SNR is low, and when the other users' SNR is high, it simply becomes a decorrelating detector. In this figure, we can see for low-rate user 1 which using repetition code, the performance of low-rate mode decorrelating detector is same as the performance of high-rate decorrelating detector. But for LMMSE receiver, the near-far resistance of low-rate user 1 when it working over the low-rate mode may not equal to the near-far resistance of the high-rate mode LMMSE receiver, which means the near-far resistance of low-rate LMMSE receiver for low-rate users is greater than or equal to the near-far resistance of high-rate mode LMMSE receiver. In Fig. 3.4, the same results for the high-rate user 3 are provided. Here, we can see the near-far resistance of the high-rate user for the high-rate LMMSE receiver is always worse than the near-far resistance of the low-rate LMMSE receiver at the high SNR region of the interferers'. That means for high-rate users, there is always a performance degradation when its information is detected through high-rate LMMSE receiver. Fig. 3.5 show the performance comparison of low-rate user between the LMMSE receiver with VSL access method and MC access method. There are two disadvantages for MC access method. Firstly, no orthogonality exists in the pairs of virtual high-rate users. Secondly, when the code is not totally orthogonal, MC access method introduces multiple access interference or inter symbol interference every M bits for each high-rate user itself through the serial to parallel transition and the code assignment. For the Gold sequences which we are using, the VSL system outperforms the MC system. Fig. 3.6 shows us the performance comparison of high-rate user between the VSL system and MC system. It also shows that the VSL system outer-performs the MC system. Another performance comparison we want to make in the simulation results of this chapter between the VSL and MC system is the influence of the rate ratio M over the bit error rate. Let's consider a five users VSL system with $K_l = 4$ low-rate user and
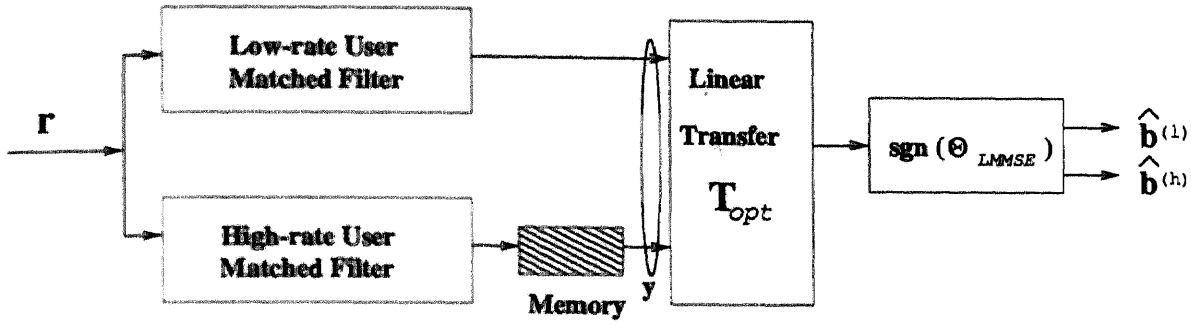
Figure 3.1 Low-rate LMMSE receiver for VSL dual-rate CDMA system
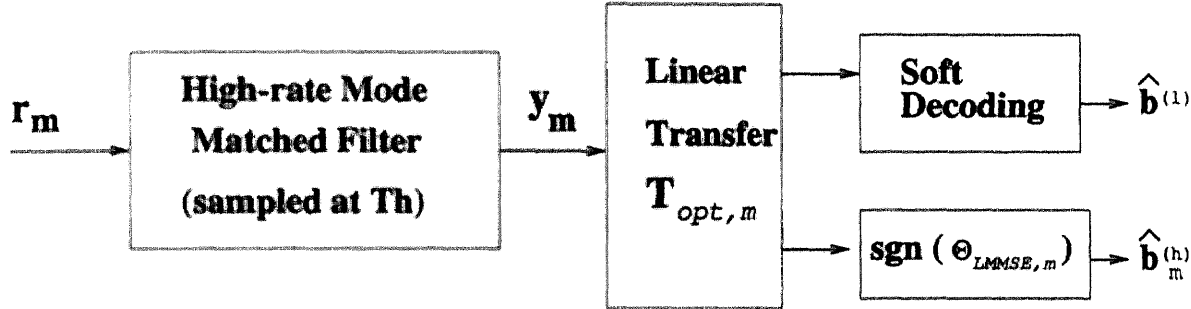
Figure 3.2 High-rate mode LMMSE receiver for dual-rate CDMA system

$K_h = 1$ high-rate user. The chip rate is selected to be consistent with the system's bandwidth for both access methods, where low-rate users still use repetition coding. Fig. 3.7 shows us the results of BER versus rate ratio M of the VSL virtual low-rate system. We observed that for both low-rate decorrelating detector and low-rate LMMSE receiver, the performance of low-rate user is insensitive to the rate-ratio M. But the performance of high-rate user becomes better as M increases. Fig. 3.8 show us the BER versus rate-ratio M for MC system. It can be observed here that both low-rate user and high-rate users' performance for different receivers is going to deteriorate as M increases.
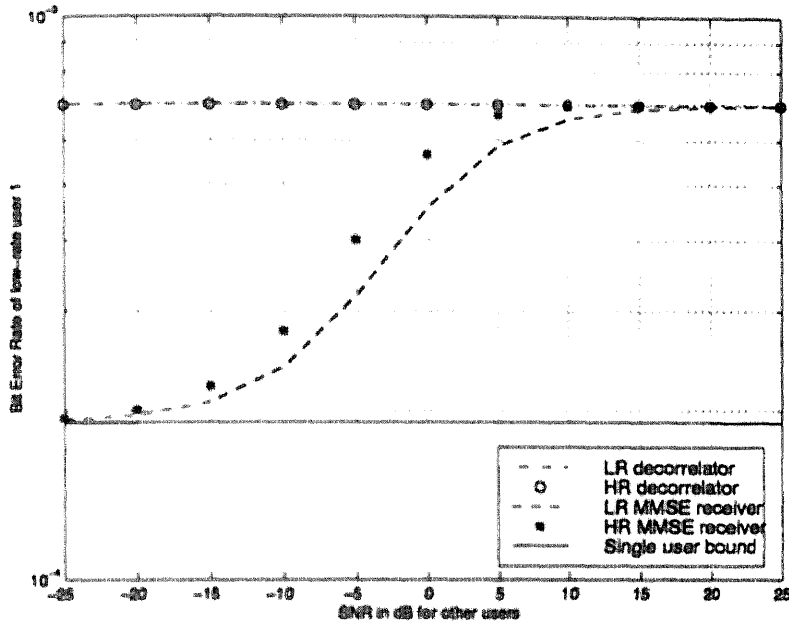
Figure 3.3 Three user VSL system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$, SNR of low-rate user 1 $SNR1 = 8dB$
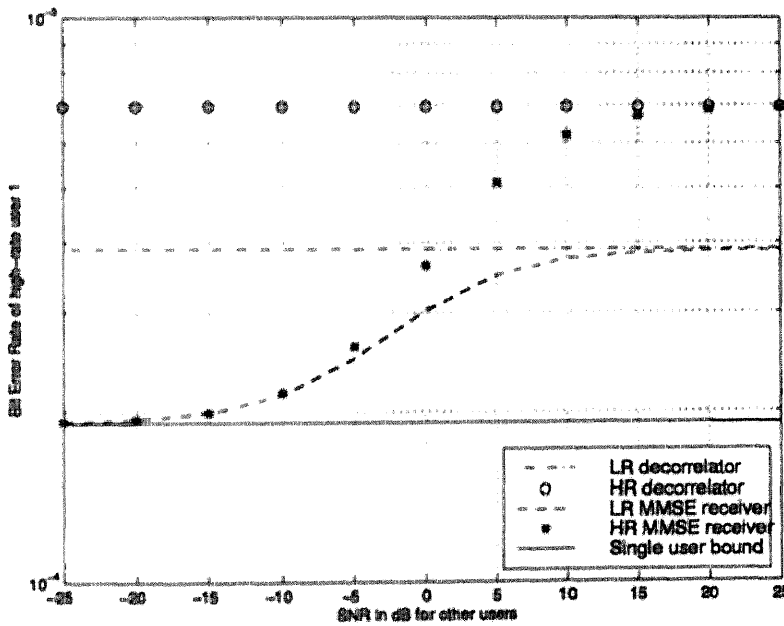
Figure 3.4 Three user VSL system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$, SNR of high-rate user 1: $SNR3 = 8dB$
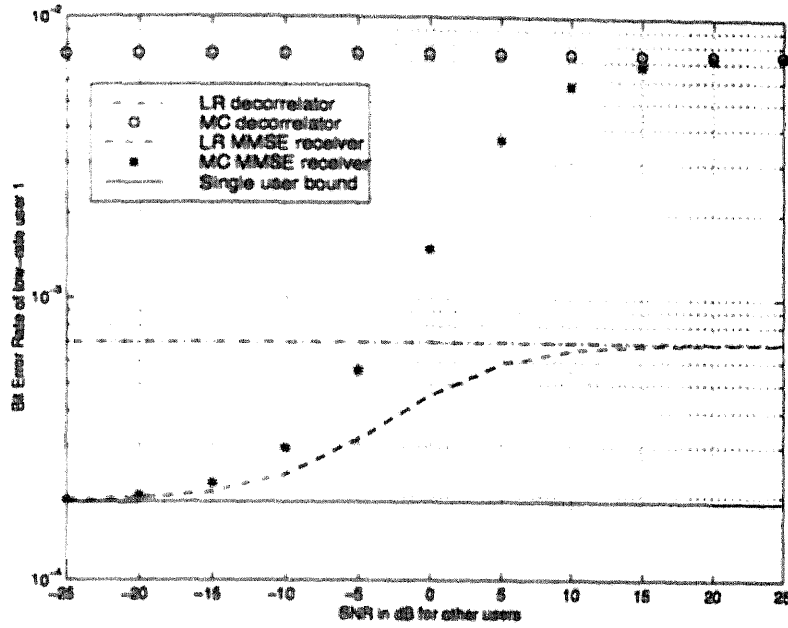
Figure 3.5 Three user system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$ for VSL system, $L = 14$ for MC system. SNR of low-rate user 1: $SNR1 = 8dB$
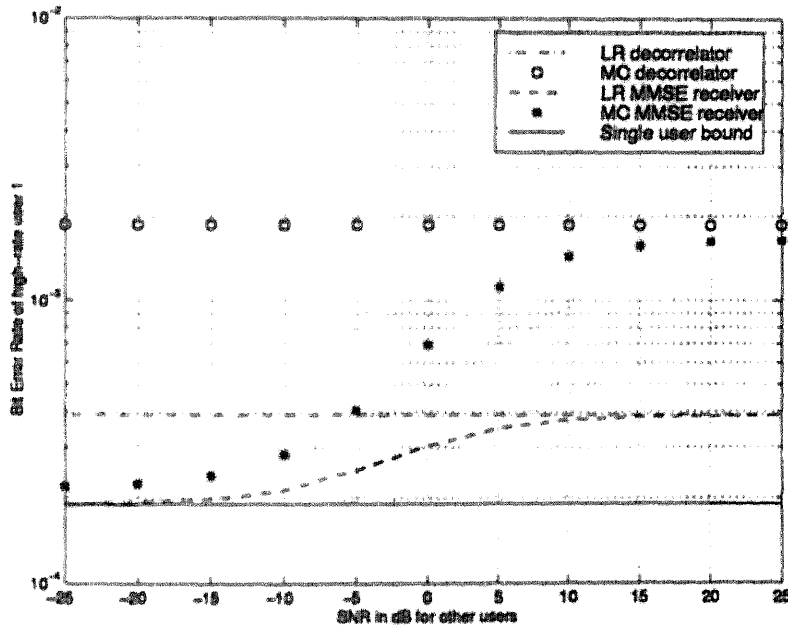
Figure 3.6 Three user system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$ for VSL system, $L = 14$ for MC system. SNR of high-rate user 1: $SNR3 = 8dB$
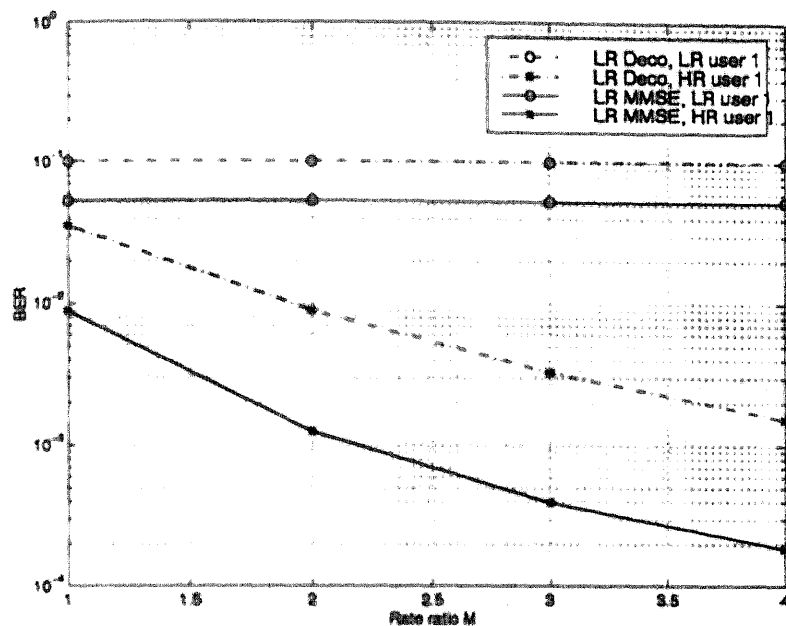
Figure 3.7 Five user VSL virtual low-rate system, $K_l = 4, K_h = 1, SNR = 10dB$ for all users, $L_l = 14, L_h = 7$
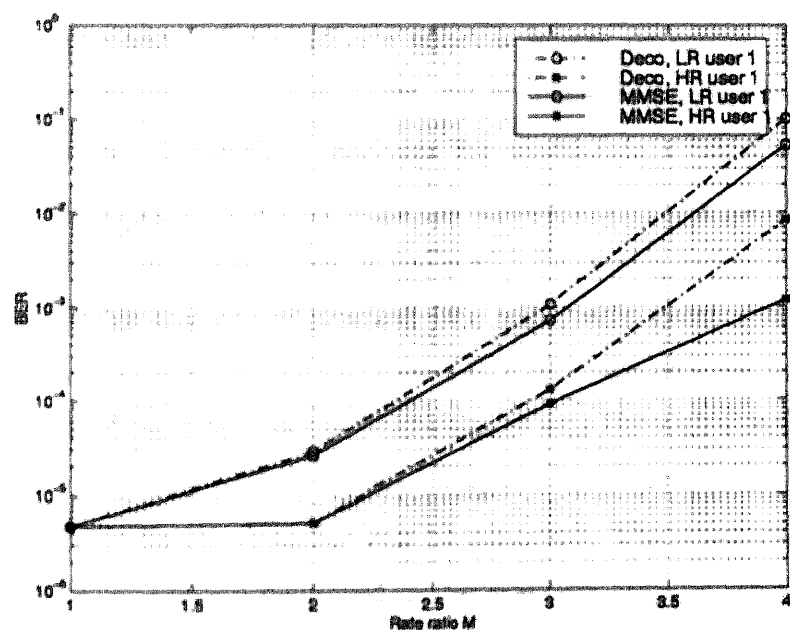
Figure 3.8 Two user MC virtual low-rate system, $K_l = 1, K_h = 1, SNR = 10dB$ for all users, $L = 14$

CHAPTER 4

MULTI-SHOT RECEIVER FOR ASYNCHRONOUS MULTI-RATE CDMA SYSTEMS

4.1 Introduction

The previous two chapters consider the receiver design for dual-rate synchronous CDMA transmission channels. In this chapter, we discuss the multi-rate receiver design for asynchronous CDMA transmission channel. The asynchronous decorrelating detector for single rate CDMA systems was proposed by [6]. LMMSE receiver for single rate asynchronous CDMA systems was proposed by [15]. Decorrelating detector and LMMSE receiver achieve the optimum near-far resistance [5] [6] by extending the observation window length to infinity. They are ideally infinite memory-length detectors. To obtain practical detectors which have low implementation complexity, linear, finite-memory-length multiuser detectors were studied by [53]. The one-shot matched filter and multi-shot matched filter in reference [54] are shown in Figure 4.1 and Figure 4.2. A multi-shot approach with debiasing method was proposed by [22]. References [53] [22] showed that a decorrelating-based detector with moderate observation window length can asymptotically achieves the performance of the infinite window length detectors. Based on dual-rate model of synchronous CDMA systems, we propose dual-rate model for asynchronous CDMA systems. Multishot receivers for multi-rate CDMA signals are proposed in this chapter. It is shown through the simulation results that the performance of multishot receiver improves monotonically with increasing the number of stacked bits. However, the large number of stacked bits will yields great computational complexity. The debiasing method which is used in multi-rate receiver for asynchronous transmission channel can improve the performance only through a moderate number of staked bits, which means without great computational complexity, the performance of infinite memory length detector can be achieved. The adaptive implementation scheme for LMMSE method proposed in the previous chapter can handle both synchronous and asynchronous channels as long as the signature and timing information on the desired user is available.

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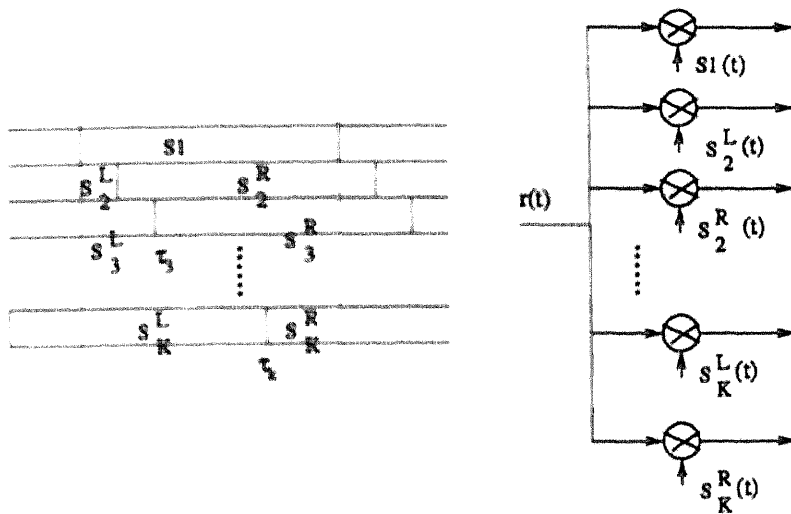
Figure 4.1 One-Shot matched filter timing and structure
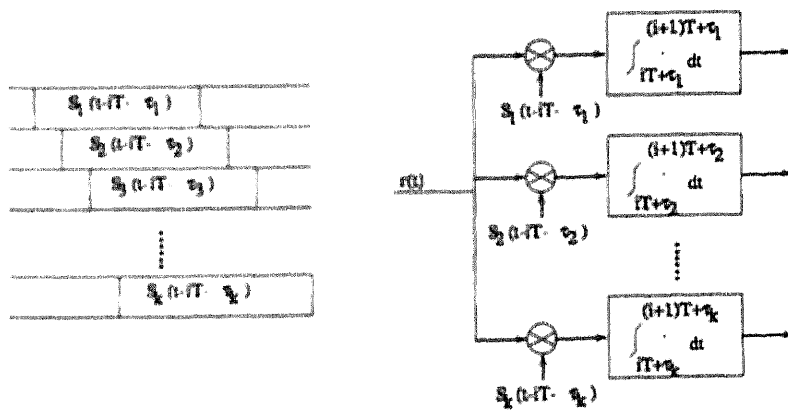
Figure 4.2 Multi-Shot matched filter timing and structure

4.2 Preliminaries

Let's consider our dual-rate system as a virtual low-rate system that transmit the signals asynchronously. Similar to the synchronous dual-rate system in chapter 2, the active users are still classified into two groups. Low-rate users with bit interval T_l, high-rate users with bit interval T_h. The bit rate ratio is equal to $M = \frac{T_l}{T_h}$, which is set to be an integer. The signature waveforms of different users are normalized to the bit interval of interest. There are total $K_v = K_l + M K_h$ virtual users in a virtual low-rate system where K_l, K_h stand for the number of low-rate and high-rate users respectively. Coherent detection is assumed in this chapter and the signals are sent over an additive white Gaussian noise channel in which the noise $n(t)$ has zero mean and variance $\sigma^2 = \frac{N_0}{2}$. The base-band received signal for virtual low-rate system during the time interval $[0, T_l)$ is:

$$
\begin{aligned}
r(t) \;=\; & \sum_{i=1}^{K_l} \sqrt{w_i^{(l)}} b_i^{(l)} s_i^{(l)}(t - \tau_i) \\
& + \sum_{j=1}^{K_h} \sum_{m=0}^{M-1} \sqrt{w_{j,m}^{(h)}} b_{j,m}^{(h)} s_{j,m}^{(h)}(t - mT_h - \tau_j) + n(t),
\end{aligned}
\tag{4.1}
$$

where superscripts (l) and (h) denote low-rate user and high-rate user. $b_i^{(l)}, w_i^{(l)}, s_i^{(l)}$ are the corresponding transmitted information bits, received energy and signature sequences of low-rate user i, where $i = 1, \cdots, K_l$. $b_{j,m}^{(h)}, w_{j,m}^{(h)}, s_{j,m}^{(h)}$ are the corresponding transmitted information bits, received energy and signature sequences of virtual high-rate user $K_l + (m-1)K_h + j$, where $m = 1, \cdots, M; j = 1, \cdots, K_h$. τ_i, τ_j represent transmission delay for low-rate user i and high-rate user j. We assume the receivers know the time delay τ for each user. If the desired user is user k, without lost of generality, we will assume $\tau_k = 0$ and the other τ's are computed relative to $\tau_k = 0$. When all the τ's are equal, the channel is a synchronous transmission channel. Under the synchronous condition, the output of a bank of matched filters for single rate system forms the sufficient statistics of all the users' information bit within one symbol interval. When the delay τ's are not equal, the information of different users is not transmitted at the same time and make the received signals to be an asynchronous form. In general, we assume that delays of all $K_l + K_h$ users satisfy the relationship of $0 \le \tau_1 \le \cdots \le \tau_{K_l} \le \tau_{K_l+1} \le \cdots \le \tau_{K_l+K_h} < T_h$. It is noted that between the pairs of virtual users in the VSL virtual low-rate system, there is

no delay involved. The signature sequences for the MC and VSL access methods in a virtual low-rate system have different properties. In MC method, $s_{k_l}^{(l)}(t) \neq 0$ and $s_{k_h,m}^{(h)}(t) \neq 0$ for $(i-1)T_l \leq t \leq iT_l$. In the VSL method, $s_{k_l}^{(l)}(t) \neq 0$ for all t, but $s_{k_h,m}^{(h)}(t) \neq 0$ only for every $(i-1)T_l - (m-1)T_h \leq t \leq (i-1)T_l + mT_h$. The VCR signals are also consistent with this model if sampled at high chip rate and has the same constraints as the VSL signals, besides, for low-rate user's signature sequences, every M chips form a group which take on the same value.

The vector form of the equation 4.1 can be written as:

$$
\begin{aligned}
\mathbf{r} &= \mathbf{SWb} + \mathbf{n} \\
&= \begin{bmatrix} \mathbf{s}^{(l)} & \mathbf{s}^{(h)} \end{bmatrix} \begin{bmatrix} \mathbf{W}^{(l)} & \mathbf{0} \\ \mathbf{0}^T & \mathbf{W}^{(h)} \end{bmatrix} \begin{bmatrix} \mathbf{b}^{(l)} \\ \mathbf{b}^{(h)} \end{bmatrix} + \mathbf{n},
\end{aligned}
\tag{4.2}
$$

where

$$
\begin{aligned}
\mathbf{s}^{(l)}(t) &= [s_1^{(l)}(t-\tau_1), \cdots, s_{K_l}^{(l)}(t-\tau_{K_l})], \\
\mathbf{W}^{(l)} &= \mathrm{diag}\left\{ \sqrt{w_1^{(l)}}, \cdots, \sqrt{w_{K_l}^{(l)}} \right\} \\
\mathbf{b}^{(l)} &= [b_1^{(l)}, \cdots, b_{K_l}^{(l)}]^T \\
\mathbf{s}^{(h)}(t) &= [s_{1,1}^{(h)}(t-\tau_1), \cdots, s_{K_h,1}^{(h)}(t-\tau_{K_h}), \cdots, s_{1,M}^{(h)}(t-\tau_1), \cdots, s_{K_h,M}^{(h)}(t-\tau_{K_h})] \\
\mathbf{W}^{(h)} &= \mathrm{diag}\left\{ \sqrt{w_{1,1}^{(h)}}, \cdots, \sqrt{w_{K_1,1}^{(h)}}, \cdots, \sqrt{w_{1,M}^{(h)}}, \cdots, \sqrt{w_{K_1,M}^{(h)}} \right\} \\
\mathbf{b}^{(h)} &= [b_{1,1}^{(h)}, \cdots, b_{K_1,1}^{(h)}, \cdots, b_{1,M}^{(h)}, \cdots, b_{K_1,M}^{(h)}]^T \\
\mathbf{n} &\sim \mathcal{N}(0, \mathbf{I}_{N_l})
\end{aligned}
$$

$((\cdot)^{(l)}, (\cdot)^{(h)}$ denote low-rate user and high-rate user, $m = 1, 2, \cdots, M)$

All the users are assumed to be in the order of K_l low-rate users followed by K_h high-rate users. The output of k_lth low-rate matched filter which is sampled at the end of ith symbol interval ($i = 0, 1, 2, \ldots$) is:

$$
y_{k_l}^{(l)}(i) = \int_{iT_l + \tau_{k_l}}^{(i+1)T_l + \tau_{k_l}} r(t) s_{k_l}^{(l)}(t - iT_l - \tau_{k_l}) dt.
\tag{4.3}
$$

The vector form of all low-rate users' matched filter output in the time interval $[iT_l, (i+1)T_l)$ is a $K_l \times 1$ vector.

$$
\mathbf{y}^{(l)}(i) = [y_1^{(l)}(i), y_2^{(l)}(i), \cdots, y_{K_l}^{(l)}(i)]^T.
\tag{4.4}
$$

The matched filter output of $((m-1)k_h+j)$th virtual high-rate use which is sampled at the end of mth subinterval of $[iT_l, (i+1)T_l)$ is:

$$y_{k_h,m}^{(h)}(i) = \int_{iT_l+(m-1)T_h+\tau_{k_h}}^{iT_l+mT_h+\tau_{k_h}} r(t)s_{k_h,m}(t - iT_l - (m-1)T_h - \tau_{k_h})dt. \tag{4.5}$$

The vector form of all MK_h virtual high-rate users' matched filter output during the time interval $[iT_l, (i+1)T_l)$ is the $MK_h \times 1$ vector

$$\mathbf{y}^{(h)}(i) = [y_{1,1}^{(h)}(i) \cdots y_{K_h,1}^{(h)}(i) \cdots y_{1,M}^{(h)}(i) \cdots y_{K_h,M}^{(h)}(i)]^T. \tag{4.6}$$

The total output vector of $K = K_l + K_h$ matched filters during the time interval $[iT_l, (i+1)T_l)$ is a $(K_l + MK_h) \times 1$ vector.

$$\begin{aligned}
\mathbf{y}(i) &= [\mathbf{y}^{(l)T}(i), \mathbf{y}^{(h)T}(i)]^T \\
&= \mathbf{R}(1)\mathbf{W}\mathbf{b}(i-1) + \mathbf{R}(0)\mathbf{W}\mathbf{b}(i) + \mathbf{R}(-1)\mathbf{W}\mathbf{b}(i+1) + \mathbf{n}(i), \quad (4.7)
\end{aligned}$$

where

$$\mathbf{W} = \begin{bmatrix} \mathbf{W}^{(l)} & \mathbf{0} \\ \mathbf{0} & \mathbf{W}^{(h)} \end{bmatrix},$$

$$\mathbf{b}(i) = \begin{bmatrix} \mathbf{b}^{(l)}(i) \\ \mathbf{b}^{(h)}(i) \end{bmatrix}$$

and $\mathbf{R}^{(i)}(j), j \in \{0, \pm 1\}$, is the cross-correlation matrix with klth element $R_{kl}^{(i)}(j) = \int_{iT_l}^{(i+1)T_l} s_k(t - iT_l - \tau_k)s_l(t - iT_l + jT_l - \tau_l)dt$ and $\mathbf{R}^{(i)}(j) = \mathbf{0}, \forall |j| > 1$. $\mathbf{R}^{(i)}(-1) = \mathbf{R}^{(i+1)}(1)$. The sampled output vector of ith bit within time interval $[iT_l, (i+1)T_l)$ can be written as

$$\mathbf{y}(i) = \begin{bmatrix} \mathbf{R}^{(i)}(1) & \mathbf{R}^{(i)}(0) & \mathbf{R}^{(i)}(-1) \end{bmatrix} \cdot \begin{bmatrix} \mathbf{z}(i-1) \\ \mathbf{z}(i) \\ \mathbf{z}(i+1) \end{bmatrix} + \mathbf{n}(i), \tag{4.8}$$

where $\mathbf{z}(i) = \mathbf{W}\mathbf{b}(i)$, $\mathbf{n}(i)$ is a Gaussian noise vector with zero mean and covariance matrix $\sigma^2\mathbf{R}(0)$. In the case of synchronous transmission channel, delays of all the users are equal, $\mathbf{R}(1), \mathbf{R}(-1)$ vanish. That means when the channel is asynchronous transmission channel, there exists MAI which is induced by the previous symbol and the next symbol. It appears in $\mathbf{R}(1)\mathbf{z}(i-1)$ and $\mathbf{R}(-1)\mathbf{z}(i+1)$. The multiuser interference introduced by the present bit appears in the term $\mathbf{R}(0)\mathbf{z}(i)$.

The matched filter output of $((m-1)k_h+j)$th virtual high-rate use which is sampled at the end of mth subinterval of $[iT_l, (i+1)T_l)$ is:

$$y_{k_h,m}^{(h)}(i) = \int_{iT_l+(m-1)T_h+\tau_{k_h}}^{iT_l+mT_h+\tau_{k_h}} r(t)s_{k_h,m}(t - iT_l - (m-1)T_h - \tau_{k_h})dt. \qquad (4.5)$$

The vector form of all MK_h virtual high-rate users' matched filter output during the time interval $[iT_l, (i+1)T_l)$ is the $MK_h \times 1$ vector

$$\mathbf{y}^{(h)}(i) = [y_{1,1}^{(h)}(i) \cdots y_{K_h,1}^{(h)}(i) \cdots y_{1,M}^{(h)}(i) \cdots y_{K_h,M}^{(h)}(i)]^T. \qquad (4.6)$$

The total output vector of $K = K_l + K_h$ matched filters during the time interval $[iT_l, (i+1)T_l)$ is a $(K_l + MK_h) \times 1$ vector.

$$\begin{aligned}
\mathbf{y}(i) &= [\mathbf{y}^{(l)T}(i), \mathbf{y}^{(h)T}(i)]^T \\
&= \mathbf{R}(1)\mathbf{W}\mathbf{b}(i-1) + \mathbf{R}(0)\mathbf{W}\mathbf{b}(i) + \mathbf{R}(-1)\mathbf{W}\mathbf{b}(i+1) + \mathbf{n}(i), \quad (4.7)
\end{aligned}$$

where

$$\mathbf{W} = \begin{bmatrix} \mathbf{W}^{(l)} & \mathbf{0} \\ \mathbf{0} & \mathbf{W}^{(h)} \end{bmatrix},$$

$$\mathbf{b}(i) = \begin{bmatrix} \mathbf{b}^{(l)}(i) \\ \mathbf{b}^{(h)}(i) \end{bmatrix}$$

and $\mathbf{R}^{(i)}(j), j \in \{0, \pm 1\}$, is the cross-correlation matrix with klth element $R_{kl}^{(i)}(j) = \int_{iT_l}^{(i+1)T_l} s_k(t - iT_l - \tau_k)s_l(t - iT_l + jT_l - \tau_l)dt$ and $\mathbf{R}^{(i)}(j) = \mathbf{0}, \forall |j| > 1$. $\mathbf{R}^{(i)}(-1) = \mathbf{R}^{(i+1)}(1)$. The sampled output vector of ith bit within time interval $[iT_l, (i+1)T_l)$ can be written as

$$\mathbf{y}(i) = \begin{bmatrix} \mathbf{R}^{(i)}(1) & \mathbf{R}^{(i)}(0) & \mathbf{R}^{(i)}(-1) \end{bmatrix} \cdot \begin{bmatrix} \mathbf{z}(i-1) \\ \mathbf{z}(i) \\ \mathbf{z}(i+1) \end{bmatrix} + \mathbf{n}(i), \qquad (4.8)$$

where $\mathbf{z}(i) = \mathbf{W}\mathbf{b}(i)$, $\mathbf{n}(i)$ is a Gaussian noise vector with zero mean and covariance matrix $\sigma^2\mathbf{R}(0)$. In the case of synchronous transmission channel, delays of all the users are equal, $\mathbf{R}(1), \mathbf{R}(-1)$ vanish. That means when the channel is asynchronous transmission channel, there exists MAI which is induced by the previous symbol and the next symbol. It appears in $\mathbf{R}(1)\mathbf{z}(i-1)$ and $\mathbf{R}(-1)\mathbf{z}(i+1)$. The multiuser interference introduced by the present bit appears in the term $\mathbf{R}(0)\mathbf{z}(i)$.

4.3 Multi-shot Receiver
for Asynchronous Dual-rate CDMA Signals

Based on the previous discussion, the matched filtered output of ith bit is:

$$\mathbf{y}(i) = \begin{bmatrix} \mathbf{R}(1) & \mathbf{R}(0) & \mathbf{R}(-1) \end{bmatrix} \cdot \begin{bmatrix} \mathbf{z}(i-1) \\ \mathbf{z}(i) \\ \mathbf{z}(i+1) \end{bmatrix} + \mathbf{n}(i). \tag{4.9}$$

The received signal will be processed in the processing windows of length $N = 2P+1$, where P is a positive integer and the window length N is also called detector memory length measured in symbol duration T_l. When $N \to \infty$, the detectors will become ideal infinite memory length multiuser detectors. In general, the number of information bits is very large. That means if we want to achieve the performance of the ideal linear multiuser detector, N is needed to be very large. This will unavoidably increase the computational complexity of the detectors. In order to make detectors work more efficiently, the averaging method and the de-biasing method are suggested later. The averaging method can improve the performance much, but the better performance the receiver can provide the larger number of N is needed. The de-biasing method is performed through the decomposition of detector $\mathcal{D}$, i.e., $\mathcal{D} = (\mathcal{D}(i), \mathcal{D}(i+1), \cdots, \mathcal{D}(i+N-1))^T$ which has the size of $NK_v \times K_v$ and $K_v \times K_v$ matrix $\mathcal{D}(i)$. This method will be shown later that it can gain much even in the worst case of $N = 3$, which means $P = 1$.

The available N-bit data sequence can be arranged in the following way:

$$\mathbf{Y}^{(N)}(i) = \begin{bmatrix} \mathbf{y}^T(i) & \mathbf{y}^T(i+1) & \cdots & \mathbf{y}^T(i+N-1) \end{bmatrix}^T \tag{4.10}$$

$$= \begin{bmatrix} \mathbf{R}(1) & \mathbf{R}(0) & \mathbf{R}(-1) & 0 & \cdots & 0 \\ 0 & \mathbf{R}(1) & \mathbf{R}(0) & \mathbf{R}(-1) & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & 0 & \cdots & \mathbf{R}(-1) \end{bmatrix} \begin{bmatrix} \mathbf{z}(i-1) \\ \mathbf{z}(i) \\ \vdots \\ \mathbf{z}(i+N) \end{bmatrix}$$

$$+ \begin{bmatrix} \mathbf{n}(i) \\ \mathbf{n}(i+1) \\ \vdots \\ \mathbf{n}(i+N-1) \end{bmatrix} \tag{4.11}$$

$$= \begin{bmatrix} \mathbf{R}(0) & \mathbf{R}(-1) & 0 & \cdots & 0 \\ \mathbf{R}(1) & \mathbf{R}(0) & \mathbf{R}(-1) & \cdots & 0 \\ \vdots & \vdots & \vdots & \cdots & \mathbf{R}(-1) \\ 0 & 0 & 0 & \cdots & \mathbf{R}(0) \end{bmatrix} \begin{bmatrix} \mathbf{z}(i) \\ \mathbf{z}(i+1) \\ \vdots \\ \mathbf{z}(i+N-1) \end{bmatrix}$$

$$+ \begin{bmatrix} \mathbf{R}(1) & \mathbf{0} \\ \mathbf{0} & \mathbf{0} \\ \vdots & \vdots \\ \mathbf{0} & \mathbf{R}(-1) \end{bmatrix} \begin{bmatrix} \mathbf{z}(i-1) \\ \mathbf{z}(i+N) \end{bmatrix} + \begin{bmatrix} \mathbf{n}(i) \\ \mathbf{n}(i+1) \\ \vdots \\ \mathbf{n}(i+N-1) \end{bmatrix}$$

$$= \mathcal{R}\mathcal{Z} + \mathcal{R}_e\mathcal{Z}_e + \mathcal{N} \tag{4.12}$$

$$= \mathcal{R}\mathcal{W}\mathbf{b}^{(N)} + \mathcal{R}_e\mathcal{W}_e\mathbf{b}_e^{(N)} + \mathcal{N}, \tag{4.13}$$

where

$$\mathcal{Z} = \mathcal{W}\mathbf{b}^{(N)} = \text{diag}\{\mathbf{W}, \cdots, \mathbf{W}\} \cdot [\mathbf{b}^T(i), \cdots, \mathbf{b}^T(i+N-1)]^T,$$

$$\mathcal{Z}_e = \mathcal{W}_e\mathbf{b}_e^{(N)} = \text{diag}\{\mathbf{W}\mathbf{W}\} \cdot [\mathbf{b}^T(i-1), \mathbf{b}^T(i+N)]^T$$

In equation 4.13, the first term $\mathcal{R}\mathcal{Z}$ is contributed by the bits inside the processing window. The second term $\mathcal{R}_e\mathcal{Z}_e$ is contributed by the bits on the edge of the processing window. The first term of equation 4.13 is an approximation of equation 4.12 since the bias term is almost a zero vector except for the first and the last sub-block. If making the truncation, neglecting the bias term $\mathcal{R}_e\mathcal{Z}_e$, the estimation of the information bit will be easy to do [22]. It was proven that $\mathcal{R}$ in 4.13 is a $NK_v \times NK_v$ symmetric, positive definite, block-tridiagonal and full rank matrix. The noise term $\mathcal{N}$ is a $NK_v \times 1$, zero mean, Gaussian noise vector with covariance matrix $\sigma^2\mathcal{R}$.

As we all know, a linear multiuser detector processes the matched filter output vector $\mathbf{Y}$ by a linear operation $\mathbf{T}$. In other words, the detector output $\mathbf{X}$ is given by $\mathbf{X} = \mathbf{T}^T\mathbf{Y}$. If $N = N_b$, where N_b is data packet length, we have ideal detectors for packetized transmission. The decorrelating detector here is $\mathbf{T} = \mathcal{R}^{-1}$. If the information bit $b_k(i)$ are independent and uniformly distributed, the LMMSE detector is $\mathbf{T} = (\mathcal{R} + \sigma^2\mathcal{W}^{-2})^{-1}$. If the size of the data packet is very large, the ideal detectors described above may not be feasible. To obtain more practical detectors, we introduce the multi-shot receiver which was proposed for single-rate asynchronous CDMA systems [22] to dual-rate asynchronous CDMA systems. A multi-shot detector is working on the finite number (e.g. $N = 2P + 1$) of matched filter outputs. The decision statistic of N-bit multiuser information of low-rate multi-shot decorrelating detector is:

$$\mathbf{X}_D^{(N)}(i) = \mathcal{R}^{-1}\mathbf{Y}^{(N)}(i)$$

$$= \mathcal{Z}^{(N)}(i) + \mathcal{R}^{-1} \begin{bmatrix} \mathbf{R}(1)\mathbf{z}(i-1) \\ 0 \\ \vdots \\ 0 \\ \mathbf{R}(-1)\mathbf{z}(i+N) \end{bmatrix} + \mathcal{N}_D, \qquad (4.14)$$

where $\mathcal{N}_D$ is a $NK_v \times 1$ Gaussian noise vector with zero mean and covariance matrix $\sigma^2 \mathcal{R}^{-1}$. The N-bit estimated output information of low-rate multishot decorrelating detector is: $\hat{\mathbf{B}}_D^{(N)}(i) = \text{sgn}\{\mathbf{X}_D^{(N)}(i)\}$. The decision statistic of N bit multiuser information of low-rate multi-shot LMMSE receiver is:

$$\begin{aligned} \mathbf{X}_{mmse}^{(N)}(i) &= (\mathcal{R} + \sigma^2 \mathcal{W}^{-2})^{-1} \mathbf{Y}^{(N)}(i) \\ &= (\mathcal{R} + \sigma^2 \mathcal{W}^{-2})^{-1} \mathcal{R} \mathcal{Z}^{(N)}(i) \\ &\quad + (\mathcal{R} + \sigma^2 \mathcal{W}^{-2})^{-1} \begin{bmatrix} \mathbf{R}(1)\mathbf{z}(i-1) \\ 0 \\ \vdots \\ 0 \\ \mathbf{R}(-1)\mathbf{z}(i+N) \end{bmatrix} + \mathcal{N}_{mmse}, \qquad (4.15) \end{aligned}$$

where $\mathcal{N}_{mmse}$ is a $NK_v \times 1$ Gaussian noise vector with zero mean and covariance matrix $(\mathcal{R} + \sigma^2 \mathcal{W}^{-2})^{-1} \mathcal{R} (\mathcal{R} + \sigma^2 \mathcal{W}^{-2})^{-T}$. The N bit estimated output information of low-rate multishot LMMSE receiver is: $\hat{\mathbf{B}}_{mmse}^{(N)}(i) = \text{sgn}\{\mathbf{X}_{mmse}^{(N)}(i)\}$. Specifically, the matrix $\mathcal{R}$ in equation 4.13 has a block-tridiagonal structure since $\mathbf{R}(0)$ is a symmetric, positive definite matrix, $\mathbf{R}(1)$ is an upper triangle matrix, $\mathbf{R}(-1) = \mathbf{R}^T(1)$. So the above $\mathcal{R}$ is a symmetric, positive definite matrix of full rank and it is also a diagonal dominant matrix. Therefore, the inverse $\mathcal{R}^{-1}, (\mathcal{R} + \sigma^2 \mathcal{W}^{-2})^{-1}$ are also diagonal dominant matrices. For asynchronous transmission channel, it is noted that the dimension of $\mathcal{R}$ is in general larger than $K_l + MK_h$ for a virtual low-rate system, depending on the actual number of virtual bits present within the processing window. In such case, the $\mathcal{R}$ matrix can be either full rank or rank deficient, depending on the relation between all virtual signatures or bases associated with all virtual bits. When the matrix $\mathcal{R}$ is singular, one is still able to decode a desired user's bit as long as its full-length signature is present within the processing interval. However, in such cases, one should use pseudo-inverse of $\mathcal{R}$ in the decorrelating detector. If the multi-shot LMMSE detector is used, we don't need to worry about the singularity problem, since our filtering matrix is never singular, as long as there is noise present in data.

According to the QoS requirement, we can set our processing interval equal to either the minimum value specified by the symbol interval of high-rate users, T_h, or the maximum value specified by the symbol interval of low-rate users, T_l. Correspondingly, we refer to these receivers as high-rate mode receivers and low-rate mode receivers. As pointed out in the previous chapters, the complexity of the problem and the processing delay of the detectors change with the choice of the processing interval. Note that in asynchronous case, the interference pattern changes with each sub-interval of length T_h. Therefore, in actually implementing the data-driven adaptive high-rate mode LMMSE receiver, firstly, we need to translate received data r_m for each subinterval into M parallel streams, and use M parallel channels to decode the desired high-rate user's bit streams. The matrix $\hat{\Sigma}_{r_m}^{-1}$ associated with each parallel channel can then be sequentially estimated and updated from each data stream $\{r_m(i)\}$. Considering the low-rate user in a high-rate mode LMMSE receiver, since its information is inbedded within a longer symbol interval $T_l = MT_h$, using partial data r_m of the mth subinterval to decode low-rate users' information causes performance degradation for low-rate users. In order to alleviate this problem, we also need to use the signal to interference plus noise ratio ($SINR_k$) of desired low-rate user within each sub-interval to weight partial results obtained in each parallel channel to make a better decision on the whole low-rate user's information bit. The total computational overhead associated with the high-rate mode LMMSE detector makes its application to asynchronous multi-rate CDMA less attractive than the low-rate mode LMMSE receiver proposed in this chapter. Intuitively, using the low-rate mode LMMSE detector, one can more reliably estimate/update the interference pattern involved in $\hat{\Sigma}_r^{-1}$ over a longer processing interval T_l. Since within each interval of length T_l, the interference pattern repeats itself periodically, it is also computationally efficient in using the low-rate LMMSE detector for the asynchronous applications.

4.3.1 Two Processing Methods of Dual-rate Multi-shot LMMSE Receiver

- Averaging Method

  The averaging method is performed by making every length N memory as a sliding window and taking an average through this N bit available detector

$$\mathcal{D}_d^{(3)} = (\mathcal{R}^{(3)})^{-1} \mathbf{u}_3$$

$$= \begin{bmatrix} \mathbf{R}(0) & \mathbf{R}(-1) & \mathbf{0} \\ \mathbf{R}(1) & \mathbf{R}(0) & \mathbf{R}(-1) \\ \mathbf{0} & \mathbf{R}(1) & \mathbf{R}(0) \end{bmatrix}^{-1} \cdot \mathbf{u}_3. \tag{4.20}$$

The decision statistic of the middle item is:

$$\mathbf{x}_d^{(3)}(i+1) = \underbrace{((\mathcal{R}^{(3)})^{-1} \mathbf{u}_3)}_{\mathcal{D}_d^{(3)}}^T \cdot \begin{bmatrix} \mathbf{y}(i) \\ \mathbf{y}(i+1) \\ \mathbf{y}(i+2) \end{bmatrix}. \tag{4.21}$$

The estimated information bit can be obtained by taking the sign of the corresponding decision statistic, $\hat{b}_d^{(N)}(i+1) = \text{sgn}\{\mathbf{x}_d^{(N)}(i+1)\}$. In general, the $NK_v \times K_v$ matrix $\mathcal{D}$ can be calculated in advance by iteratively using the results of block matrix (of less dimension) inversion and the nice structure of $\mathcal{R}$. For the $N = 3$ case, the $3K_v \times K_v$ matrix can be calculated as follow:

$$\mathcal{D}_d^{(3)} = \mathbf{V}_0 \begin{bmatrix} \mathbf{R}(1)\mathbf{R}^{-1}(0) & -\mathbf{I}_{K_v} & \mathbf{R}(-1)\mathbf{R}^{-1}(0) \end{bmatrix}^T, \tag{4.22}$$

where

$$\mathbf{V}_0 = -(\mathbf{R}(0) - \mathbf{R}(-1)\mathbf{R}^{-1}(0)\mathbf{R}(1) - \mathbf{R}(1)\mathbf{R}^{-1}(0)\mathbf{R}(-1))^{-1}$$

For the LMMSE detector $N = 3$ case, the generalized multi-shot detector $\mathcal{D}$ should satisfy the condition

$$(\mathcal{R}^{(3)} + \sigma^2 \mathcal{W}_3^{-2})\mathcal{D}_{mmse}^{(3)} = \mathbf{u}_3$$

$$\mathcal{D}_{mmse}^{(3)} = (\mathcal{R}^{(3)} + \sigma^2 \mathcal{W}_3^{-2})^{-1} \mathbf{u}_3. \tag{4.23}$$

The decision statistic of the middle item equals to:

$$\mathbf{x}_{mmse}^{(3)}(i+1) = ((\mathcal{R}^{(3)} + \sigma^2 \mathcal{W}_3^{-2})^{-1} \mathbf{u}_3)^T \cdot \begin{bmatrix} \mathbf{y}(i) \\ \mathbf{y}(i+1) \\ \mathbf{y}(i+2) \end{bmatrix}, \tag{4.24}$$

The estimated information bit can be obtained by taking the sign of the corresponding decision statistic. $\hat{b}_{mmse}^{(N)}(i+1) = \text{sign}\{\mathbf{x}_{mmse}^{(N)}(i+1)\}$. Through the above de-biasing process, we can achieve the near-far resistance with $N = 3$, and instead of doing the matrix inversion of $\mathcal{R}$ which has the size of $NK_v \times NK_v$, only the matrix inversion of $\mathbf{R}$ which has the size of $K_v \times K_v$ is involved.

4.4 Simulation Results and Discussions

In chapter 3, we numerically evaluate and compare the performance of the LMMSE receiver over the high-rate bit interval and low-rate bit interval for the synchronous transmission channel. We also compare the performance difference between the VSL system and MC system for dual-rate LMMSE receivers for synchronous transmission channel in the previous chapter. In this chapter, we provide the simulation results for the detectors which we proposed in this chapter. Firstly, we provide the simulation results which show the performance improvement with averaging method and debiasing method for a dual-rate virtual low-rate asynchronous CDMA system. Secondly, we provide the simulation results which show the performance improvement between the multi-shot LMMSE receiver and multi-shot decorrelating detector for multi-rate CDMA signals. Also in the second part of simulation results, we show the performance improvement between three-shot detectors with de-biasing and without debiasing for multi-rate signals. Thirdly, we provide simulation results for comparison of the VSL and MC systems.

A direct sequence CDMA system with BPSK data and coherent detection is considered. There are four users in this system, which are two low-rate users and two high-rate users. This means $K_l = 2$ and $K_h = 2$. The rate-ratio is set to be equal to $M = 2$. For the VSL system, the processing gain for high-rate users is set to be $L_h = 7$, the processing gain of low-rate users is set to be $L_l = 2 \times 7 = 14$. The repetition coding scheme is used for low-rate users which means that the code repeat itself M times during one low-rate bit interval. The Gold sequences are chosen to be the spreading sequences. The SNR of the kth user of the system will be $10 \log \frac{w_k}{N_0/2}$. The transmission delays associated with all four users are chosen as: $\tau_1 = 0, \tau_2 = T_c, \tau_3 = 2T_c, \tau_4 = 3T_c$.

In Fig. 4.4, the bit error rate (BER) of low-rate user 1 is plotted as a function of increasing SNR of the other users. As expected, for the virtual low-rate VSL system with users transmitting at two different data rates, the three-shot decorrelating detector with averaging method and debiasing method can provide performance improvement over three-shot decorrelating detector. Three-shot debiasing decorrelating detector can almost attain near-far resistance. Fig. 4.5 shows the same simulation results of the performance comparison of high-rate user 3 between
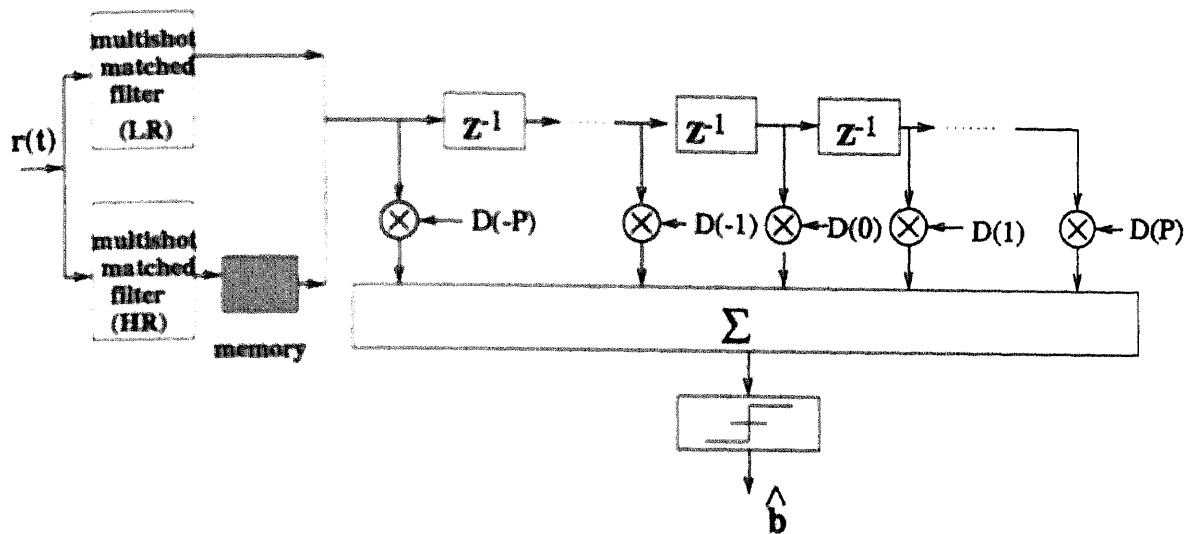
Figure 4.3 Low-rate mode multi-shot receiver

the three-shot decorrelator, three-shot decorrelator with averaging and three-shot debiasing detector. In Fig. 4.6, the bit error rate of low-rate user 1 is again plotted as a function of increasing SNR of the other users and Fig. 4.7 is the similar results of the high-rate user 1. Comparing to the multi-shot decorrelating detector, multi-shot LMMSE receiver can provide better performance when the other user's SNR is lower than the desired user's SNR. The proposed multi-shot LMMSE debiasing detector can attain the upper bound of multi-shot debiasing decorrelator when the other users' SNR is high and single user lower bound when the other users' SNR is low comparing to the desired user's SNR.

The performance comparison of VSL and MC system is shown in Fig. 4.8 and Fig. 4.9 for low-rate user 1 and high-rate user 1. Comparing to the corresponding results of the VSL systems, it is clear that the VSL access systems offer a distinct advantage over the MC system.
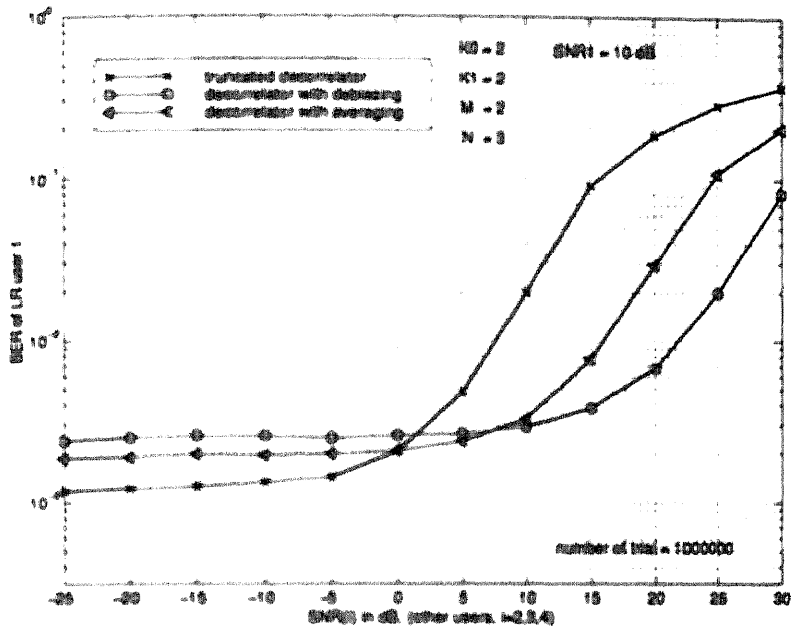
Figure 4.4 Four user VSL system, $K_l = 2, K_h = 2, M = 2, L_l = 14, L_h = 7$, window length $N = 3$, SNR of low-rate user 1 $SNR1 = 10dB$
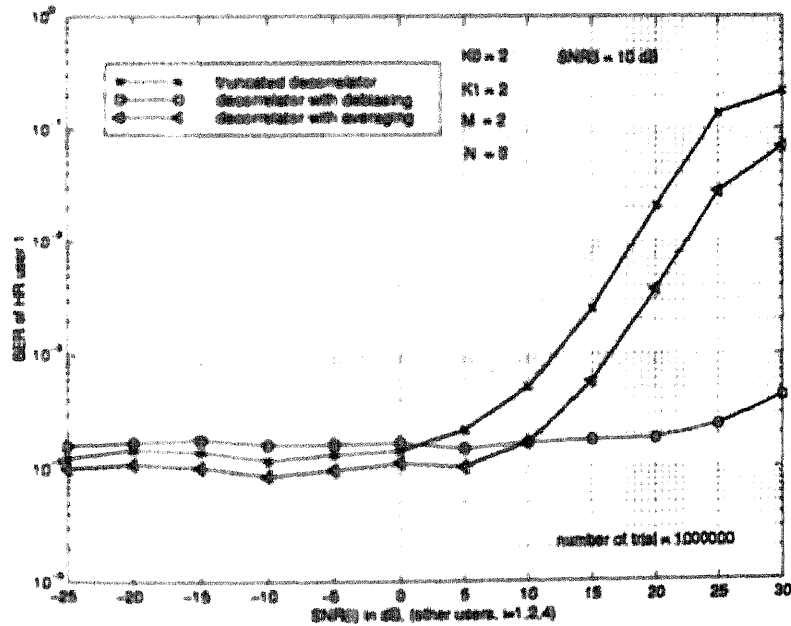
Figure 4.5 Four user VSL system, $K_l = 2, K_h = 2, M = 2, L_l = 14, L_h = 7$, window length $N = 3$, SNR of high-rate user 3 $SNR3 = 10dB$
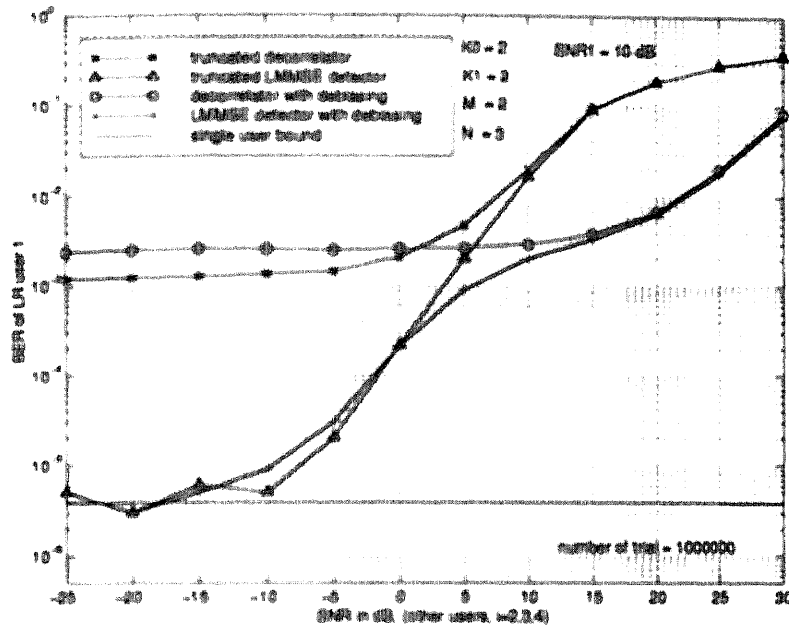
Figure 4.6 Four user VSL system, $K_l = 2, K_h = 2, M = 2, L_l = 14, L_h = 7$, window length $N = 3$, SNR of low-rate user 1 $SNR1 = 10dB$
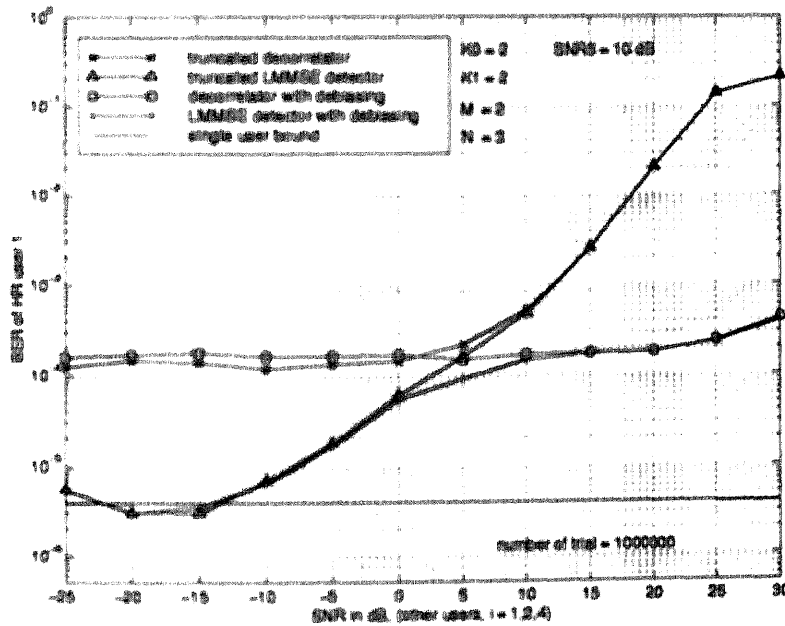
Figure 4.7 Four user VSL system, $K_l = 2, K_h = 2, M = 2, L_l = 14, L_h = 7$, window length $N = 3$, SNR of high-rate user 3 $SNR3 = 10dB$
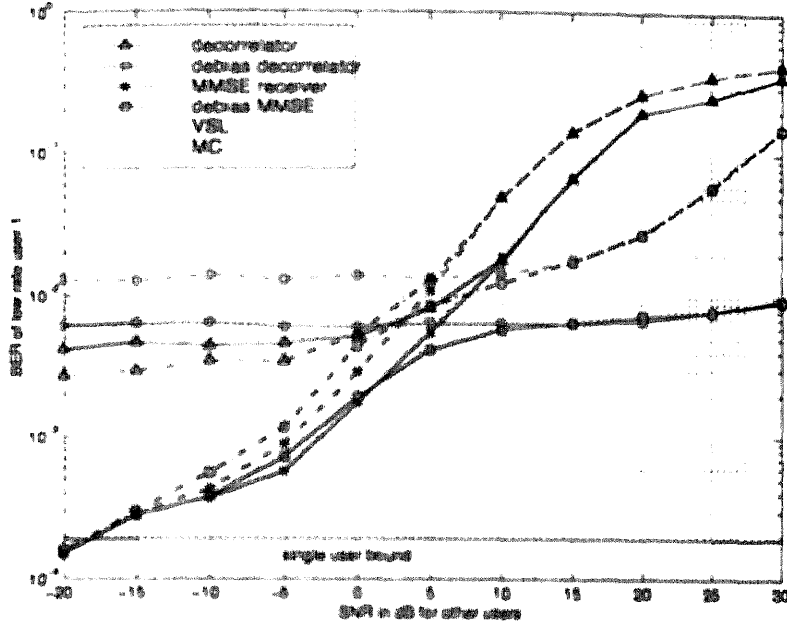
Figure 4.8 Three user system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$ for VSL system, $L = 14$ for MC system. SNR of low-rate user 1: $SNR1 = 8dB$
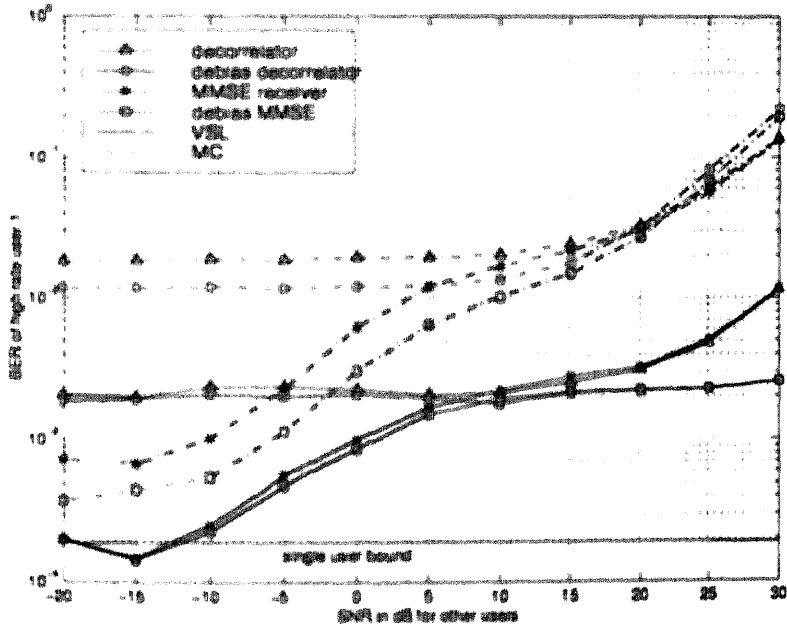
Figure 4.9 Three user system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$ for VSL system, $L = 14$ for MC system. SNR of high-rate user 1: $SNR3 = 8dB$

CHAPTER 5

LINEAR SPACE-TIME RECEIVER FOR MULTI-RATE CDMA SYSTEMS

5.1 Introduction

Due to the growing user demands for wireless communication services, there has been much interest in finding alternative methods of increasing capacities of wireless systems. For CDMA systems, since the spreading sequences are long pseudo-random sequences of 1's and -1's, users' transmitted signals are spread over a bandwidth which is much wider than the data rate. By definition, signals with these characteristics are spread spectrum signals, and their bandwidth is wide enough so that the channel affects the subbands differently. Such channels are said to be frequency selective. Because of signal scatters in wireless channels, the received signal contains multipath components which are delayed and distorted replicas of the original transmitted signal. While multipath signals are often viewed as a detriment to many communication systems, the frequency selective nature of the CDMA channel allows the detector to take advantage of them by resolving their individual signal components, thereby providing a form of frequency diversity. The well-known Rake receiver [13] is optimal for combating multipath fading in the absence of MAI. The multiuser detection [3] is optimal for suppressing the MAI. More recently, one emerging concept that has been receiving much attention is Space Division Multiple Access. This technique is focused on using multiple antennas at the base station receivers to provide antenna or diversity gain and to exploit the spatial distribution of users through spatial filtering [55]. The application of SDMA techniques to wireless CDMA systems has been investigated by many researchers. In [56], a linear multiuser array detector is derived which used a front end array of sensors and which exploits knowledge of the received signals' directions of arrival. In doing so, the linear multiuser array detector provides significant performance gains over the single sensor linear multiuser decorrelating detector. By combining these space-domain techniques with the time domain techniques like Rake receiver and multiuser detection, the resulting space-time detectors [28], [55] show potential of improving the capacity and performance of CDMA systems over traditional time-domain-only detectors.

Firstly, the low-rate mode LMMSE space-time receiver for dual-rate CDMA signals is proposed. In this receiver, there are multiple antennas at the front end of the receiver. After that, the code array/channel matched filters which correspond to low-rate users and high-rate users are followed. The LMMSE estimation-based detection which is working over the low-rate users' information bit interval tries to minimize the MSE between the output data and the transmitted information after the code array/channel matched filters. The disadvantage of this receiver is that the explicit knowledge of the array and channel coefficients (spatial signatures) are needed in code array/channel matched filter part. Also, the estimation of the channel coefficients cannot be incorporated into the adaptive LMMSE algorithms. In order to solve the problems, an alternative approach for this receiver is proposed and is called type two low-rate mode LMMSE space-time receiver for dual-rate CDMA signals. In this alternative approach, the array/channel weighting, combining and linear multiuser detection are collected into a single matrix followed the code matched filters. No explicit knowledge of the array and channel coefficients are needed in this approach, and the estimation of the channel coefficients can be incorporated into certain LMMSE adaptive algorithms. The above two approaches have the disadvantages of low-rate mode receivers which are high dimension, computational complexity and processing delay for high rate users.

Secondly, the two-stage LMMSE space-time receiver for dual-rate CDMA signals is proposed. The proposed two-stage LMMSE space-time receiver generates the decisions for high rate users at every interval of T_h. After the first stage which is working over the $(M - 1)$ subintervals of T_h, the low rate users' decision is made in the second stage based on the information provided by the previous stage. Also the high-rate user's decision of Mth subinterval is made in the second stage. The two-stage LMMSE space-time receiver reduces the computational complexity of the low-rate mode LMMSE space-time receiver and suppress the processing delay for high rate users with sacrificing a minor performance. There are also two approaches for the two-stage LMMSE space-time receiver. The first approach requires the explicit knowledge of the array and channel coefficients at the code array/channel matched filter part. The array and channel coefficients need to be estimated in advance. The second approach can incorporate the channel estimation into the

certain LMMSE adaptive algorithm and only the knowledge of the desired user's signature waveform is required.

5.2 Preliminaries

A two-rate CDMA system using variable spreading length access method is considered. The active users are classified into two groups. Low-rate users with bit interval T_l, high-rate users with bit interval T_h. Bit rate ratio equals to $M = \frac{T_l}{T_h}$, which is set to be an integer. The signature waveforms of different users are normalized to the interested bit interval. There are total number of $K = K_l + K_h$ users where K_l, K_h stand for the number of low-rate and high-rate users. $K_v = K_l + MK_h$ are the total number of virtual users in a virtual low-rate system. Users transmit synchronously over frequency selective Rayleigh fading channels, such as those found in indoor wireless and mobile radio systems. This channel can be represented by a tapped delay line with tap spacing $T_c = 1/B_w$, where B_w is the baseband signal bandwidth. The tap weight coefficients of ith user vth path $c_{i,v}(t)$, $v = 0, 1, \cdots, V - 1$, is modeled as zero mean, complex Gaussian random variables. There are a total of $V = B_w T_m + 1$ [57] resolvable paths where T_m is the channel multipath spread. It is assumed that the symbol interval T_l and T_h is selected such that $T_l \gg T_m, T_h \gg T_m$. This will allow us to ignore the effects of inter-symbol interference. We further assume that the coherence time of the channel is much longer than the symbol intervals, so that a slow-fading channel whose coefficients remain constant at least for one symbol interval is assumed. We consider a P element receiving antenna in the front end of the receiver. They are arranged as a linear array with $\lambda/2$ spacing where λ is the wave-length of the RF signal carrier. We assume that this spacing is sufficient for the users' signals to arrive as planar wavefronts and for their relative time delays at each sensor to be modeled as phase shifts. The phase offset at pth sensor with respect to the first sensor due to ith low-rate user's vth path signal is $(p - 1)\pi \sin \alpha_{i,v}^{(l)}$. The phase offset at pth sensor with respect to the first sensor due to jth high-rate user's vth path signal is $(p - 1)\pi \sin \alpha_{j,v}^{(h)}$, where $\alpha_{i,v}^{(l)}$ or $\alpha_{j,v}^{(h)}$ is the signal's direction of arrival measured with respect to the linear array. The resulting base band received signal at the pth sensor

during the time interval $[0, T_l)$ is:

$$
\begin{aligned}
r_p(t) &= \sum_{i=1}^{K_l} \sqrt{w_i^{(l)}} b_i^{(l)} \left(\sum_{v=1}^{V} a_{i,v,p}^{(l)} c_{i,v}^{(l)} s_{i,v}^{(l)}(t) \right) \\
&\quad + \sum_{j=1}^{K_h} \left\{ \sum_{m=1}^{M} \sqrt{w_j^{(h)}} b_{j,m}^{(h)} \left(\sum_{v=1}^{V} a_{j,v,p}^{(h)} c_{j,v}^{(h)} s_{j,v}^{(h)}(t - (m-1)T_h) \right) \right\} + n_p(t) \\
&= \sum_{i=1}^{K_l} \sqrt{w_i^{(l)}} b_i^{(l)} \left(\sum_{v=1}^{V} h_{i,v,p}^{(l)} s_{i,v}^{(l)}(t) \right) \\
&\quad + \sum_{j=1}^{K_h} \left\{ \sum_{m=1}^{M} \sqrt{w_j^{(h)}} b_{j,m}^{(h)} \left(\sum_{v=1}^{V} h_{j,v,p}^{(h)} s_{j,v}^{(h)}(t - (m-1)T_h) \right) \right\} + n_p(t), \quad (5.1)
\end{aligned}
$$

where $h_{i,v,p}^{(l)} = a_{i,v,p}^{(l)} c_{i,v}^{(l)}$ and $a_{i,v,p}^{(l)} = \exp[j(p-1)\pi \sin \alpha_{i,v}^{(l)}]$, $c_{i,v}^{(l)}$ is the vth path channel coefficient of low-rate user i, $s_{i,v}^{(l)}(t)$ corresponds to vth path of the ith low-rate user's spreading code, $w_i^{(l)}$ and $b_i^{(l)}$ are ith low-rate user's transmit power and information bit respectively. In the similar way, $h_{j,v,p}^{(h)} = a_{j,v,p}^{(h)} c_{j,v}^{(h)}$ and $a_{j,v,p}^{(h)} = \exp[j(p-1)\pi \sin \alpha_{k,v}^{(h)}]$, $c_{j,v}^{(h)}$ is the vth path channel coefficient of high-rate user j, $s_{j,v}^{(h)}(t)$ corresponds to vth path of the jth high-rate user's spreading code, $w_j^{(h)}$ and $b_{j,m}^{(h)}$ are jth high-rate user's transmit power and information bit of over $[(m-1)T_h, mT_h)$. The correlations among the array gains $a_{i,v,p}^{(l)}$ and $a_{j,v,p}^{(h)}$ depend on factors such as the antenna spacing, the signal's angle of arrival, and it's scattering angle (beamwidth) [58]. We assume that $h_{i,v,p}^{(l)}$ and $h_{j,v,p}^{(h)}$ are deterministic quantities which we can estimate.

Let

$$
\mathbf{A}^{(l)} \equiv \left[\mathbf{A}_1^{(l)} \cdots \mathbf{A}_{K_l}^{(l)} \right] \tag{5.2}
$$

is the $P \times V K_l$ array gain matrix of low-rate users where ith low-rate user's array gain matrix is:

$$
\mathbf{A}_i^{(l)} = \begin{bmatrix} a_{i,1,1}^{(l)} & \cdots & a_{i,V,1}^{(l)} \\ \vdots & \ddots & \vdots \\ a_{i,1,P}^{(l)} & \cdots & a_{i,V,P}^{(l)} \end{bmatrix}. \tag{5.3}
$$

By defining the follow vectors:

$$
\mathbf{a}_{i,p}^{(l)} = [a_{i,1,p}^{(l)} \cdots a_{i,V,p}^{(l)}]^T, \tag{5.4}
$$

the $\mathbf{A}_i^{(l)}$ can be written as:

$$\mathbf{A}_i^{(l)} = \begin{bmatrix} \mathbf{a}_{i,1}^{(l)T} \\ \vdots \\ \mathbf{a}_{i,P}^{(l)T} \end{bmatrix}. \tag{5.5}$$

For the high-rate users, we assume the array gain does not change within the time interval $[0, T_l)$, so for any subinterval $[(m-1)T_h, mT_h)$, the array gain matrix for high-rate users is defined as:

$$\mathbf{A}^{(h)} \equiv \begin{bmatrix} \mathbf{A}_1^{(h)} \cdots \mathbf{A}_{K_h}^{(h)} \end{bmatrix}, \tag{5.6}$$

which is a $P \times VK_h$ matrix with each component $\mathbf{A}_j^{(h)}$ being the size $P \times V$ array gain matrix of jth high-rate user,

$$\mathbf{A}_j^{(h)} = \begin{bmatrix} a_{j,1,1}^{(h)} & \cdots & a_{j,V,1}^{(h)} \\ \vdots & \ddots & \vdots \\ a_{j,1,P}^{(h)} & \cdots & a_{j,V,P}^{(h)} \end{bmatrix}. \tag{5.7}$$

By defining the follow vectors:

$$\mathbf{a}_{j,p}^{(h)} = [a_{j,1,p}^{(h)} \cdots a_{j,V,p}^{(h)}]^T, \tag{5.8}$$

$\mathbf{A}_j^{(h)}$ can be written as:

$$\mathbf{A}_j^{(h)} = \begin{bmatrix} \mathbf{a}_{j,1}^{(h)T} \\ \cdots \\ \mathbf{a}_{j,P}^{(h)T} \end{bmatrix}. \tag{5.9}$$

The noiseless chip rate sampled version of received information of K_l low-rate users at the pth sensor during $[(m-1)T_h, mT_h)$ is the N_h vector:

$$\mathbf{r}_{m,p}^{(l)} = \mathbf{S}_m^{(l)}(\mathbf{A}_p^{(l)} \circ \mathbf{C}^{(l)})\mathbf{W}^{(l)}\mathbf{b}^{(l)} = \mathbf{S}_m^{(l)}\mathbf{H}_p^{(l)}\mathbf{W}^{(l)}\mathbf{b}^{(l)}, \tag{5.10}$$

where

$$
\begin{aligned}
\mathbf{H}_p^{(l)} &= \mathbf{A}_p^{(l)} \circ \mathbf{C}^{(l)} & VK_l \times K_l \\
\mathbf{S}^{(l)} &= [\mathbf{s}_{1,1}^{(l)} \cdots \mathbf{s}_{1,V}^{(l)} \cdots \mathbf{s}_{K_l,1}^{(l)} \cdots \mathbf{s}_{K_l,V}^{(l)}] & N_l \times VK_l \\
\mathbf{A}_p^{(l)} &= \mathrm{diag}(\mathbf{a}_{1,p}^{(l)} \cdots \mathbf{a}_{K_l,p}^{(l)}) & VK_l \times K_l \\
\mathbf{C}^{(l)} &= \mathrm{diag}(\mathbf{c}_1^{(l)} \cdots \mathbf{c}_{K_l}^{(l)}) & VK_l \times K_l \\
\mathbf{c}_i^{(l)} &= [c_{i,1}^{(l)} \cdots c_{i,V}^{(l)}]^T & V \times 1 \\
\mathbf{W}^{(l)} &= \mathrm{diag}\left(\sqrt{w_1^{(l)}} \cdots \sqrt{w_{K_l}^{(l)}}\right) & K_l \times K_l \\
\mathbf{b}^{(l)} &= [b_1^{(l)} \cdots b_{K_l}^{(l)}]^T & K_l \times 1
\end{aligned}
$$

($\mathbf{x} \circ \mathbf{y}$ denotes the component-wise product of two similarly sized matrices $\mathbf{x}, \mathbf{y}$). Noting that the signature waveform $\mathbf{S}_m^{(l)}(n) = \mathbf{S}^{(l)}(n)$ for $(m-1)N_h \leq n < mN_h$. $\mathbf{S}_m^{(l)}$ has the size of $N_h \times VK_l$. When the repetition coding scheme is used, $\mathbf{S}_1^{(l)} = \cdots = \mathbf{S}_M^{(l)}$ and $\mathbf{S}^{(l)} = [\mathbf{S}_1^{(l)^T} \cdots \mathbf{S}_M^{(l)^T}]^T$. The noiseless chip-rate sampled version of received information of K_h high-rate users at the pth sensor during $[(m-1)T_h, mT_h)$ is the N_h vector:

$$\mathbf{r}_{m,p}^{(h)} = \mathbf{S}_m^{(h)}(\mathbf{A}_{m,p}^{(h)} \circ \mathbf{C}_m^{(h)})\mathbf{W}_m^{(h)}\mathbf{b}_m^{(h)} = \mathbf{S}_m^{(h)}\mathbf{H}_{m,p}^{(h)}\mathbf{W}_m^{(h)}\mathbf{b}_m^{(h)}, \tag{5.11}$$

where

$$
\begin{aligned}
\mathbf{H}_{m,p}^{(h)} &= \mathbf{A}_{m,p}^{(h)} \circ \mathbf{C}_m^{(h)} & VK_h \times K_h \\
\mathbf{S}_m^{(h)} &= [\mathbf{s}_{1,1}^{(h)} \cdots \mathbf{s}_{1,V}^{(h)} \cdots \mathbf{s}_{K_h,1}^{(h)} \cdots \mathbf{s}_{K_h,V}^{(h)}] & N_h \times VK_h \\
\mathbf{A}_{m,p}^{(h)} &= \mathrm{diag}(\mathbf{a}_{1,p}^{(h)} \cdots \mathbf{a}_{K_h,p}^{(h)}) & VK_h \times K_h \\
\mathbf{C}_m^{(h)} &= diag(\mathbf{c}_1^{(h)} \cdots \mathbf{c}_{K_h}^{(h)}) & VK_h \times K_h \\
\mathbf{c}_j^{(h)} &= [c_{j,1}^{(h)} \cdots c_{j,V}^{(h)}]^T & V \times 1 \\
\mathbf{W}_m^{(h)} &= \mathrm{diag}\left(\sqrt{w_1^{(h)}} \cdots \sqrt{w_{K_h}^{(h)}}\right) & K_h \times K_h \\
\mathbf{b}_m^{(h)} &= [b_{1,m}^{(h)} \cdots b_{K_h,m}^{(h)}]^T & K_h \times 1
\end{aligned}
$$

The resulting chip-rate sampled version of received signal during $[(m-1)T_h, mT_h)$ at pth sensor is:

$$\mathbf{r}_{m,p} = \mathbf{r}_{m,p}^{(l)} + \mathbf{r}_{m,p}^{(h)} + \mathbf{n}_{m,p} = \mathbf{S}_m\mathbf{H}_{m,p}\mathbf{W}_m\mathbf{b}_m + \mathbf{n}_{m,p}, \tag{5.12}$$

where

$$
\begin{aligned}
\mathbf{S}_m &= [\mathbf{S}_m^{(l)}\mathbf{S}_m^{(h)}] & N_h \times VK \\
\mathbf{H}_{m,p} &= \begin{bmatrix} \mathbf{H}_{m,p}^{(l)} & 0 \\ 0 & \mathbf{H}_{m,p}^{(h)} \end{bmatrix} & VK \times K \\
\mathbf{W}_m &= \begin{bmatrix} \mathbf{W}^{(l)} & 0 \\ 0 & \mathbf{W}_m^{(h)} \end{bmatrix} & K \times K \\
\mathbf{b}_m &= \begin{bmatrix} \mathbf{b}^{(l)} \\ \mathbf{b}_m^{(h)} \end{bmatrix} & K \times 1
\end{aligned}
$$

$\mathbf{n}_{m,p}$ is complex Gaussian noise vector with i.i.d real and imaginary components with zero mean and variance $\sigma^2\mathbf{I}_{N_h}$, and the noise at each of the sensors is uncorrelated. The resulting chip-rate sampled version of received signal at pth sensor over the time interval $[0, T_l)$ is a N_l vector, and it can be written as:

$$
\begin{aligned}
\mathbf{r}_p &= [\mathbf{r}_{1,p}^{(l)^T} \cdots \mathbf{r}_{M,p}^{(l)^T}]^T + [\mathbf{r}_{1,p}^{(h)^T} \cdots \mathbf{r}_{M,p}^{(h)^T}]^T + \mathbf{n}_p \\
&= \mathbf{S}\mathbf{H}_p\mathbf{W}\mathbf{b} + \mathbf{n}_p, \tag{5.13}
\end{aligned}
$$

where

$$
\begin{aligned}
\mathbf{S} &= [\mathbf{S}^{(l)}\mathbf{S}^{(h)}] & N_l \times VK_v \\
\mathbf{S}^{(h)} &= \mathrm{diag}(\mathbf{S}_1^{(h)}\cdots\mathbf{S}_M^{(h)}) & N_l \times VMK_h \\
\mathbf{H}_p &= \begin{bmatrix} \mathbf{H}_p^{(l)} & 0 \\ 0 & \mathbf{H}_p^{(h)} \end{bmatrix} & VK_v \times K_v \\
\mathbf{H}_p^{(h)} &= diag(\mathbf{H}_{1,p}^{(h)}\cdots\mathbf{H}_{M,p}^{(h)}) & VMK_h \times MK_h \\
\mathbf{W}^{(h)} &= diag(\mathbf{W}_1^{(h)}\cdots\mathbf{W}_M^{(h)}) & MK_h \times MK_h \\
\mathbf{W} &= \begin{bmatrix} \mathbf{W}^{(l)} & 0 \\ 0 & \mathbf{W}^{(h)} \end{bmatrix} & K_v \times K_v \\
\mathbf{b} &= \begin{bmatrix} \mathbf{b}^{(l)} \\ \mathbf{b}^{(h)} \end{bmatrix} & K_v \times 1 \\
\mathbf{b}^{(h)} &= [\mathbf{b}_1^{(h)^T}\cdots\mathbf{b}_M^{(h)^T}]^T & MK_h \times 1.
\end{aligned}
$$

$\mathbf{n}_p$ is complex Gaussian noise vector with i.i.d real and imaginary components with zero mean and variance $\sigma^2\mathbf{I}_{N_l}$, and the noise at each of the sensors is uncorrelated. We make the following assumptions in this chapter:

1. The zero mean noise vector $\mathbf{n}_p, \mathbf{n}_{m,p}$ are temporally and specially white with $E[\mathbf{n}_p\mathbf{n}_p^T] = 0$, $E[\mathbf{n}_p\mathbf{n}_p^H] = \sigma^2\mathbf{I}_{N_l}$ and $E[\mathbf{n}_{m,p}\mathbf{n}_{m,p}^T] = 0$, $E[\mathbf{n}_{m,p}\mathbf{n}_{m,p}^H] = \sigma^2\mathbf{I}_{N_h}$

2. The information bits are i.i.d. with unit energy.

3. All channels $\mathbf{h}_{i,v}$ are LTI within finite duration $[0, VT_c)$.

4. The array gains are all magnitude one: $\|a_{i,v,p}^{(l)}\| = 1$, $\|a_{j,v,p}^{(h)}\| = 1$ for all i, j, p, v. So the array gains represent a simple phase offset with respect to a reference sensor.

5. The V multipath signals for a given user are independent.

5.3 Low-rate Mode LMMSE Space-Time Detection

The low-rate mode LMMSE space-time detector has a bank of the code array/channel matched filters at the front end of the receiver for de-spreading, space filtering and multipath fading cancellation. First, the detectors use a bank of correlators at each antenna, matched to the KV delayed version of multipath spreading codes. The matched filter output for low-rate users at antenna p can be written as:

$$
\mathbf{y}_p^{(l)} = \mathbf{S}^{(l)^T}\mathbf{r}_p. \tag{5.14}
$$

The matched filter output for high-rate users at pth sensor during the time interval $[(m-1)T_h, mT_h)$ can be written as:

$$y_{m,p}^{(h)} = S_m^{(h)^T} r_{m,p}. \tag{5.15}$$

Then the correlator output for each user's vth multipath component at antenna p is weighted by the conjugate of the estimated array/channel coefficients, which is $\hat{h}_{i,v,p}^{(l)} = \hat{a}_{i,v,p}^{(l)} \hat{c}_{i,v}^{(l)}$ for low-rate user i and $\hat{h}_{j,v,p}^{(h)} = \hat{a}_{j,v,p}^{(h)} \hat{c}_{j,v}^{(h)}$ for high-rate user j, to form the code channel matched filter output. The code channel matched filter output at pth sensor for low-rate users can be written as:

$$z_p^{(l)} = \hat{H}_p^{(l)H} y_p^{(l)}. \tag{5.16}$$

The code channel matched filter output at pth sensor for high-rate users' mth subinterval can be written as:

$$z_{m,p}^{(h)} = \hat{H}_{m,p}^{(h)\,H} y_{m,p}^{(h)}. \tag{5.17}$$

Then the multi-path combined information for all P element receiving antennas are going to be summed again to form the final code array/channel matched filter output. Let's define the real part of the summation of P low-rate users' code channel matched filter output as:

$$x^{(l)} = Re\left[\sum_{p=1}^{P} z_p^{(l)}\right] = Re\left[\sum_{p=1}^{P} \hat{H}_p^{(l)H} y_p^{(l)}\right]. \tag{5.18}$$

The real part of the summation of P high-rate users' code channel matched filter output is:

$$x_m^{(h)} = Re\left[\sum_{p=1}^{P} z_{m,p}^{(h)}\right] = Re\left[\sum_{p=1}^{P} \hat{H}_{m,p}^{(h)\,H} y_{m,p}^{(h)}\right]. \tag{5.19}$$

Since rate difference exists between the different users, the code array/channel matched filter output for high-rate users, $x_m^{(h)}$, is put into the memory to form the MVK_h vector $x^{(h)}$, where $x^{(h)} = [x_1^{(h)^T} \cdots x_M^{(h)^T}]^T$. This MVK_h vector is going to be jointly processed with the low rate users' code array/channel matched filter output vector $x^{(l)}$. The structure of the code array/channel matched filter is shown

in Figure 5.1. All together, the code array/channel matched filter output during the time interval $[0, T_l)$ can be written as:

$$\mathbf{x} = [\mathbf{x}^{(l)T} \mathbf{x}^{(h)T}]^T = Re \left[\sum_{p=1}^{P} \hat{\mathbf{H}}_p^H \mathbf{S}^T \mathbf{r}_p \right] = \mathcal{R} \mathbf{W} \mathbf{b} + \mathbf{n}_x, \qquad (5.20)$$

where

$$\begin{aligned} \mathcal{R} &= Re \left[\sum_{p=1}^{P} \hat{\mathbf{H}}_p^H \mathbf{R} \mathbf{H}_p \right], \\ \mathbf{R} &= \mathbf{S}^T \mathbf{S}, \\ \mathbf{n}_x &= Re \left[\sum_{p=1}^{P} \hat{\mathbf{H}}_p^H \mathbf{S}^T \mathbf{n}_p \right], \end{aligned}$$

$\mathbf{n}_x$ is a Gaussian noise vector with zero mean and covariance matrix $\sigma^2 \mathcal{R}$.

All the linear processor is going to work on the data of the code array/channel matched filter output. Fig.5.2 shows the system structure of low-rate mode LMMSE space-time receiver. After the code array/channel matched filtering, the minimum mean-squared error detection is used to suppress the MAI by minimizing the mean-square error between the output data vector and the transmitted information vector.

Suppose $\mathbf{b}$ is the transmitted information vector. $\mathbf{x}$ is the $(K_l + MK_h)$ vector which is the input to the linear transformation part. We need to find the $\mathbf{T}_{opt}$ such that

$$\mathbf{T}_{opt} = \arg \min_{\mathbf{T}} \{ \mathbf{E}[\|\mathbf{T}\mathbf{x} - \mathbf{W}\mathbf{b}\|^2] \}. \qquad (5.21)$$

Suppose $\Theta = \mathbf{W}\mathbf{b}$. We assume that all the components of the random multiuser information bits $\mathbf{b}$ are independent and identically distributed with zero mean and unit variance, the random vector Θ and noise vector $\mathbf{n}_x$ are statistically independent. The solution to equation 5.21 can be found in [17][15], which is

$$\begin{aligned} \hat{\Theta}_{LMMSE} &= \mathbf{E}(\Theta) + \Sigma_{\Theta x} \Sigma_{xx}^{-1} (\mathbf{x} - \mathbf{E}(\mathbf{x})) \\ &= \Sigma_{\Theta x} \Sigma_{xx}^{-1} \mathbf{x} \\ &= (\mathcal{R} + \sigma^2 \mathbf{W}^{-2})^{-1} \mathbf{x}, \qquad (5.22) \end{aligned}$$

where $\Sigma_{\Theta x} = \mathbf{W}^2 \mathcal{R}, \Sigma_{xx} = \mathcal{R} \mathbf{W}^2 \mathcal{R} + \sigma^2 \mathcal{R}$. So the solution of $\mathbf{T}_{opt}$ is going to be

$$\mathbf{T}_{opt} = (\mathcal{R} + \sigma^2 \mathbf{W}^{-2})^{-1}. \qquad (5.23)$$
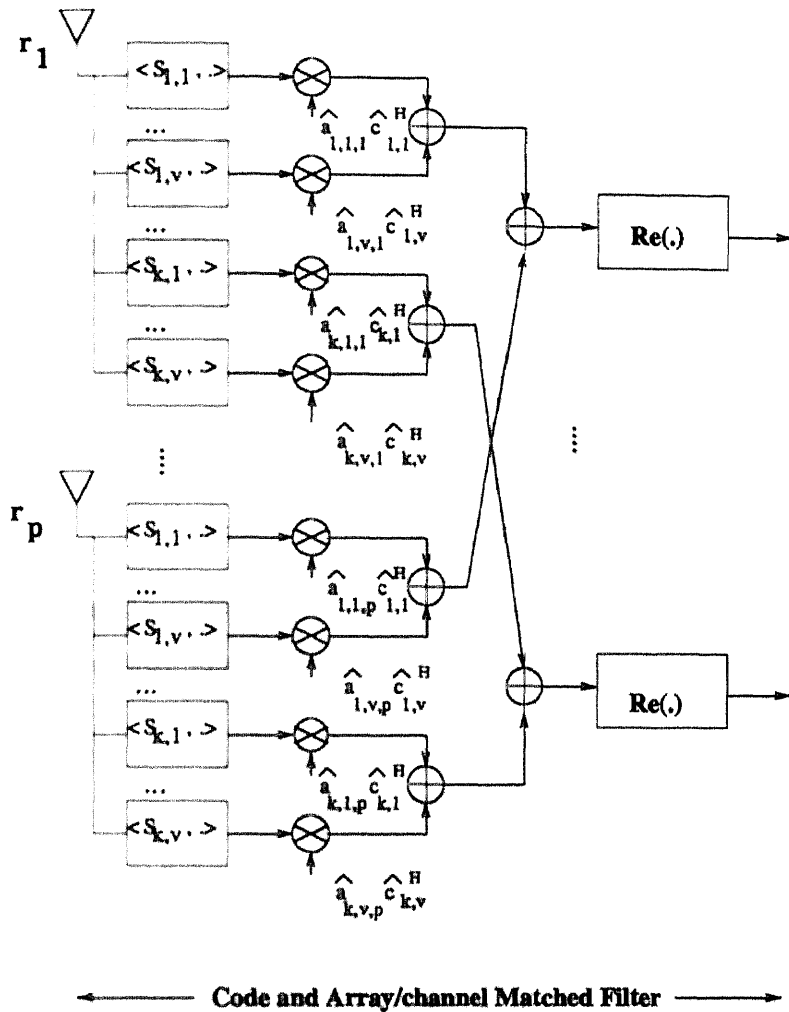
Figure 5.1 Code and array/channel matched filter for single rate CDMA
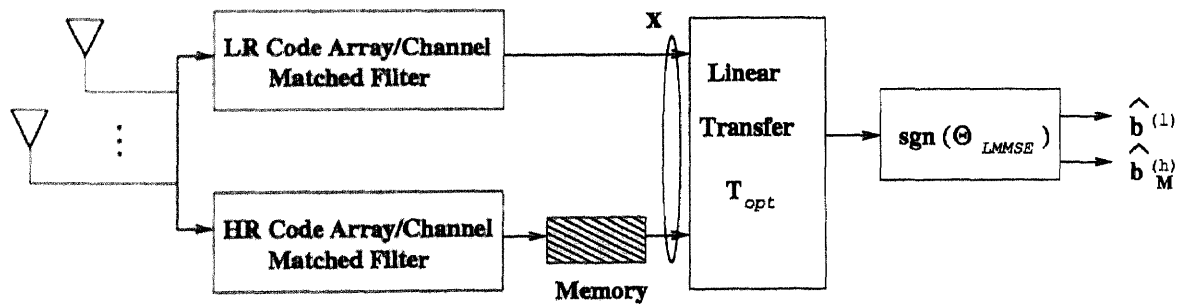
Figure 5.2 Low-rate mode LMMSE space-time receiver

The final decision is made by $\hat{b} = \mathrm{sgn}(\hat{\Theta}_{LMMSE})$ and the $\hat{\Theta}_{LMMSE}$ can be get from

$$\hat{\Theta}_{LMMSE} = \mathbf{T}_{opt}\mathbf{x} = \mathbf{T}_{opt}\mathcal{R}\mathbf{W}\mathbf{b} + \mathbf{n}_\Theta, \tag{5.24}$$

where $\mathbf{n}_\Theta$ is the Gaussian noise vector with zero mean and covariance matrix $\sigma^2 \mathbf{T}_{opt}\mathcal{R}\mathbf{T}_{opt}^T$. In order to find the estimation error of $\hat{\Theta}_{MMSE}$, we can further decompose the $\hat{\Theta}_{MMSE}$ as follow,

$$
\begin{aligned}
\hat{\Theta}_{MMSE} &= \mathbf{T}_{opt}\mathbf{x} = (\mathcal{R} + \sigma^2\mathbf{W}^{-2})^{-1}\mathbf{x} \\
&= (\mathcal{R} + \sigma^2\mathbf{W}^{-2})^{-1} \cdot (\mathcal{R}\mathbf{W}\mathbf{b} + \mathbf{n}_x) \\
&= \Theta - \sigma^2\mathbf{T}_{opt}\mathbf{W}^{-1}\mathbf{b} + \mathbf{T}_{opt}\mathbf{n}_x \\
&= \Theta + \mathbf{e},
\end{aligned}
\tag{5.25}
$$

where the estimation error, $\mathbf{e} = -\sigma^2\mathbf{T}_{opt}\mathbf{W}^{-1}\mathbf{b} + \mathbf{T}_{opt}\mathbf{n}_x$, contains both bias and noise components. The error covariance matrix of MMSE can be found to be equal to

$$E\left\{(\Theta - \hat{\Theta})(\Theta - \hat{\Theta})^T\right\} = cov(\mathbf{e}) = \sigma^2\mathbf{T}_{opt}. \tag{5.26}$$

The bit error rate for kth LR/HR user average over the interferers' bits is

$$P_k(\sigma) = \frac{1}{2^{K_v-1}} \sum_{\mathbf{b}\in\{\pm1\}^{K_v}, b_k=1} Q\left(\frac{(\mathbf{T}_{opt}\mathcal{R}\mathbf{W}\mathbf{b})_k}{\sigma\sqrt{(\mathbf{T}_{opt}\mathcal{R}\mathbf{T}_{opt}^T)_{k,k}}}\right). \tag{5.27}$$

Because of the convexity of the mean-squared error expression in equation 5.21, the MMSE solution in equation 5.23 represents a global minimum. But the explicit knowledge of the array and channel coefficients are needed in code array/channel matched filter part. Another drawback of this MMSE detector is that the estimation of the channel and array coefficients cannot be incorporated into the adaptive algorithm for obtaining $\mathbf{T}_{opt}$. The estimates must be explicitly obtained using some other means (like a training or pilot signal). In order to solve the problems, an alternative approach which is called is called type 2 low-rate mode LMMSE space-time detection is proposed here. The system construction of the type 2 receiver is shown in Figure 5.3. In this receiver, the linear MMSE detection is cascaded together with array/channel weighting and combining.
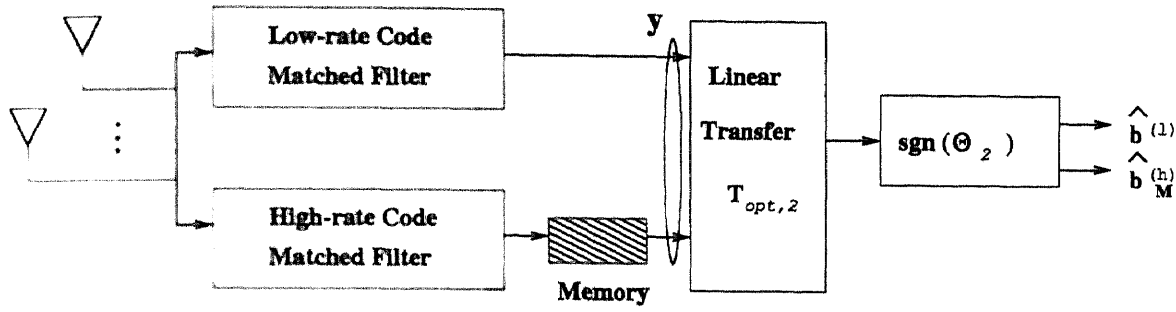
Figure 5.3 Type 2 low-rate mode LMMSE space-time receiver

In this receiver, the code matched filter output for high-rate users at pth sensor is put into the memory every T_h. At pth sensor, at the end of the time interval $[0, T_l)$, the code matched filter output can be written as:

$$\mathbf{y}_p = [\mathbf{y}_p^{(l)T} \mathbf{y}_{1,p}^{(h)T} \cdots \mathbf{y}_{M,p}^{(h)\ T}]^T$$

$$= \mathbf{S}^T \mathbf{r}_p = \mathbf{R}\mathbf{H}_p\mathbf{W}\mathbf{b} + \mathbf{S}^T\mathbf{n}_p. \tag{5.28}$$

The code matched filter output of all P sensors can be stacked into a size PVK_v vector. This vector can be written as:

$$\mathbf{y} = [\mathbf{y}_1^T \cdots \mathbf{y}_P^T]^T = \mathcal{R}_2\mathbf{H}\mathbf{W}\mathbf{b} + \mathbf{n}_y, \tag{5.29}$$

where

$$
\begin{array}{lll}
\mathcal{R}_2 & = & \mathbf{I}_P \otimes \mathbf{R} \qquad & PVK_v \times PVK_v \\
\mathbf{H} & = & [\mathbf{H}_1^T \cdots \mathbf{H}_P^T]^T & PVK_v \times K_v \\
\mathbf{n}_y & = & [(\mathbf{S}^T\mathbf{n}_1)^T \cdots (\mathbf{S}^T\mathbf{n}_P)^T]^T & PVK_v \times 1
\end{array}
$$

(where $\otimes$ denotes the Kronecker product of two matrices).

After the code matched filter, the linear multiuser detector, together with array/channel weighting and combining, is trying to minimize the MSE between the output data and the transmitted information. That means we try to find $\mathbf{T}_{opt,2}$ such that

$$\mathbf{T}_{opt,2} = \arg\min_{\mathbf{T}} E[\|\mathbf{T}\mathbf{y} - \mathbf{b}\|^2] \tag{5.30}$$

It can be found that the solution of equation 5.30 is:

$$
\begin{aligned}
\mathbf{T}_{opt,2} & = E[\mathbf{b}\mathbf{y}^H](E[\mathbf{y}\mathbf{y}^H])^{-1} \\
& = \mathbf{W}\mathbf{H}^H\mathcal{R}_2\Sigma_{\mathbf{yy}}^{-1}, \tag{5.31}
\end{aligned}
$$

where $\Sigma_{\mathbf{yy}} = E[\mathbf{y}\mathbf{y}^H]$, and $\mathbf{T}_{opt,2}$ can be written as:

$$\mathbf{T}_{opt,2} = \mathbf{W}\mathbf{H}^H(\mathbf{H}\mathbf{W}^2\mathbf{H}^H\mathcal{R}_2 + \sigma^2\mathbf{I}_{PVK_v})^{-1} \tag{5.32}$$

Noting that we suppose the matrix $\mathcal{R}_2$ is full rank when $\mathbf{R}$ is full rank, so that the matrix inversion exist. If $\mathcal{R}_2$ is not full rank, the pseudoinverse based on the singular-value decomposition can be used. So the decision statistic can be written as:

$$\hat{\Theta}_2 = \mathbf{T}_{opt,2}\mathbf{y} \tag{5.33}$$

The final decision is made by taking the sign of the real part of $\hat{\Theta}_2$, which means $\hat{b} = \text{sgn}\{Re[\hat{\Theta}_2]\}$. An attractive feature of this type 2 LMMSE space-time receiver is that it can be obtained adaptively using the data driven adaptive method which we introduced in chapter 3, or the other well-known adaptive algorithms such as least-mean squares or recursive least squares with only the knowledge of the desired user's signature and timing information. When the recursive least square adaptive method is used, this type two space-time receiver can be used to jointly detect the desired user's information and estimate the channel parameters [59][60][61].

5.4 Two-stage Space-time Detection for Dual-rate CDMA Systems

The objective of proposing another two-stage LMMSE space-time detector is that: firstly, the detectors should be able to work at the mode specified by the symbol interval according to the QoS requirements; secondly, the previous detector which is working over the maximum symbol interval has a large dimension and thus high computational complexity; thirdly, there exists $(M - 1)$ bits processing delay for high-rate users. In order to solve these problems, the two-stage LMMSE space-time detection is proposed. The system construction is shown in Fig. 5.4. Inside the code array/channel matched filter, the processing interval is all set to $[(m - 1)T_h, mT_h)$. After one processing interval, low-rate users only get the partial received information and high-rate users can get the whole received bit information. The delay which is induced by the low rate receiver for high rate users is eliminated. In addition, the cross correlation matrix of the signature sequences, $\mathbf{R}_m = \mathbf{S}_m^T \mathbf{S}_m$, has the smaller size of $(K_h + K_l) \times (K_h + K_l)$ comparing to the size of $(MK_h + K_l) \times (MK_h + K_l)$. The two-stage LMMSE space-time receiver works based on the following principles over $[0, T_l)$:
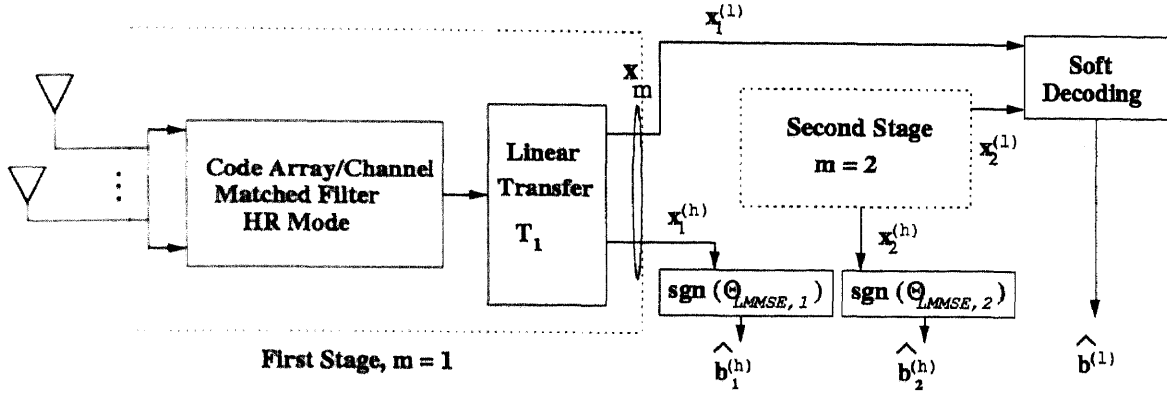
Figure 5.4 Two-stage space-time LMMSE receiver

1. For the first $(M-1)$ subinterval where $1 \leq m \leq (M-1)$:

   (a) match the received information of $[(m-1)T_h, mT_h)$ with code array/channel matched filter bank

   (b) perform the LMMSE estimation-based detection for the code array/channel matched filter output

   (c) make the decision for high-rate users for mth subinterval

   (d) put low-rate users' partial detected information into memory

2. For the last subinterval M where $m = M$:

   (a) the detection of high-rate users follow $1(a)$ to $1(c)$

   (b) make final decision for low-rate users using maximum ratio combining of the M detected low-rate users' partial information

The code array/channel matched filter output $\mathbf{x}_m$ (m corresponding to the mth subinterval of T_l) is a $(K_l + K_h)$ vector.

$$
\begin{aligned}
\mathbf{x}_m &= Re\left[\sum_{p=1}^{P} \hat{\mathbf{H}}_{m,p}^{H} \mathbf{R}_m \mathbf{H}_{m,p}\right] \mathbf{W}_m \mathbf{b}_m + Re\left[\sum_{p=1}^{P} \hat{\mathbf{H}}_{m,p}^{H} \mathbf{S}_m^{T} \mathbf{n}_{m,p}\right] \\
&= \mathcal{R}_m \mathbf{W}_m \mathbf{b}_m + \mathbf{n}_{x,m},
\end{aligned}
\tag{5.34}
$$

where

$$
\mathcal{R}_m = Re\left[\sum_{p=1}^{P} \hat{\mathbf{H}}_{m,p}^{H} \mathbf{R}_m \mathbf{H}_{m,p}\right],
$$

$$\mathbf{n}_{x,m} = Re\left[\sum_{p=1}^{P} \hat{\mathbf{H}}_{m,p}^{H} \mathbf{S}_{m}^{T} \mathbf{n}_{m,p}\right].$$

The definition of $\mathbf{H}_{m,p}, \mathbf{W}_m, \mathbf{b}_m$ can be found in the equation 5.12. The high-rate cross correlation matrix is a $(K_l + K_h) \times (K_l + K_h)$ matrix which can be written as:

$$\mathbf{R}_m = \left[\begin{array}{cc} \mathbf{R}_{ll,m} & \mathbf{R}_{lh,m} \\ \mathbf{R}_{hl,m} & \mathbf{R}_{hh} \end{array}\right], \tag{5.35}$$

where $\mathbf{R}_{ll,m}$ is a $K_l \times K_l$ cross-correlation matrix of K_l low-rate users within mth subinterval of a time period T_l, $R_{ll,m}(i,j) = \int_{(m-1)T_h}^{mT_h} s_i^{(l)}(t)s_j^{(l)}(t)dt.(i,j = 1, \cdots, K_l)$. $\mathbf{R}_{lh,m}$ is a $K_l \times K_h$ cross-correlation matrix of K_l low-rate users with K_h high-rate users during a time period T_h and $\mathbf{R}_{hl,m} = \mathbf{R}_{lh,m}^T$ with $R_{lh,m}(i,j) = \int_{(m-1)T_h}^{mT_h} s_i^{(l)}(t)s_j^{(h)}(t - (m-1)T_h)dt.(i = 1, \cdots, K_l; j = 1, \cdots, K_h)$. $\mathbf{R}_{hh}$ is a $K_h \times K_h$ cross-correlation matrix of K_h high-rate users within a time period T_h with $R_{hh}(i,j) = \int_{0}^{T_h} s_i^{(h)}(t)s_j^{(h)}(t)dt.(i,j = 1, \cdots, K_h)$. $\mathbf{n}_{x,m}$ is a zero mean Gaussian noise with covariance matrix $\sigma^2 \mathcal{R}_m$. According to MMSE criteria, the linear transformation for the first stage can be found by:

$$\begin{aligned} \mathbf{T}_{opt,m} &= \arg\min_{\mathbf{T}_m}\{E[\|\mathbf{T}_m\mathbf{x}_m - \mathbf{W}_m\mathbf{b}_m\|^2]\}, \\ \mathbf{T}_{opt,m} &= (\mathcal{R}_m + \sigma^2\mathbf{W}_m^{-2})^{-1}. \end{aligned} \tag{5.36}$$

The decision statistic for the first stage is:

$$\hat{\Theta}_{MMSE,m} = \mathbf{T}_{opt,m}\mathbf{x}_m = \mathbf{T}_{opt,m}\mathcal{R}_m\mathbf{W}_m\mathbf{b}_m + \mathbf{T}_{opt,m}\mathbf{n}_{x,m}. \tag{5.37}$$

For high-rate users, we can decode the information at every subinterval $[(m-1)T_h, mT_h)$ at both stages. The estimated information bit of jth high-rate user for both stages is:

$$\hat{b}_{j,m}^{(h)} = \mathrm{sgn}\{(\hat{\Theta}_{MMSE,m})_{K_l+j}\}. \tag{5.38}$$

The probability of error for high-rate user j for both stages averaged over the interferers' bits is

$$P_{j,m}^{(h)}(\sigma) = \frac{1}{2^{K-1}} \sum_{\mathbf{b}_m \in \{\pm 1\}^K, b_{j,m}=1} Q\left(\frac{(\mathbf{T}_{opt,m}\mathcal{R}_m\mathbf{W}_m\mathbf{b}_m)_{K_l+j}}{\sigma\sqrt{(\mathbf{T}_{opt,m}\mathcal{R}_m\mathbf{T}_{opt,m}^T)_{K_l+j,K_l+j}}}\right). \tag{5.39}$$

For low-rate users, only the partial information can be obtained during the first stage. At the last stage, a soft decoding rule is going to be applied for low-rate users. The soft decoding rule which is used here is the maximum ratio combining. So the decision of ith low-rate user is made by taking the sign of $\sum_{m=1}^{M} \left(\frac{1}{(T_{opt,m} \mathcal{R}_m T_{opt,m}^T)_{i,i}} (\hat{\Theta}_{MMSE,m})_k \right)$. There is also an alternative approach for two-stage LMMSE space-time detection and is called type 2 two-stage LMMSE space-time detection. For this alternative approach, the code matched filter output at pth sensor, at the end of the time interval $[(m-1)T_h, mT_h)$, can be written as:

$$
\begin{aligned}
\mathbf{y}_{m,p} &= [\mathbf{y}_{m,p}^{(l)}{}^T \mathbf{y}_{m,p}^{(h)}{}^T]^T \\
&= \mathbf{S}_m^T \mathbf{r}_{m,p} = \mathbf{R}_m \mathbf{H}_{m,p} \mathbf{W}_m \mathbf{b}_m + \mathbf{S}_m^T \mathbf{n}_{m,p}
\end{aligned}
\tag{5.40}
$$

The code matched filter output of all P sensors can be stacked in to a size PVK vector. This vector can be written as:

$$
\mathbf{y}_m = [\mathbf{y}_{m,1}^T \cdots \mathbf{y}_{m,P}^T]^T = \mathcal{R}_{m,2} \mathbf{H}_m \mathbf{W}_m \mathbf{b}_m + \mathbf{n}_{y,m}
\tag{5.41}
$$

where

$$
\begin{aligned}
\mathcal{R}_{m,2} &= \mathbf{I}_P \otimes \mathbf{R}_m & PVK \times PVK \\
\mathbf{H}_m &= [\mathbf{H}_{m,1}^T \cdots \mathbf{H}_{m,P}^T]^T & PVK \times K \\
\mathbf{n}_{y,m} &= [(\mathbf{S}_m^T \mathbf{n}_{m,1})^T \cdots (\mathbf{S}_m^T \mathbf{n}_{m,P})^T]^T & PVK \times 1
\end{aligned}
$$

The definition of $\mathbf{W}_m, \mathbf{b}_m, \mathbf{n}_{m,p}$ can be found in equation 5.12. After the code matched filter, the linear multiuser detector, together with array/channel weighting and combining, is trying to minimize the MSE between the output data and the transmitted information. That means we try to find $\mathbf{T}_{opt2,m}$ such that

$$
\mathbf{T}_{opt2,m} = \arg\min_{\mathbf{T}_m} E[\|\mathbf{T}_m \mathbf{y}_m - \mathbf{b}_m\|^2]
\tag{5.42}
$$

It can be found that the solution of equation 5.42 is:

$$
\begin{aligned}
\mathbf{T}_{opt2,m} &= E[\mathbf{b}_m \mathbf{y}_m^H](E[\mathbf{y}_m \mathbf{y}_m^H])^{-1} \\
&= \mathbf{W}_m \mathbf{H}_m^H \mathcal{R}_{m,2} \Sigma_{y_m}^{-1},
\end{aligned}
\tag{5.43}
$$

where $\Sigma_{y_m} = E[\mathbf{y}_m \mathbf{y}_m^H]$, and $\mathbf{T}_{opt2,m}$ can be written as:

$$
\mathbf{T}_{opt2,m} = \mathbf{W}_m \mathbf{H}_m^H (\mathbf{H}_m \mathbf{W}_m^2 \mathbf{H}_m^H \mathcal{R}_{m,2} + \sigma^2 \mathbf{I}_{PVK})^{-1}
\tag{5.44}
$$

Noting that we suppose the matrix $\mathcal{R}_{m,2}$ is full rank when $\mathbf{R}_m$ is full rank, so that the matrix inversion exist. If $\mathcal{R}_{m,2}$ is not full rank, the pseudoinverse based on the singular-value decomposition can be used. So the decision statistic can be written as:

$$\hat{\Theta}_{m,2} = \mathbf{T}_{opt2,m}\mathcal{Y}_m \tag{5.45}$$

The final decision for high-rate users is made by taking the sign of the real part of the $\hat{\Theta}_{m,2}$, which means for high-rate user j, the estimated information for the mth subinterval of T_l is $\hat{b}_{j,m}^{(h)} = \text{sgn}\{Re[(\hat{\Theta}_{m,2})_{K_l+j}]\}$. At the last stage, the low-rate users' information is going to be combined to make the final decision. The soft decoding rule which is used here is the maximum ratio combining. So the decision of ith low-rate user is made by taking the sign of the real part of $\sum_{m=1}^{M} \left(\frac{1}{(\mathbf{T}_{opt2,m}\mathcal{R}_{2,m}\mathbf{T}_{opt2,m}^T)_{i,i}}(\mathbf{T}_{opt2,m}\mathcal{Y}_m)_i \right)$.

The type 2 two-stage LMMSE space-time receiver can be implemented adaptively without the knowledge of the spatial signatures, but the desired user's signature sequence and timing information is still required.

5.5 Numerical Analysis

The simulation results of the low-rate mode LMMSE space-time receiver and two-stage LMMSE space-time receiver for VSL access method are shown in this part. We compare the performance of 1-stage and 2-stage space-time MMSE receiver for the low-rate users and high-rate users. We also compare the performance between the VSL and MC access methods of space-time receivers. Performance of different combining methods, which are Maximum Ratio Combining (MRC) and MMSE combining, for the multi-rate space-time receivers are also compared.

A direct sequence CDMA systems is considered. There are two different data rates with rate ratio M. For low-rate users, the repetition coding scheme is used, which means that the codes repeat themselves M times during one low-rate bit interval. V is the multipath delay spread. P is the number of sensors. For a given maximum symbol period T_l, the directions of arrival for each users' multipath components $\alpha_{i,v}^{(l)}$, $\alpha_{j,v}^{(h)}$ are supposed to be uniformly distributed random variables over $[0, 2\pi]$.

Firstly, the performance of low-rate mode and 2-stage space-time receivers for low-rate user 1 is compared in Fig. 5.5 and Fig. 5.6. Since the repetition coding scheme is used for low-rate users, the performance for the low-rate mode and 2-stage space-time receiver for low-rate users are same. The performance of low-rate mode and 2-stage space-time receivers for high-rate user 3 is compared in Fig. 5.7. For the high-rate users, the performance of 2-stage space-time receiver is not as good as low-rate mode space-time receiver in order to get lower computational complexity and eliminate processing delay. It can be seen from the above three figures that there is significant performance gain when $P = 3, V = 3$ comparing to when $P = 1, V = 3$ which corresponding to time domain only receiver.

Secondly, the performance between the VSL and MC access methods of low-rate mode space-time receiver for different users is compared. Fig. 5.9 is the performance of the high-rate user 1. It can be seen that for both decorrelator and LMMSE detector, the performance for MC system is worse than VSL system, since for high-rate users, the MC access method is going to introduce self interference because of the multiple channels that are assigned to one user. But for the low-rate user, whose performance is shown in Fig. 5.8, the MC access method does not always perform worse than VSL access method for the MMSE detection.

Thirdly, MMSE combining is compared with the MRC for VSL access method with $P = 2, V = 1$. For a single-rate system with users uniformly distributed in space, MMSE combining performance only slightly better than MRC. The improvement associated with MMSE is mainly due to the coherent combining, since the number of degrees of freedom is small compared with the number of users. The difference in performance becomes much more pronounced in multi-rate system. The coexistence of high-rate and low-rate users creates the received power imbalance similar to the near-far problem. MMSE filter can suppress the greater interference caused by high-rate users, whereas the maximum ratio filter does not account for power variation for users with different data rates. The performance comparison of MRC and MMSE combining is shown in Fig. 5.10, Fig. 5.11 and Fig. 5.12. The difference between the MMSE and MRC increases with increasing the rate-ratio M.
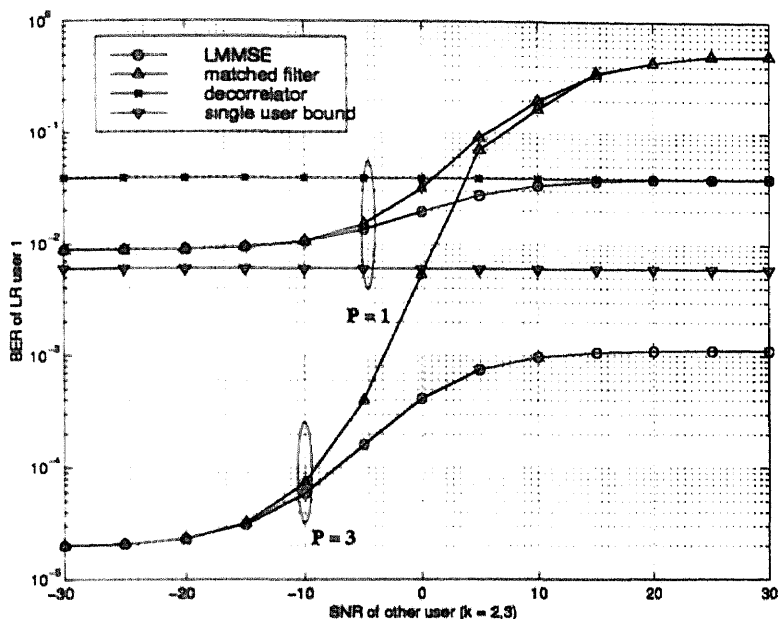
Figure 5.5 Performance of low-rate mode space-time receiver of low-rate user 1, three user VSL system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$, SNR of low-rate user 1 $SNR1 = 8dB$
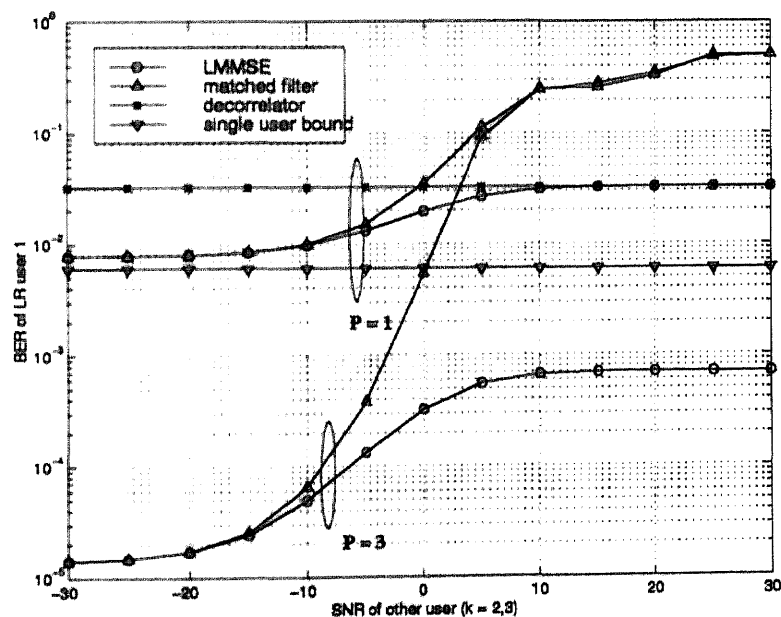
Figure 5.6 Performance of 2-stage space-time receiver of low-rate user 1, 3 user VSL system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$, SNR of low-rate user 1 $SNR1 = 8dB$
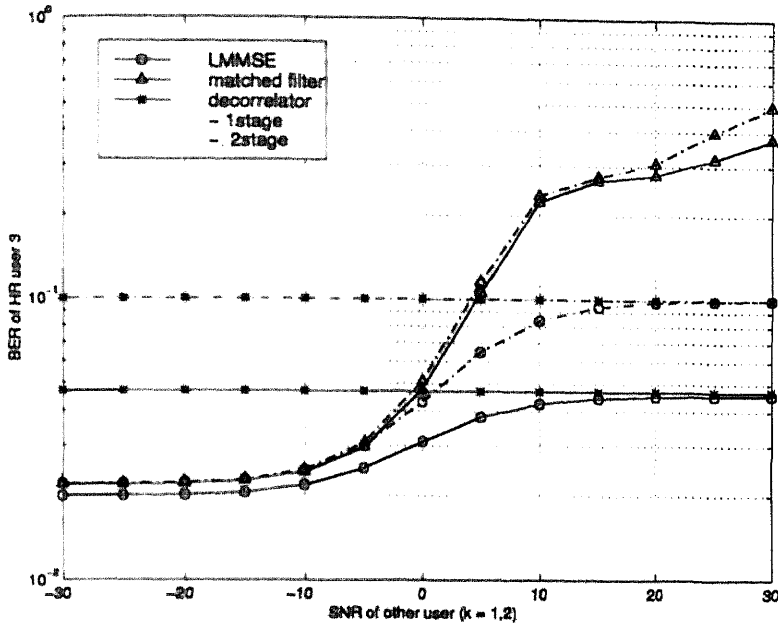
Figure 5.7 Performance comparison of low-rate mode and 2stage space-time (P=1, V=3) receiver of high-rate user 1, 3 user VSL system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7$, SNR of high-rate user 1 $SNR3 = 8dB$
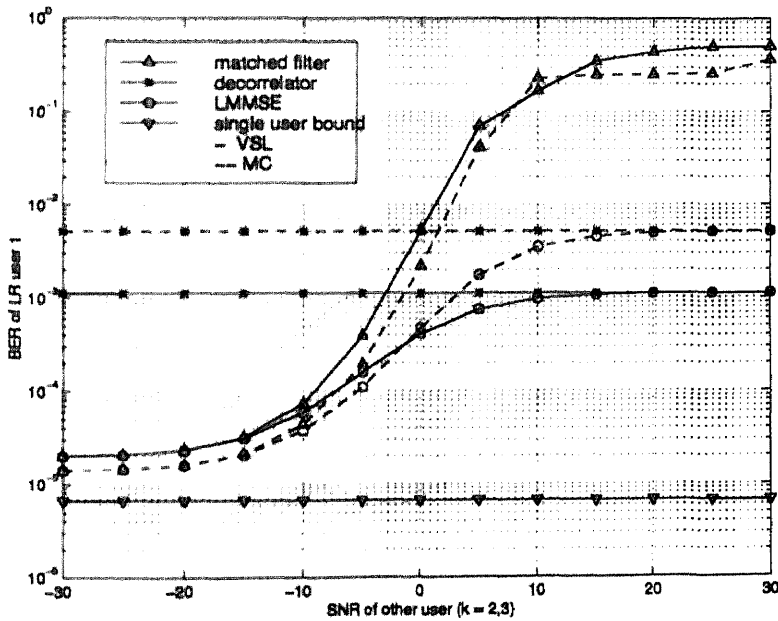
Figure 5.8 Performance comparison of VSL & MC access methods for low-rate mode space-time $(P = 3, V = 3)$ receiver for low-rate user 1, 3 user system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7, L_{mc} = 7, SNR1 = 8dB$
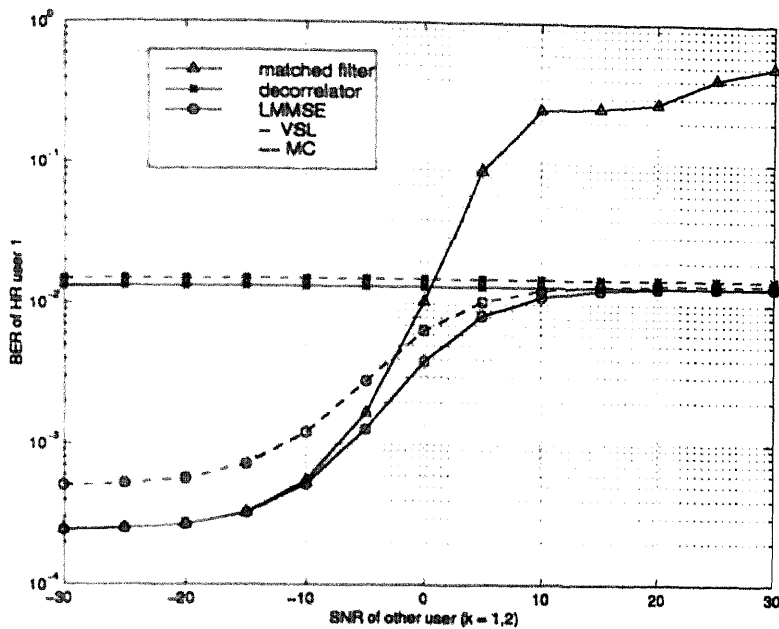
Figure 5.9 Performance comparison of VSL & MC access methods for low-rate mode space-time $(P = 3, V = 3)$ receiver for high-rate user 3, 3 user system, $K_l = 2, K_h = 1, M = 2, L_l = 14, L_h = 7, L_{mc} = 7, SNR3 = 8dB$
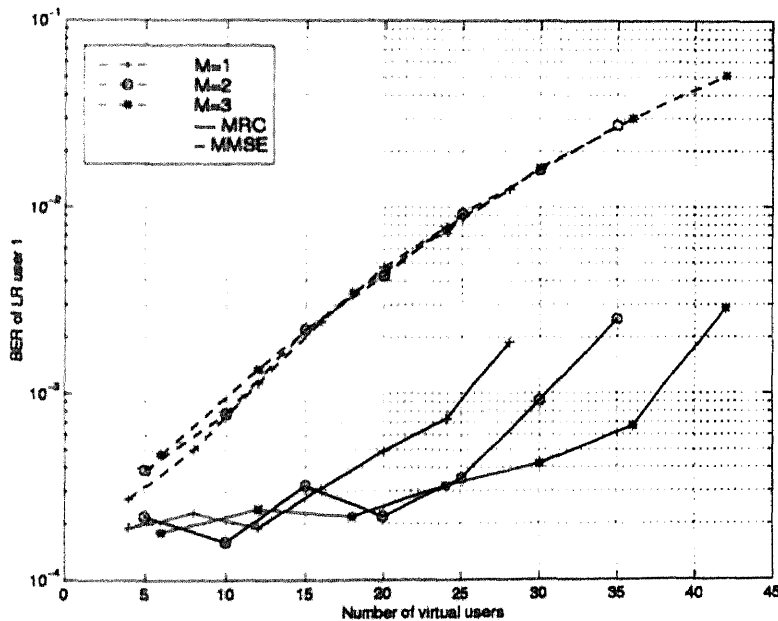
Figure 5.10 Performance comparison of MRC and MMSE combining for low-rate mode space-time $(P = 2, V = 1)$ receiver of low-rate user 1. VSL system, $L_h = 31, 25\%$ high-rate users, 75% low-rate users, desired user $SNR = 8dB$
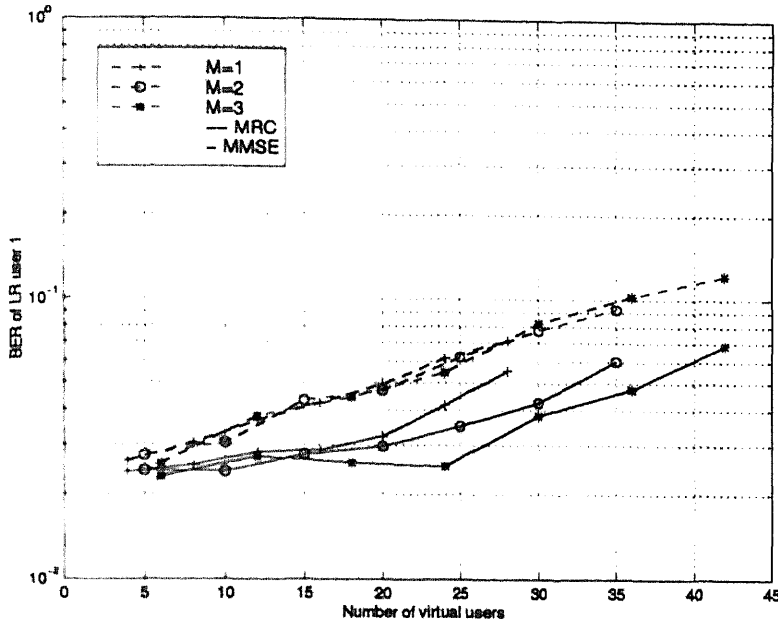
Figure 5.11 Performance comparison of MRC and MMSE combining for low-rate mode space-time ($P = 2, V = 1$) receiver of low-rate user 1. VSL system, $L_h = 31,25\%$ high-rate users, 75% low-rate users, desired user $SNR = 3dB$
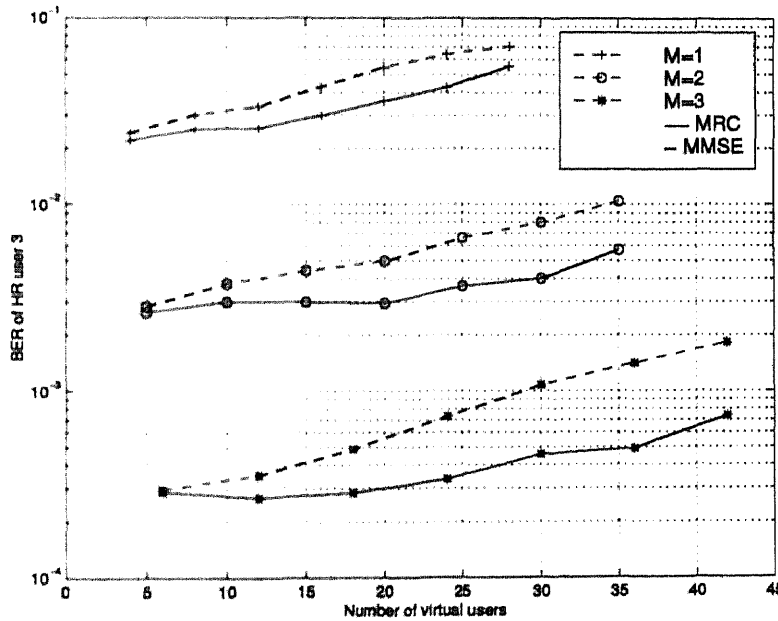
Figure 5.12 Performance comparison of MRC and MMSE combining for low-rate mode space-time ($P = 2, V = 1$) receiver of low-rate user 1. VSL system, $L_h = 31,25\%$ high-rate users, 75% low-rate users, desired user $SNR = 3dB$

CHAPTER 6

DECISION FEEDBACK INTERFERENCE CANCELLATION FOR MULTI-RATE CDMA SIGNALS BSED ON MMSE CRITERION

6.1 Introduction

In the previous chapter, our approach to multiuser detections for multi-rate CDMA signals is to use linear MMSE multiuser detection in combining with other signal processing techniques. Increasing the number of dimensions available through diversity leads to the algorithms which have good performance with moderate complexity. In addition, diversity provides robustness in the presence of low SNR and fading. The low-rate mode LMMSE space-time receiver and the two-stage LMMSE space-time receiver are proposed in chapter 5. They can successfully detect the multi-rate users' information simultaneously under certain conditions. Namely, good performance requires the number of available dimensions (e.g. antennas times multi-path) exceeds the number of users and the received powers must be disparate (i.e. greater than $6dB$ difference).

To over come the limitations imposed by linear receivers, we are going to study the Decision-Feedback Detection (DFD) in this chapter. The decision-feedback detection for the single rate CDMA signals was studied in [62][63]. The multistage detection for single-rate CDMA signals was studied in [4][7]. The successive Interference cancellation for single-rate CDMA signals was studied in [64]. The decision-feedback decorrelating detector for dual-rate CDMA signals was studied in [65]. The DFD can be viewed as a two-stage detector where decisions made by the first stage are used to cancel the interference successively in the second stage. The canceler does not require remodulation, which enhances robustness in the presence of multi-path. To reduce the effect of error propagation, the users to be demodulated are ordered with respect to decreasing received power, so that weak users benefit from reliable cancellation of stronger users' interference. For the multi-rate communications, if the symbol energy is fixed for different users with different data rates, the high rate users have high power in it's signal. This feature naturely invites the decision feedback interference cancellation scheme into multi-rate receiver design. The emphasis of our study in this chapter is on a group-wise serial decision feedback detection for multi-rate CDMA signals. The proposed detector is working based on the MMSE

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criterion and has the promise of DFD which has superior performance comparing to the linear MMSE receiver at the cost of a moderate increase in complexity. In group-wise detection, the users are grouped according to the processing gain they used, i.e., users with the lowest processing gain are detected first neglecting the presence of the other users in the system. Interference between the groups is cancelled in a successive order. The bit error rate of the groupwise MMSE DFD is compared with the multi-rate linear MMSE detector and multi-rate decorrelating detector. The results show that the groupwise MMSE DFD yields better performance than multi-rate linear MMSE detector and multi-rate decorrelating detector. It is also shown that when the system is highly loaded, the groupwise MMSE DFD can offer better performance than multi-rate linear MMSE detector and multi-rate conventional detector.

6.2   Groupwise Multi-rate System Description

In a Variable Spreading Length (VSL) multi-rate CDMA system, all the active users are classified into G groups. The group G users have the lowest data rate $\frac{1}{T_G}$, where T_G is the bit interval of group G users, and there are total number of K_G users in this group. The group 1 users have the highest data rate $\frac{1}{T_1} = \frac{m_1}{T_G}$, where T_1 is the bit interval of group 1 users, and $m_1 = \frac{T_G}{T_1} = 2^{G-1}$ is the rate ratio between the highest bit rate and the lowest bit rate. There are total number of K_1 users in group 1. In general, for group g, we assume that the integer $m_g = \frac{T_G}{T_g}$ represents the rate ratio between the rate of group g and the lowest bit rate. It is supposed that the group number g and the corresponding m_g has the following relationship, $m_g = 2^{G-g}$, where $g = 1, 2, \cdots, G$. There are total $K = K_1 + K_2 + \cdots + K_G$ users which are grouped into G groups according to different bit rates. In a VSL CDMA system, users with different data rates will have different processing gain. Total G groups users have G different spreading factors, which are $L_1 = \frac{T_1}{T_c}, L_2 = \frac{T_2}{T_c} = 2L_1, \cdots, L_G = \frac{T_G}{T_c} = m_1 L_1$. The signature waveform of kth user of group g, $s_k^{(g)}(t)$, has the following property: $\int_0^{T_g} s_k^{(g)^2}(t)dt = 1$. In the interval $[0, T_G)$, each group G user transmit one bit, while each group g $(g < G)$ user transmit $m_g = 2^{G-g}$ bits.

The total base band received signal during the time interval $[0, T_G)$ can be written as:

$$
\begin{aligned}
r(t) &= \sum_{i=1}^{K_1} \sum_{m=1}^{m_1} \sqrt{w_i^{(1)}} b_{i,m}^{(1)} s_i^{(1)}(t - (m-1)T_1 - \tau_i^{(1)}) \\
&+ \sum_{j=1}^{K_2} \sum_{m=1}^{m_2} \sqrt{w_j^{(2)}} b_{j,m}^{(2)} s_j^{(2)}(t - (m-1)T_2 - \tau_j^{(2)}) \\
&+ \cdots + \sum_{k=1}^{K_G} \sqrt{w_k^{(G)}} b_k^{(G)} s_k^{(G)}(t - \tau_k^{(G)}) + n(t),
\end{aligned}
\tag{6.1}
$$

where

$b_k^{(G)}$ is the transmitted information bit of kth user of group $G, k = 1, \cdots, K_G$

$b_{k,m}^{(g)}$ is the transmitted information bit of kth group g user's mth subinterval, $g = 1, \cdots, G-1; m = 1, \cdots m_g; k = 1, \cdots, K_g$

$\sqrt{w_k^{(g)}}$ is the received information amplitude of kth user of group g, $g = 1, \cdots, G; k = 1, \cdots, K_g$

$s_k^{(g)}$ is the signature waveform of kth user of group g, $g = 1, \cdots, G; k = 1, \cdots, K_g$

$n(t)$ is white Gaussian noise with power spectral density σ^2.

(all the superscript $\cdot^{(x)}$ represents the group number x)

6.3   Decision Feedback Interference Cancellation for Multi-rate CDMA Signals Based on MMSE Criterion

Equation (6.1) is a representation of the base band received signal for the multi-rate CDMA systems. This system representation is equivalent to a single rate system with K_v virtual users each transmitting only one bit during the time interval $[0, T_G)$, where $K_v = m_1 K_1 + m_2 K_2 + \cdots + K_G$. The chip-rate sampled version of the received information $r(t)$ is a $(L_G \times 1)$ vector and is given by

$$
\mathbf{r} = \mathbf{SWb} + \mathbf{n} = \mathbf{Pb} + \mathbf{n}, \tag{6.2}
$$

where $\mathbf{P} = \mathbf{SW}, \mathbf{n} \sim \mathcal{N}(0, \sigma^2 \mathbf{I}_{L_G})$ and $\mathbf{I}_{L_G}$ is the identity matrix of size L_G. The $L_G \times K_v$ matrix $\mathbf{S}$ contains the spreading sequences and can be written as

$$
\mathbf{S} = [\mathbf{s}_{1,1}^{(1)} \cdots \mathbf{s}_{K_1,1}^{(1)} \cdots \mathbf{s}_{1,m_1}^{(1)} \cdots \mathbf{s}_{K_1,m_1}^{(1)} \cdots \mathbf{s}_1^{(G)} \cdots \mathbf{s}_{K_G}^{(G)}]. \tag{6.3}
$$

For VSL access method, when $g < G$, $\mathbf{s}_{k,m_g}^{(g)}$ are simply the time shifted version of each other, which means $s_{k,m_g}^{(g)}(t) = s_k^{(g)}(t - (m_g - 1)T_g)$ for $t \in [(m-1)T_g, mT_g)$. $\mathbf{b}$ is
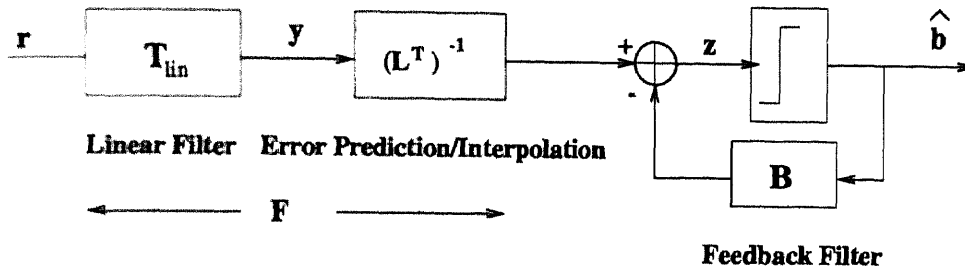
Figure 6.1 System construction of a typical decision feedback interference canceler

the column vector which contains all the K_v transmitted bits in the interval $[0, T_G)$ and can be written as

$$\mathbf{b} = [b_{1,1}^{(1)} \cdots b_{K_1,1}^{(1)} \cdots b_{1,m_1}^{(1)} \cdots b_{K_1,m_1} \cdots b_1^{(G)} \cdots b_{K_G}^{(G)}]^T. \qquad (6.4)$$

$\mathbf{W}$ is the $K_v \times K_v$ diagonal matrix with the diagonal elements corresponding to the received amplitude of the specific bit. The block diagram of a typical decision feedback interference canceler is shown in Figure 6.1, where the filter $\mathbf{F}$ is the feed forward filter, and filter $\mathbf{B}$ is a feedback filter which is lower diagonal with zero along the diagonal. The output z can be written as:

$$\mathbf{z} = \mathbf{Fr} - \mathbf{B}\hat{\mathbf{b}}, \qquad (6.5)$$

where $\hat{\mathbf{b}}$ is the output of the decision device. The associated estimation error can be written as:

$$\mathbf{e} = \mathbf{b} - \mathbf{z}. \qquad (6.6)$$

The Mean Square Error between the output data z and the transmitted information bits $\mathbf{b}$ is going to be minimized. The forward filter $\mathbf{F}$ and the backward filter $\mathbf{B}$ can be jointly determined through the MSE optimization approach as follow:

$$\min_{\mathbf{F},\mathbf{B}} MSE = \min_{\mathbf{F},\mathbf{B}}(E[\|\mathbf{b} - \mathbf{z}\|^2]). \qquad (6.7)$$

In order to get the solution of the optimum feed forward and feedback filter, we suppose that there is no error propagation, $\hat{\mathbf{b}} = \mathbf{b}$, then MSE can be written as:

$$MSE = E(\|\mathbf{b} - \mathbf{Fr} + \mathbf{Bb}\|^2). \qquad (6.8)$$

The covariance matrix of the received signal $\mathbf{r}$ is

$$\mathbf{E}[\mathbf{rr}^T] = \Sigma_{\mathbf{rr}} = \mathbf{PP}^T + \sigma^2 \mathbf{I}_{L_G}. \tag{6.9}$$

The correlation between the received signal $\mathbf{r}$ and the transmitted information $\mathbf{b}$ is

$$E[\mathbf{rb}^T] = E[(\mathbf{Pb} + \mathbf{n})\mathbf{b}^T] = \mathbf{P}. \tag{6.10}$$

Through the joint optimization based on MMSE criterion, the forward and feedback filters, $\mathbf{F}$ and $\mathbf{B}$ can be found to be:

$$\mathbf{F} = (\mathbf{I}_{K_v} + \mathbf{B})\mathbf{P}^T \Sigma_{\mathbf{rr}}^{-1}, \tag{6.11}$$

$$\mathbf{B} = (\mathbf{L}^T)^{-1} - \mathbf{I}_{K_v}. \tag{6.12}$$

$\mathbf{F}$ is essentially a concatenation of the linear MMSE filter $\mathbf{T}_{lin}$ and a error whitening filter $(\mathbf{L}^T)^{-1}$, where

$$\mathbf{T}_{lin}^T = \mathbf{P}^T \Sigma_{\mathbf{rr}}^{-1}. \tag{6.13}$$

The associated linear MMSE filter output error vector is

$$\mathbf{e}_{lin} = \mathbf{b} - \mathbf{T}_{lin}{}^T \mathbf{r}. \tag{6.14}$$

The covariance matrix of $\mathbf{e}_{lin}$ is

$$\mathbf{Q} = cov[\mathbf{e}_{lin}] = E[\mathbf{e}_{lin}\mathbf{e}_{lin}^T] = \mathbf{I}_{K_v} - \mathbf{P}^T \Sigma_{\mathbf{rr}}^{-1} \mathbf{P}. \tag{6.15}$$

The matrix $\mathbf{L}$ is defined by the Cholesky factorization of the error covariance matrix associated with the linear MMSE filter as follow:

$$\mathbf{L}^T \mathbf{DL} = \mathbf{Q}. \tag{6.16}$$

The users are demodulated successively as

$$\hat{b}_{1,1}^{(1)} = \text{sgn}[z_1]$$
$$\hat{b}_{2,1}^{(1)} = \text{sgn}[z_2 - \mathbf{B}_{2,1}\hat{b}_{1,1}^{(1)}]$$
$$\vdots$$
$$\hat{b}_{K_G}^{(G)} = \text{sgn}[z_{K_v} - \sum_{i=1}^{K_v-1} \mathbf{B}_{K_v,K_v-i}\hat{b}_i] \tag{6.17}$$

For a single rate CDMA system, it has been shown in [66] [67][68] that this structure performs ideal successive interference cancellation in the absence of error propagation. That is when detecting user i, the effect of all previously detected users 1 through $i-1$ is completely removed. This is also true for the system model of equation (6.1), since this system is essentially a virtual low-rate system with K_v virtual users when the detection is working over the received information over the period $[0, T_G)$.

6.4 Groupwise Serial Interference Cancellation for Multi-rate CDMA Signals Based on MMSE Criterion

The objective of this section is to develop the structure of the group-wise serial interference canceler for multi-rate CDMA signals. Users are grouped according to their processing gain. The receiver construction is shown in Figure 6.2. The working principle of this proposed receiver is as follow:

(1) Form the K users into G groups according to their processing gain L_g. Total $K = K_1 + K_2 + \cdots + K_G$ matched filters are sampled at different symbol interval T_g, $g = 1, 2, \cdots, G$.

(2) For $g = 1$, form the forward filter by concatenating the linear MMSE filter $\mathbf{T}_{lin}$ and error whitening filter $(\mathbf{L}^T)^{-1}$, using the feedback filter $\mathbf{B}$ to subtract the interference introduced by the user with high power in this group.

(3) For $1 < g \leq G$, form the forward and backward filters using the received information of group 1 to $g - 1$, the interference introduced by the high rate groups and the user with high power in group g can be cancelled through backward filter B.

Recall the equation (6.1), $r(t)$ contains all the information over the time interval $[0, T_G)$. Based on this equation, the base band signal which is received during the interval $[0, T_1)$ can be written as

$$r^{(1)}(t) = \sum_{i=1}^{K_1} \sqrt{w_i^{(1)}} b_i^{(1)} s_i^{(1)}(t) + partial\ information\ of\ the\ other\ groups$$
$$+ n^{(1)}(t), \tag{6.18}$$

where $n^{(1)}(t)$ is the white Gaussian noise at the receiver end. The group 1 users which have the highest bit rate are detected first neglecting the presence of the other users in the system. The MSE which is going to be minimized is:

$$MSE^{(1)} = E[\|\mathbf{b}_1 - \mathbf{z}^{(1)}\|^2], \tag{6.19}$$

where

$$\mathbf{z}^{(1)} = \mathbf{F}^{(1)}\mathbf{r}^{(1)} - \mathbf{B}^{(1)}\hat{\mathbf{b}}_1. \tag{6.20}$$

Through the joint optimization, the forward filter can be found to be:

$$\mathbf{F}^{(1)} = (\mathbf{L}^{(1)^T})^{-1}\mathbf{T}_{lin}.$$

$\mathbf{b}_1$ is the transmitted information bit vector of group 1 users during the time interval $[0, T_1)$. $\mathbf{r}^{(1)}$ is chip-rate sampled version of $r^{(1)}(t)$. The forward filter $\mathbf{F}^{(1)}$ is a concatination of linear MMSE filter and error whitening filter, where $\mathbf{T}_{lin}^{(1)^T} = \mathbf{P}^{(1)^T}\Sigma_{\mathbf{r}^{(1)}}^{-1}$, $\mathbf{P}^{(1)}$ is a $L_1 \times K_1$ matrix with ith column being the chip rate sampled version of $p_i^{(1)}(t)$,

$$p_i^{(1)}(t) = \sqrt{w_i^{(1)}}s_i^{(1)}(t), 0 \le t < T_1. \tag{6.21}$$

$\Sigma_{\mathbf{r}^{(1)}}$ is the covariance matrix of the received information vector $\mathbf{r}^{(1)}$, $\Sigma_{\mathbf{r}^{(1)}} = E[\mathbf{r}^{(1)}\mathbf{r}^{(1)^T}]$. The associated output error vector after the linear filter is:

$$\mathbf{e}_{lin}^{(1)} = \mathbf{b}_1 - \mathbf{T}_{lin}^{(1)^T}\mathbf{r}^{(1)}. \tag{6.22}$$

The covariance matrix of the above error vector is

$$\mathbf{Q}^{(1)} = E[\mathbf{e}_{lin}^{(1)}\mathbf{e}_{lin}^{(1)^T}] = \mathbf{I}_{K_1} - \mathbf{P}^{(1)^T}\Sigma_{\mathbf{r}^{(1)}}^{-1}\mathbf{P}^{(1)}, \tag{6.23}$$

All K_1 group 1 users are going to be detected successively using a forward error whitening filter $(\mathbf{L}^{(1)^T})^{-1}$ and a feedback filter $\mathbf{B}^{(1)}$ after the linear MMSE filter, where

$$\mathbf{Q}^{(1)} = (\mathbf{L}^{(1)})^T\mathbf{D}^{(1)}\mathbf{L}^{(1)} \tag{6.24}$$

$$\mathbf{B}^{(1)} = (\mathbf{L}^{(1)^T})^{-1} - \mathbf{I}_K. \tag{6.25}$$

$\mathbf{L}^{(1)}$ is defined by Cholesky factorization of the error covariance $\mathbf{Q}^{(1)}$. $\mathbf{D}^{(1)}$ is the diagonal matrix.

For the detection of the group 2 users, the received signal during the time interval $[0, 2T_1)$ can be written as:

$$r^{(2)}(t) = \sum_{i=1}^{K_1}\sum_{m=1}^{2} \sqrt{w_i^{(1)}} b_{i,m}^{(1)} s_i^{(1)}(t - (m-1)T_1) + \sum_{j=1}^{K_2} \sqrt{w_j^{(2)}} b_j^{(2)} s_j^{(2)}(t)$$
$$+ partial\ information\ of\ the\ other\ groups + n^{(2)}(t). \qquad (6.26)$$

It is noted that the group 2 users are lower rate users comparing to the group 1 users. If the energy per symbol is same for all users, the group 1 users will have higher power in their signals. Thus, the group 1 users are expected to cause more interference to the other group users which is not negligible. The MSE which is going to be minimized for the detection of the group 2 users is changed to

$$MSE^{(2)} = E\{\|\mathbf{b}_2 - \mathbf{z}^{(2)}\|^2\}, \qquad (6.27)$$

where

$$\mathbf{z}^{(2)} = \mathbf{F}^{(2)}\mathbf{r}^{(2)} - \mathbf{B}^{(2)}\hat{\mathbf{b}}_2$$
$$\mathbf{F}^{(2)} = (\mathbf{L}^{(2)^T})^{-1}\mathbf{T}_{lin}^{(2)}$$

In the above equation, $\mathbf{b}_2 = [\mathbf{b}_1^{(1)}\mathbf{b}_2^{(1)}\mathbf{b}_1^{(2)}]^T$. $\mathbf{b}_2$ include the information bits of group 1 users during the time interval $[0, T_1)$ which is $\mathbf{b}_1^{(1)}$, the information bits of group 1 users during the time interval $[T_1, 2T_1)$ which is $\mathbf{b}_2^{(1)}$, and the information bits of group 2 users during the time interval $[0, 2T_1)$ which is $\mathbf{b}_1^{(2)}$. $\mathbf{r}^{(2)}$ is chip-rate sampled version of $r^{(2)}(t)$. The linear MMSE filter part can be written as:

$$\mathbf{T}_{lin}^{(2)^T} = \mathbf{P}^{(2)^T}\Sigma_{\mathbf{r}^{(2)}}^{-1}, \qquad (6.28)$$

where $\mathbf{P}^{(2)}$ is a $L_2 \times (2K_1 + K_2)$ matrix with jth column being the chip-rate sampled version of $p_j^{(2)}(t)$ for $0 \le t < 2T_1$,

$$p_j^{(2)}(t) = \begin{cases} \sqrt{w_j^{(1)}} s_j^{(1)}(t - (m-1)T_1) & (m-1)K_1 + 1 \le j \le mK_1, m = 1, 2 \\ \sqrt{w_j^{(2)}} s_j^{(2)}(t) & 2K_1 + 1 \le j \le 2K_1 + K_2 \end{cases} \qquad (6.29)$$

$\Sigma_{\mathbf{r}^{(2)}}$ is the covariance matrix of the received signal vector $\mathbf{r}^{(2)}$ and $\Sigma_{\mathbf{r}^{(2)}} = E[\mathbf{r}^{(2)}\mathbf{r}^{(2)^T}]$. The associated error vector of the linear MMSE filter is going to be

$$\mathbf{e}_{lin}^{(2)} = \mathbf{b}_2 - \mathbf{T}_{lin}^{(2)^T}\mathbf{r}^{(2)} \qquad (6.30)$$

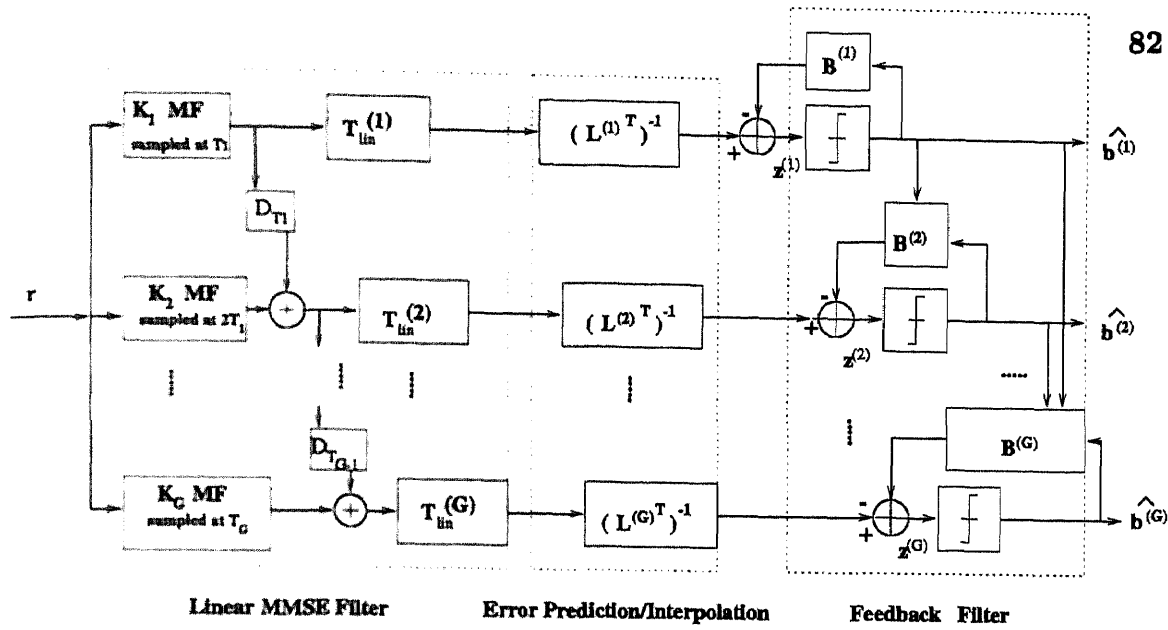Linear MMSE Filter   Error Prediction/Interpolation   Feedback Filter

Figure 6.2 System construction of groupwise successive MMSE interference canceler

which has covariance $\mathbf{Q}^{(2)} = E[\mathbf{e}_{lin}^{(2)}\mathbf{e}_{lin}^{(2)T}]$. In the same way, the error whitening filter can be found to be $(\mathbf{L}^{(2)T})^{-1}$ and the feed back filter can be found to be $\mathbf{B}^{(2)}$, where $\mathbf{Q}^{(2)} = (\mathbf{L}^{(2)})^T\mathbf{D}^{(2)}\mathbf{L}^{(2)}, \mathbf{B}^{(2)} = (\mathbf{L}^{(2)T})^{-1} - \mathbf{I}$. $\mathbf{D}^{(2)}$ is a diagonal matrix. Noted that at the end of the time interval $[0, 2T_1)$, the group 1 user's bits of the subinterval $[T_1, 2T_2), \mathbf{b}_2^{(1)}$, will be jointly decoded with the group 2 bits of the time interval $[0, 2T_1)$, $\mathbf{b}_1^{(2)}$. At the end of time interval $[0, 4T_1)$, the group 1 bits of the subinterval $[3T_1, 4T_1), \mathbf{b}_4^{(1)}$, the group 2 bits of the subinterval $[2T_1, 4T_1), \mathbf{b}_2^{(2)}$, will be jointly decoded with the group 3 bits $\mathbf{b}_1^{(3)}$. At last, and at the end of time interval $[0, T_G)$, the group 1 bits of the subinterval $[(m_1 - 1)T_1, T_G), \mathbf{b}_{m_1}^{(1)}$; the group 2 bits of subinterval $[(m_2 - 1)T_2, m_2T_2), \mathbf{b}_{m_2}^{(2)}$; the group $(G - 1)$ bits of subinterval $[\frac{1}{2}T_G, T_G), \mathbf{b}_2^{(G-1)}$; will be jointly decoded with the group G bits $\mathbf{b}^{(G)}$. Figure 6.2 shows the system construction of this proposed groupwise successive MMSE interference canceler. For the lowest rate group G users which have been interfered most, the structure of this groupwise interference canceler will perform ideal successive interference cancellation when there is no error propagation.

6.5   Simulation Results

Simulation results of the proposed groupwise MMSE interference canceler for multi-rate CDMA system are presented in this part. Direct-sequence spread-spectrum

signature waveforms and BPSK data and spreading modulation with coherent detection were used to simulate a synchronous CDMA system with multiple data rates. Length 7 Gold sequence is chosen to form the signature sequence for different users. The repetition coding scheme is used to generate lower rate users' signature sequences. This means the group g (where $g > 1$) users' signature sequences are generated by repeating length 7 signature sequences m_g times, and the processing gain of the group g users is going to be $L_g = 7 * m_g$, where $m_g = 2^{g-1}$. Table 1. shows examples of 4-user system with different processing gains. In example 1, all the users have the same processing gain $L_1 = 7$, and they all have the highest bit rate $\frac{1}{T_1}$. In example 2, there are two users with processing gain 7 and the bit rate $\frac{1}{T_1}$. Another two users have the processing gain $L_2 = 2L_1 = 14$ and bit rate $\frac{1}{2T_1}$. There are three spreading factors in example 3. First user, which has the bit rate $\frac{1}{T_1}$, has spreading factor L_1. Second user, which has bit rate $\frac{1}{2T_1}$, has spreading factor $2L_1$. The last two users, which have bit rate $\frac{1}{4T_1}$, have spreading factor $4L_1$.

The simulation results of Fig. 6.3 is performed for the system with one processing gain which equals to 31 and the received bit energy for all the active users are equal to $8dB$. The performance show that when the system is highly loaded, the successive MMSE interference cancellation scheme is performed especially well than the linear MMSE detector and the conventional detector. The bit error rates for several detection schemes, the conventional matched filter, the decorrelator, the linear MMSE detector and the proposed groupwise MMSE interference canceler, are compared in Fig. 6.4. The results are simulated for example 2, and all the users have the equal received energies per bit. It can be seen from this figure that the groupwise successive MMSE DFD outer performs the linear MMSE detection and decorrelator. Fig. 6.5 is the near-far curve of the performance of user #4, which is low-rate user 2. It can be seen from this figure that without error propagation, the performance of the low-rate user 2 can attain single user bound even when the interference level is very high. Fig. 6.6is the near-far curve of the user #2, which is high-rate user 2. It can be seen from the results that for the higher-rate users, the performance of the groupwise successive MMSE interference canceler is still better than the linear MMSE detector and decorrelator.

From the above simulation results, we can conclude that the proposed groupwise successive MMSE interference canceler outer perform both multi-rate LMMSE detector and multi-rate decorrelator. It eliminates the bit processing delay for the low-rate mode linear detectors. It is also observed that when the interfering user become stronger, the bit error rate of the lowest rate user approaches the single user bound. In addition , this work has shown that given sufficient SNR, the spectral efficiency of proposed groupwise successive MMSE DFD is significantly higher than multi-rate LMMSE detector.

Table 1: Example specifications

example	PG	rate	No. of users
1	7	$1/T_1$	4
2	7	$1/T_1$	2
	14	$1/2T_1$	2
3	7	$1/T_1$	1
	14	$1/2T_1$	1
	28	$1/4T_1$	2
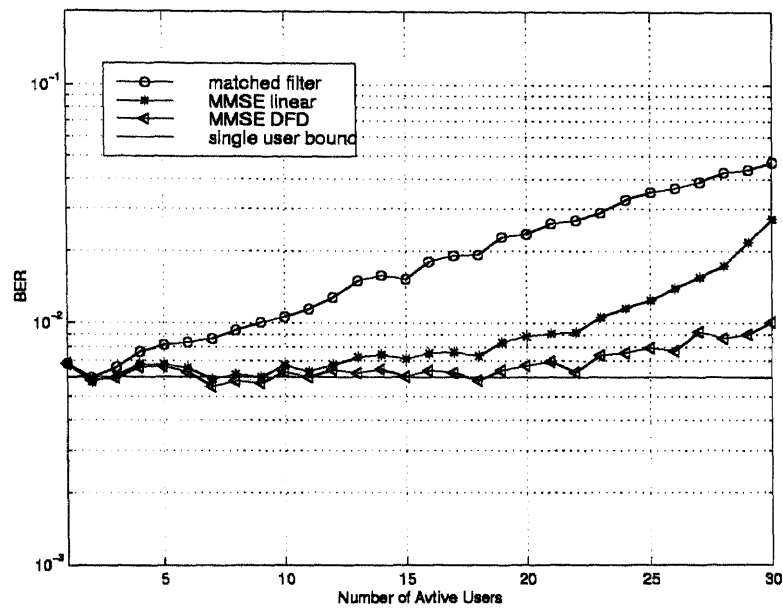
Figure 6.3 Bit error rate of multiuser detectors for a desired user in one-processing gain system, SNR = 8dB and it is same for all users
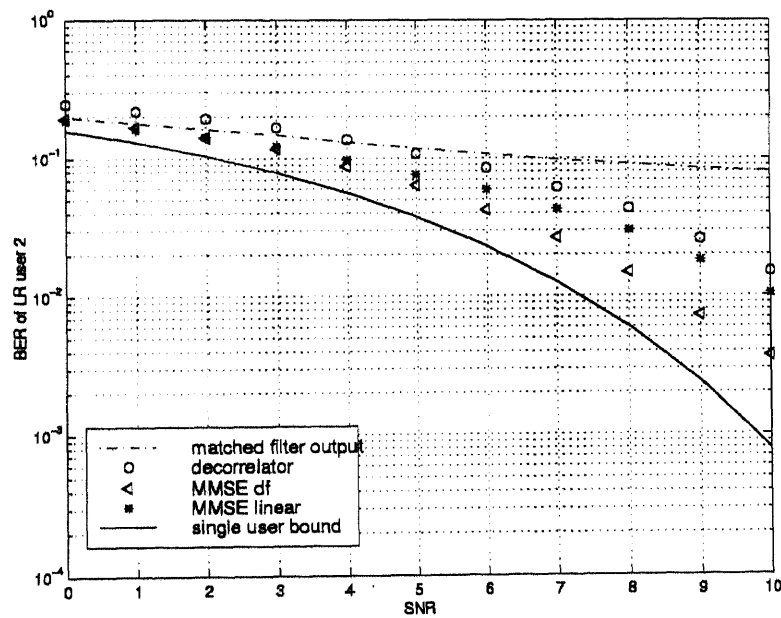
Figure 6.4 Bit error rate of multiuser detectors in a two-processing gain system with equal received energies
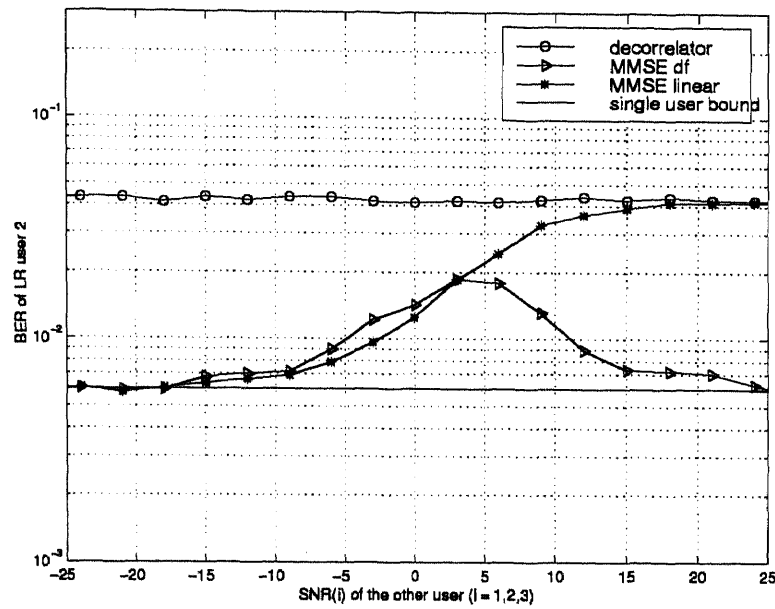
Figure 6.5 Bit error rate of multiuser detectors for desired user # 4, which is the 2nd low-rate user, in a two-processing gain system, SNR(4) = 8dB
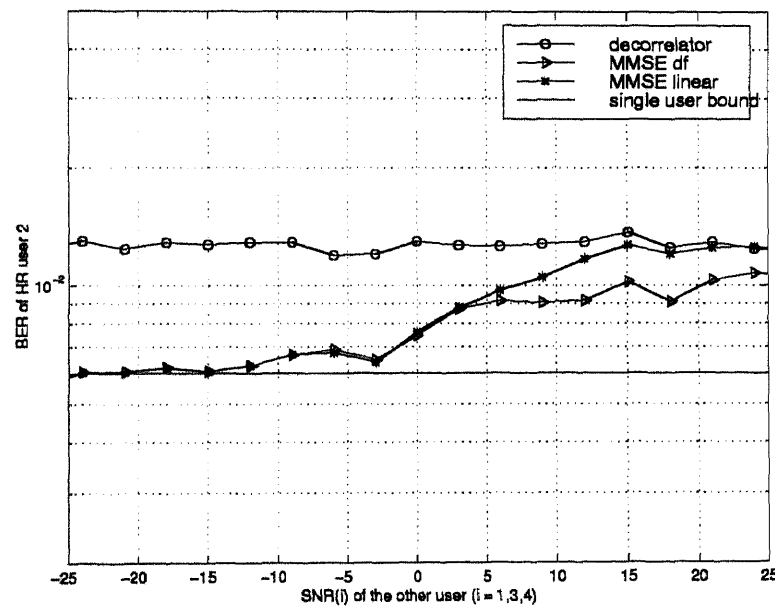
Figure 6.6 Bit error rate of multiuser detectors for desired user # 2, which is the 2nd high-rate user, in a two-processing gain system, SNR(2) = 8dB

CHAPTER 7

CONCLUSIONS

7.1 Summary

In this thesis, we explore the application of different interference cancellation techniques to multi-rate CDMA systems that serve users with different QoS.

We propose the linear MMSE receivers for synchronous downlink dual-rate CDMA system. Comparing it with dual-rate decorrelating detectors, the proposed LMMSE multi-rate receiver can offer better performance than the multi-rate decorrelator due to the fact that LMMSE detector offers best trade-off between the MAI cancellation and noise variance enhancement. In addition, dual-rate LMMSE receiver can be adaptively implemented only with the knowledge of the desired user while dual-rate decorrelating detector requires the knowledge of all the active users.

For asynchronous uplink dual-rate CDMA system, we propose the dual-rate multi-shot decorrelating detectors and dual-rate multi-shot MMSE detectors. The performance of multi-shot detectors can be improved monotonically with increasing the number of stacked bits, but a great computational complexity is going to be introduced in order to get better performance. Debiasing method is introduced to dual-rate multi-shot linear detectors. Debiasing method optimizes dual-rate detectors based on the dual-rate multi-shot model. Debiasing multi-shot MMSE detector for dual-rate signals can offer better performance than the corresponding debiasing multi-shot decorrelating detector.

Based on the above investigation for the receiver design of multi-rate systems, we extend our work to frequency selective multi-path fading channels and propose low-rate mode LMMSE space-time detection and two-stage LMMSE space-time detection. Both detection is based on minimum mean-squared error criteria. Two-stage LMMSE space-time receiver eliminate the processing delay for high-rate users and reduce the computational complexity. Two kinds of approaches are introduced to both detections. The first approach need the explicit knowledge of the array and channel information in the code array/channel matched filter part. This need to be estimated in advance using training or pilot sequences. The type 2 approach cascade the array/channel weighting, combining and multiuser detection together.

No explicit knowledge of the array/channel information is needed. The channel estimation can be incorporated in to certain adaptive implementation scheme, and only the knowledge of the desired user's timing and signature waveform is needed. We also perform a comparative study on the dual-rate receiver which use multi-path (temporal) processing, array (spatial) processing and the one which use both array and multi-path processing. The space-time receiver for the dual-rate CDMA signals give us the potential of improving the capacity of multi-rate systems. The space-time processing combined with multi-user detection have the advantages of combating multipath fading through temporal processing, reducing MAI through MMSE method and provide antenna or diversity gain through spatial processing to increase the capacity of the multi-rate CDMA systems.

The nature of the unequal received amplitudes for multi-rate CDMA signals motivates us to propose the group-wise interference cancellation methods for multi-rate CDMA signals. The non-linear decision feedback detection (DFD) schemes are used in the proposed receivers. Users with same data rate are grouped together. Users with the highest data-rate are detected first. Interference between the groups is cancelled in a successive order. The results show that the group-wise MMSE DFD yields better performance than multi-rate linear MMSE detector and multi-rate decorrelating detector, especially for highly loaded CDMA systems.

7.2 Future Work

- Based on the previous work, we found out that the choice of the access method and the set of signature sequences will influence the receiver performance, and the simulation results show the performance is very sensitive to the choice of spreading codes. In order to remove the influence of the signature sequences, further work need to be done for the random signature sequence analysis for different access methods for our proposed receivers.

- Most current CDMA systems (including IS-95 as well as military DS communication systems) are based on long spreading waveforms. For such systems, the structure of the multiple-access interference changes from symbol to symbol. This is a disadvantage for implementing multiuser detection, since the detector

must be time-varying, and explicit knowledge of interference parameters is required. However, even short spreading waveform is adopted in future, a set of relative delays and spreading waveforms that leads to poor performance could persist for a long period of time for short spreading waveforms. The research results in adaptive interference suppression can reach the point that a complete system design based on it is within reach (assuming application of sufficient engineering effort and ingenuity), and that such system would provide large performance gains even over severely time varying channels compared to existing DS/CDMA systems based on long spreading waveforms and conventional reception.

- For both MC and VSL access method, the number of effective users in a virtual low-rate system is higher than the real number of users. The multiple modulation access scheme will generate the same number of effective users as the real number of users. The multiple modulation access scheme may introduce severe power imbalance problem, but the linear and non-linear MMSE estimation based detection can utilize the power imbalance problem in the multi-rate communication system. The multiuser detection and interference cancellation schemes for multi-rate CDMA systems using multiple modulation access method need to be studied.

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