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Chair: Dr. James Caffery, Jr.

Dr. Howard Fan

Dr. Qing-An Zeng

SEMIBLIND MULTIUSER DETECTION FOR MC-CDMA

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by

Hanhong Shen

B.S., Shanghai Jiao Tong University, Shanghai, China

Thesis Advisor and Committee Chair: Dr. James Caffery, Jr.

Abstract

The code-division multiple-access (CDMA) technique has emerged in recent years as the preferred multiple access technique for providing voice and multimedia services in modern mobile communications. However, the traditional use of the DS-CDMA technology does not appear realistic for very high data-rate multimedia services due to the severe multipath-induced intersymbol interference as well as multiple access interference. To mitigate the interchip and intersymbol interferences, multicarrier CDMA (MC-CDMA) communication systems, which integrate the advantages of multicarrier transmission systems with those of CDMA, has attracted much attention over the last several years.

Traditional multiuser detection techniques for MC-CDMA require either prior knowledge, such as spreading sequences, amplitudes and/or timing, regarding all active users or some explicit estimation procedures for this information. It makes such technology difficult for practical implementation for high data rate multiuser receivers. In this thesis, a different approach for MC-CDMA multiuser receiver design is proposed. Instead of the traditional multiuser vector channel model, a new semiblind multiuser vector channel model and a new semiblind multiuser detection framework are proposed. With this semiblind multiuser detection framework, several group-based semiblind multiuser detectors, which can detect the information bits of more than one spreading channel or user, are proposed for MC-CDMA wireless communication systems. These multiuser detectors are based on a semiblind signature matrix and least squares criterion. The proposed algorithms are simple and direct. Differing from other blind or semiblind multiuser detection algorithms, no search or pairing procedure is required but only the amplitude of the desired user. Besides the derived relationship between the proposed algorithms and the conventional decorrelating detector, a geometric interpretation and the criteria of choosing the size of the semiblind spreading matrix are discussed. Computer simulations are finally provided to demonstrate the performance of the algorithms.

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GLOSSARY

3G	Third Generation
4G	Fourth Generation
AWGN	Additive White Gaussian Noise
ACF	Auto-Correlation Function
ADSL	Asymmetric Digital Subscribe Line
BS	Base Station
BPSK	Binary Phase-Shift Keying
CCF	Cross-Correlation Function
CDMA	Code-Division Multiple Access
CE	Controlled Equalization
CMF	Chip Matched Filter
CP	Cyclic Prefix
DD	Decorrelating Detector/Detection
DFT	Discrete Fourier Transform
DS-CDMA	Direct Spreading Code-Division Multiple Access
DVB	Digital Video Broadcast
EGC	Equal Gain Combining
FDMA	Frequency-Division Multiple Access

FH-CDMA	Frequency-Hopping Code-Division Multiple Access
FFT	Fast Fourier Transform
IC	Interference Cancellation
ISI	Inter-Symbol Interference
ICI	Inter-Carrier Interference
IFFT	Inverted Fast Fourier Transform
FEC	Forward Error Correction
LS	Least Squares
MAI	Multiple Access Interference
MC-CDMA	Multi-Carrier Code-Division Multiple Access
MC-DS-CDMA	Multi-Carrier Direct Spreading Code-Division Multiple Access
ML	Maximum Likelihood
MLS	Mix Least Squares and Total Least Squares
MMSE	Minimum Mean Squares Error
MMSEC	Minimum Mean-Square Error Combining
MRC	Maximum-Ratio Combining
MS	Mobile Station
MT-CDMA	Multi-tone Code-Division Multiple Access
MUD	Multi-user Detection
OFDM	Orthogonal Frequency Division Multiplexing

ORC	Orthogonality Restoring Combining
PAPR	Peak to Average Power Ratio
PDF	Probability Density Function
QoS	Quality of Services
SMF	Symbol Matched Filter
SVD	Singular Value Decomposition
TDMA	Time-Division Multiple Access
TH-CDMA	Time-Hopping Code-Division Multiple Access
TLS	Total Least Squares

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Chapter 1

Introduction

Over the past few years, with the increasing demand of bandwidth for wireless multimedia applications, several wireless technologies have been intensely studied in order to provide faster and more reliable data over wireless channels, such as CDMA, orthogonal frequency-division multiplexing (OFDM), and their combination: MC-CDMA. The advantages of multicarrier modulation, on one hand, and the flexibility offered by the spreading technique, on the other hand, have motivated many researchers to investigate the combination of both techniques, known as Multi-Carrier Spreading Spectrum. It allows one to benefit from several advantages of both multicarrier modulation and spreading spectrums by offering high flexibility, high spectral efficiency, simple and robust detection techniques and narrow band interference rejection capability [1, 2, 3].

1.1. Code Division Multiple Access (CDMA)

For the past few years, considerable research and development has been done in the field of CDMA. CDMA is a multiple access technique which uses spreading sequences to modulate signals and separate different users, thereby allowing multiple users to access shared radio channel resource simultaneously and even asynchronously [1]. In CDMA, each user is assigned a unique spreading sequence, which has a sufficiently low crosscorrelation with all the other users, to encode its information-bearing signals. Since the chip rate of the spreading sequences is much higher than that of the original information-bearing signals, the encoded signals have much wider bandwidth and different users share the same frequency spectrum to transmit signals [3]. At the receiver, the information bearing signals of the desired user are recovered by despreading the received signal with known spreading sequence. Due to the special property of spreading sequences, theoretically the information bearing signals of the desired user can be fully recovered.

CDMA has several unique advantages as a multiple access technique, compared with other techniques such as frequency–division multiple access (FDMA) and time–division multiple access (TDMA) techniques [3, 6, 7, 8]:

1. **Universal frequency spectrum reuse**. Different users are able to share the same frequency spectrum because of the use of different spreading sequences. As a result, frequency-dependent transmission impairment has limited effect on signals.
2. **Security**. Since the spread sequences are noise-like, the encoded signals of each user are therefore also noise-like to all the other uses sharing the same frequency spectrum. As a result, it is difficult to detect or jam the spread signals.
3. **Flexible channel traffic pattern**. It is easy to change the spreading sequences.
4. **No absolute limit on the number of users**. There is a theoretical limit on the number of orthogonal spreading sequences that a CDMA system can offer, which means CDMA system has an upper limit on the number of users when using orthogonal spreading sequences. However, using proper multiuser detection techniques, it is feasible to use non-orthogonal spreading sequences with low crosscorrelation; therefore, more users can be supported in the same system. Theoretically, there is no limit on the number of users in non-orthogonal CDMA systems; the system performance gradually degrades when more users are added.

Because of its properties, CDMA seems to be the most plausible candidate for 3G wireless personal communication systems [3, 6]. However, CDMA also has several drawbacks that cannot be ignored.

1. **Near-far problem**. This well-known problem happens when interfering users are closer to the base station than the desired user. When transmitting the same power, the interfering signals would have higher received power than that of the desired user and would therefore

overpower the signal of the desired user. The near-far problem could impair system performance dramatically. To combat this problem, it is necessary for the CDMA system to implement strict power control, the purpose of which is to ensure that each user has the same received power levels at the base station.

2. **High complexity of receiver**. CDMA systems, especially DS-CDMA systems, are very sensitive to multipath effects [7]. To exploit all the multipath diversity, RAKE receivers and corresponding diversity combining techniques are usually applied. In addition, channel estimation is also necessary due to the time-variant property of the channel. As a result, the receiver is highly complex.
3. **Easy self-jamming**. This happens because the spreading codes used by different users are not exactly orthogonal. Contributions from other users will significantly affect the receiver's decision.

There are a number of various modulation varieties in CDMA systems, such as DS-CDMA in which spreading sequences are directly multiplied with information bearing signals, FH-CDMA in which the carrier frequency changes with the spreading sequences, TH-CDMA in which the information bearing signals are transmitted in short bursts according to spreading sequences instead of continuously, and hybrids of these techniques. Among these different techniques, DS-CDMA is the most popular since it has several attractive properties for wireless medium [6, 21, 22, 42], such as easy generation of spreading codes, easy generation of frequency synchronizer, and possibility of coherent demodulation. However, the traditional use of the DS-CDMA technology does not appear realistic for very high data-rate multimedia services at the speeds of the order of several hundred megabits per second due to the severe multipath-induced interchip and intersymbol interference (ISI) as well as multiple access interference (MAI) [42].

1.2. Orthogonal Frequency Division Multiplexing (OFDM)

The breakthrough of OFDM came in the 1990s as it was the modulation chosen for asynchronous digital subscriber lines (ADSL) in the USA and it was selected for the European digital video broadcasting (DVB-T) standard [23, 41]. OFDM is a multicarrier modulation scheme; the principle of OFDM is to convert a serial high rate data stream to multiple parallel low rate substreams. Each substream is then modulated onto a different subcarrier. The frequency spacing between subcarriers provides orthogonality, so that each individual demodulator only sees its own frequency. OFDM has an inherent property to combat multipath effects because of lower signal rate and long signal duration [8]. It has gained lots of attention and is considered to be a plausible candidate for 4G wireless communication systems.

As a multicarrier modulation scheme, the advantages of OFDM include:

1. **High spectral efficiency**. The power spectra of subcarriers overlap while orthogonal at each center frequency. The same bandwidth can therefore accommodate more users simultaneously.
2. **Insensitivity to ISI and ICI**. Because the data rate of the converted subcarrier is much lower than the original data rate, the symbol duration is much longer than the original signals and ISI due to high data rate is significantly decreased. OFDM also adds guard time to each OFDM symbol to combat ICI and ISI.
3. **Higher data rate**. Data stream is split into L parallel subcarriers, so a higher data rate with desirable performance is possible.
4. **Better performance with less complexity**. OFDM splits data into parallel independent narrowband channels, each channel encounters only flat fading. As a result, adaptive equalization is unnecessary, and it is possible to operate at high delay spread or time variant environments. In addition, by combing narrowband subcarriers with proper interleaving and error coding techniques, OFDM can effectively combat multipath effects; using guard interval offers the receiver an extremely simple way to eliminate ISI and ICI.
5. **Computational efficiency**. By using FFT, it is highly efficient to implement modulation and demodulation in OFDM systems.

The major drawbacks of OFDM are [5]:

1. **High Peak to Average Power Ratio (PAPR).** OFDM employs a large number of subcarriers; the addition of signals in each subcarrier introduces large PAPR. It is very difficult and expensive to have power amplifiers with such large linear range; the nonlinear amplifier has to be used which may cause distortion of signal waveform therefore increasing in-band and out-of-band interference.
2. **High sensitivity to carrier frequency offset**. The power spectra of subcarriers overlap. A frequency offset will result in a shift of the received signal spectrum, which causes the power spectra of OFDM subcarriers to no longer be orthogonal to each other. This leads to ICI which may potentially cause a severe performance degradation of OFDM systems.

1.3. Multicarrier Code Division Multiple Access (MC-CDMA)

For those multimedia applications which require very high data rate (up to several hundred megabits per second), DS-CDMA is not feasible because of the severe ICI and ISI at very high data rate. MC-CDMA was developed to alleviate the problem. MC-CDMA was the combination of CDMA and OFDM techniques with an aim to enhance wireless system capabilities and accommodate the benefits of both CDMA and OFDM systems [2] and may even outperform both. The principle of MC-CDMA is to spread the codes in the frequency domain: a single signal is replicated into L parallel copies; each of the copies is multiplied by a single chip of a spreading code of length L and then modulated (usually using BPSK modulation) to a subcarrier. This is shown in Fig. 1, where $\mathbf{c} = \begin{bmatrix} c_1 & c_2 & \cdots & c_L \end{bmatrix}^{\mathrm{T}}$ denotes the spreading sequence for single user. The frequency spacing between two adjacent (out of of the L) subcarriers is F/T_b, where F is an integer and T_b is the symbol duration [4].

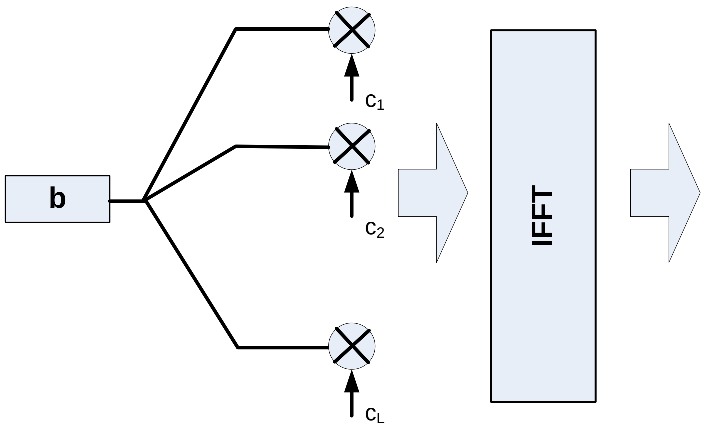

Figure 1: MC-CDMA transmitter block diagram.

Besides the advantages of CDMA and OFDM, MC-CDMA also has the following impressive advantages:

1. **Much simpler receiver design compared with DS-CDMA**. Theoretically, a DS-CDMA system with spreading sequences of length N could accommodate N users simultaneously, but it is always infeasible because of the tremendous computational load. A MC-CDMA system can handle these N users easily using a standard receiver structure. Furthermore, MC-CDMA provides guard interval for each symbol, so a RAKE receiver used to combat multipath effects is no longer necessary.
2. **Higher spectral efficiency**. OFDM often uses coding techniques to enhance the performance; as a result, the numbers of subcarriers are greater than those of the symbols transmitted simultaneously. In MC-CDMA systems, there is no coding needed at the transmitter, so the number of subcarriers is much less than that of OFDM.
3. **Inherent frequency diversity.** MC-CDMA provides inherent frequency diversity in that each individual subcarrier may encounter fading independently. Using diversity combining techniques may enhance the system performance.

However, MC-CDMA still has the following major drawbacks [5, 43]:

1. **High Peak to Average Power Ratio (PAPR).**
2. **Sensitivity to frequency offset and phase noise.** Frequency offset and phase noise will damage orthogonality between subchannels therefore will cause the system performance lose.

1.4. Multiuser Detection

Multiple access interference is the major factor that limits the capacity and performance of DS-CDMA systems. Earlier detectors did not consider MAI and performed single user detection. To mitigate MAI effects, those detectors focused on improving spreading code waveform design, power control, forward error correction (FEC) code design, sectored/adaptive antenna design, and so on [16]. This approach is simple and easy to implement. However, it ignores the crosscorrelation between spreading sequences and the (possibly) known structure of the data signals. Therefore, the performance of this scheme is poor and vulnerable to the near-far problem. To combat the near-far problem and MAI, more sophisticated detection strategies, such as multiuser detection (MUD), need to be implemented at the receiver. Multiuser detection is a strategy for mitigating MAI effects at the receiver side. For the last ten years, various multiuser detection schemes have been developed to combat the effects of MAI. Most early research work was focused on conventional multiuser detectors, which assume knowledge of all active users, including both known users and unknown interfering users, perform interference cancellation and achieve optimal or suboptimal detection results. For example, with the knowledge of all active users' spreading sequences, the classic decorrelating detector can completely eliminate the effects of MAI at the expense of enhancing background noise. This multiuser detection scheme offers superior performance compared to the traditional decorrelating detectors [15, 16, 17]; however it is also complex. The complexity increases dramatically with the number of users which makes it hard to implement in practice. Meanwhile, conventional multiuser detectors require too much information about users, such as the knowledge of all the spreading sequences, received signal amplitudes, and so on, which makes them unrealistic in practice. Compared to decorrelating detection, blind multiuser detection has been developed with only knowledge of the desired user(s) information so that such designs may be much more practical. Though blind multiuser detection has no a priori knowledge of other users' information, they normally use additional statistic signal processing procedures to estimate other users' spreading sequences or channel information [15, 16, 17]. This normally takes more time or bandwidth; otherwise, detection performance may be sacrificed.

In this thesis, we are looking for the trade-offs between the blind algorithms and the conventional detectors, and we propose several semiblind multiuser detection schemes which only require the additional amplitude information of desired users instead of all the active users.

1.5. Thesis Outline

The rest of the thesis is organized as follows:

- Chapter 2 discusses the fundamentals of CDMA and OFDM systems: signal structure, signal transmission and reception. The basic multiuser detection technique, the decorrelating detector, is discussed in detail.

- Chapter 3 presents the fundamentals of MC-CDMA systems. We discuss MC-CDMA modulation and demodulation procedures for both the reverse link and forward link of MC-CDMA systems. We represent MC-CDMA systems using a vector model and also discuss existing multiuser detection schemes for MC-CDMA.

- Chapter 4 presents the proposed group-based semiblind multiuser detection schemes, which can simultaneously detect more than one spreading channel. At first, a semiblind multiuser detection framework is presented. After that, we present least-square-based detection schemes, including least-squares detection, total least-squares detection and mixed detection.

- Chapter 5 discusses the geometric explanation and how to choose the size of the semiblind spreading matrix which is critical in the proposed group-based semiblind multiuser detection schemes.

- Chapter 6 presented the computer simulation results and final conclusion. Future work is discussed in Chapter 7.

1.6. Thesis Contribution

In this thesis, a new group-based semiblind multiuser detection framework and several detection algorithms are presented for MC-CDMA. They are synchronous multiuser detection schemes. They are simple and direct without any search or convergence procedure. The relationship with the conventional decorrelating detector is revealed. We also present the geometric explanation and give a closed-form expression for the new noise covariance matrix. These theoretical analysis results can be used to evaluate the performance of the proposed algorithms.

Chapter 2

Background

Spread spectrum systems have been utilized since the 1950s. They have advantages for anti-jamming and anti-multipath in wireless communications [3, 6, 7, 8]. Conventional DS-CDMA systems offer several advantages in cellular environments including easy frequency planning, high immunity against interference if a high processing gain is used and flexible data rate adaptation. However, it is well-known that DS-CDMA is an interference-limited system, which means it suffers from both in-cell and out-of-cell multiple access interference, even with perfect power control. Multiuser detection is one of the techniques for mitigating MAI in CDMA systems. In conventional multiuser detection schemes, it assumes that each multiuser receiver has enough priori knowledge regarding all interference-contributing users' spreading sequences, amplitudes and timing. It then utilizes this knowledge to do interference cancellation and detection. Multiuser receiver design using MUD may be able to improve the system capacity and achieve high spectral efficiency in CDMA systems. In addition, the technique of multicarrier transmission has received wide interest especially for high data-rate broadcast applications. The principle of multicarrier transmission, especially orthogonal frequency division multiplexing (OFDM), is to convert a serial high-rate data stream into multiple parallel low-rate substreams, each of which is modulated on a subcarrier. Since the symbol rate on each subcarrier is much less than the initial serial data rate, the effect of delay spread significantly decreases and also the complexity of equalization at the receiver can be reduced. OFDM is a low-complexity technique for efficiently modulating multiple subcarriers by using modern digital signal processing.

2.1. Direct-Sequence Code Division Multiple Access (DS-CDMA)

In DS-CDMA the discrete-time spreading sequences are directly multiplied with the ,information bearing signals [3].

2.1.1. DS-CDMA Signal

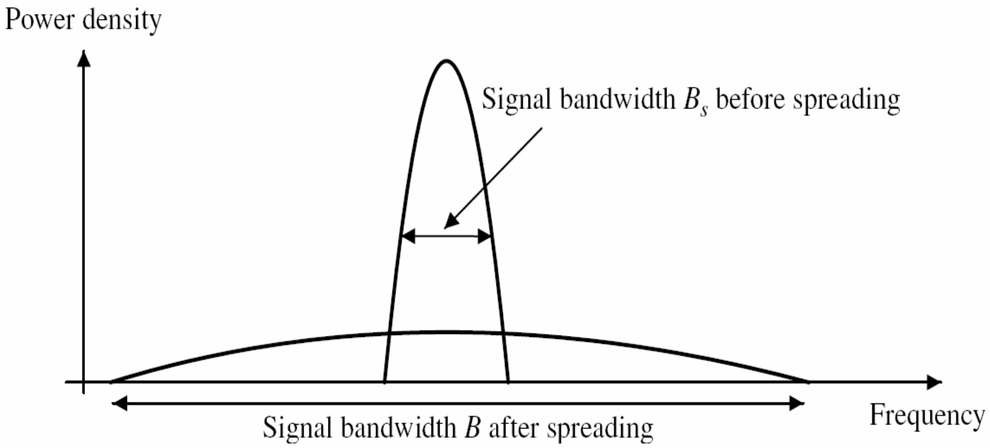

Figure 2: Power spectrum density before and after direct sequence spreading.

Literally, as shown in Fig. 2, a spread spectrum signal is a signal that is frequency-spread and transmitted over a wide frequency band, which is much wider than the minimum bandwidth required. The principle of DS-CDMA is to spread a data symbol with a spreading waveform $s_k(t)$, of length $T_s = L_c T_c$,

$$s_k(t) = \sum_{l=1}^{L_c} c_k[l] \phi(t - lT_c) \tag{2.1}$$

assigned to user k, $k = 1, 2, \ldots, K$, where K is the total number of active users. $\phi(t)$ is the pulse shaping filter. T_c is the chip duration and $c_k[l]$ is the spreading sequence. After spreading, the signal $x_k(t)$ of user k is given by

$$x_k(t) = b_k[n] \sum_{l=1}^{L_c} c_k[l] \phi\left(t - nT_s - lT_c\right) \tag{2.2}$$

for $0 \le t \le T_s$, where $b_k[n]$ is the transmitted bit of user k. The multiplication of the information $b_k[n]$ with the spreading sequence $s_k(t)$ is done bit-synchronously and the overall transmitted signal $x_k(t)$ of all K synchronous users results in

$$x(t) = \sum_{k=1}^{K} x_k(t) \quad . \tag{2.3}$$

The proper choice of spreading sequence is critical in DS-CDMA, since the MAI strongly depends on the crosscorrelation of the used spreading sequences. To minimize the MAI, the crosscorrelation function values should be as small as possible. In order to guarantee equal interference among all transmitting users, the crosscorrelation properties between different pairs of spreading sequences should be similar. Moreover, the autocorrelation function of the spreading sequences should have low out-of-phase peak magnitudes in order to achieve a reliable synchronization.

2.1.2. DS-CDMA Signal Transmission

The received signal $y(t)$ obtained at the output of a multipath channel with P different propagation paths and K users can be expressed as

$$
\begin{aligned}
y(t) &= \sum_{k=1}^{K}\sum_{p=1}^{P} y_k\left(t-\tau_p\right)+n(t) \\
&= \sum_{k=1}^{K}\sum_{p=1}^{P} \alpha_{pk} x_k\left(t-\tau_p\right)\otimes h\left(t-\tau_p\right)+n(t)
\end{aligned}
\tag{2.4}
$$

where $h(t)$ is the channel impulse response, $y_k(t)$ is the noise-free received signal of user k, $n(t)$ is additional white Gaussian noise (AWGN) with variance σ^2, α_{pk} is the p^{th} channel loss factor of user k, $\tau_1 < \tau_2 < \cdots < \tau_p$ are the delays for the P paths and $\otimes$ denotes the convolution product operation.

2.1.3. DS-CDMA Receiver

The received signal $y(t)$ is synchronized through a chip matched filter $\phi(t)$ to each path and combined by a RAKE combiner. The output of the RAKE receiver is

$$
r(t) = y_k\left(t-\tau_1\right)\otimes\phi\left(t-\tau_1\right)+m_{\mathrm{ISI}}(t)+m_{\mathrm{MAI}}(t)+n(t) \tag{2.5}
$$

where

$$
m_{\mathrm{ISI}}(t) = \sum_{p\neq q}^{p,q=P} \beta_{qk}\alpha_{pk} y_k\left(t+\tau_q-\tau_1-\tau_p\right)\otimes\phi\left(t-\tau_q\right) \tag{2.6}
$$

is the ISI to user k,

$$
m_{\mathrm{MAI}}(t) = \sum_{j\neq k}\left\{ y_j\left(t-\tau_1\right)\otimes\phi\left(t-\tau_1\right)+\sum_{p\neq q}^{p,q=P} \beta_{qk}\alpha_{pj} y_j\left(t+\tau_q-\tau_1-\tau_p\right)\otimes\phi\left(t-\tau_q\right)\right\} \tag{2.7}
$$

is the MAI to user k, β_{qk} is the weight of the q^{th} RAKE finger for user k with $\sum_{q=1}^{P} \beta_{qk}\alpha_{qk} = 1$.

User k's RAKE output can be sampled at $f_S = 1/T_S$ and expressed by

$$\begin{aligned} \mathbf{r} &= \left[r\left(nT + T_S + \tau_1\right) \quad r\left(nT + 2T_S + \tau_1\right) \quad \ldots \quad r\left(nT + LT_S + \tau_1\right) \right]^{\mathrm{T}} \\ &= \sum_{k=1}^{K} A_k b_k \mathbf{s}_k + \mathbf{n} \\ &= \mathbf{SAb} + \mathbf{n} \end{aligned} \tag{2.8}$$

where $L = T_s / T_S$ is the number of samples per symbol and should not be less than the spreading gain L_c and $\mathbf{S}$ is the unknown spreading signature matrix containing all ISI and MAI effects.

Because of $m_{\mathrm{MAI}}(t)$ in the received signal $r(t)$, the performance of conventional matched filter receiver will suffer from the so-called near-far problem where a strong or near-by user can prevent the detection of weak or far-away users. Multiuser detection is the receiver side detection technique.

2.2. Multiuser Detection

Multiuser detection is a method for minimizing MAI effects by exploiting the interference structure. It can mitigate the near-far problem in CDMA systems and therefore improve system capacity, which should be decided by the thermal noise instead of MAI. Multiuser receivers include the optimum maximum likelihood (ML) receiver, linear decorrelating receiver, linear minimum mean square error (MMSE) receiver, successive interference cancellation (SIC) receiver, decision feedback receiver, multistage interference cancellation receiver, and so on [15, 16, 17].

With the conventional vector multiuser channel model (2.8), the classic decorrelating detector $\mathbf{W}_{\mathrm{DD}}$ solves the following problem:

$$\mathbf{W}_{\mathrm{DD}} = \arg \min_{\mathbf{x} \in \mathbb{R}^{K \times L}} \left\| \mathbf{X}\mathbf{r} - \mathbf{A}\mathbf{b} \right\|_2^2 \quad . \tag{2.9}$$

Since the received spreading sequence $\mathbf{S} \in \mathbb{R}^{L \times K}$ is assumed to be a full-rank matrix with $\operatorname{rank}\{\mathbf{S}\} = K$, the least-squares solution to (2.9) is given by

$$\mathbf{b}_{\mathrm{DD}} = \operatorname{sgn}\left\{ \left(\mathbf{S}^{\mathrm{T}} \mathbf{S} \right)^{-1} \mathbf{S}^{\mathrm{T}} \mathbf{r} \right\} \quad . \tag{2.10}$$

A block diagram of the conventional decorrelating detection is shown in Fig. 3.

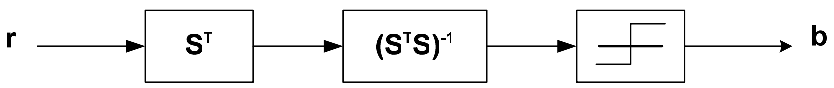

Figure 3: The block diagram of the conventional decorrelating detection.

In decorrelating detection, the desired user's detector can be described as the projection of its own spreading sequence vector onto the subspace orthogonal to the spreading vectors of the others users [1, 39]. Furthermore, in [38], it is shown that the above decorrelating detection is actually the oblique projection of the desired user's signature vector onto the space spanned by other users' signature vectors along the complement space of that space. This is illustrated in Fig. 4.

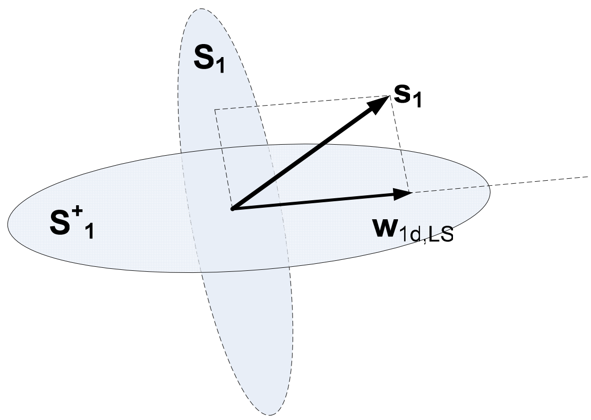

Figure 4: The geometric explanation of decorrelating detection.

The decorrelating detector is designed to completely eliminate the MAI caused by other users, at the expense of enhancing the ambient noise. Thus, when the received amplitudes are completely unknown, the decorrelating detector is a sensible choice. There are some desirable features of this multiuser detector: it does not require knowledge of the received amplitude, but it does require knowledge of all active users' spreading sequences through the matrix $\mathbf{S}^{+} = \left(\mathbf{S}^{\mathrm{T}}\mathbf{S}\right)^{-1}\mathbf{S}^{\mathrm{T}}$; it can also readily be decentralized in the sense that the demodulation of each user can be implemented completely independently.

2.3. Orthogonal Frequency Division Multiplexing (OFDM)

OFDM is a multicarrier modulation scheme. The principle of OFDM is to convert a serial high rate data stream to multiple parallel low-rate substreams.

2.3.1. OFDM Signal

Multicarrier modulation (MC) communication simultaneously transmits L complex-valued information symbols in parallel on L subcarriers. In OFDM the L substreams are modulated on subcarriers with a spacing of

$$\begin{aligned}\Delta_s &= \frac{1}{T_s} \\ &= \frac{1}{LT_d}\end{aligned} \tag{2.11}$$

in order to achieve orthogonality between the signals on the L subcarriers, where T_d is the source data symbol length. The L parallel modulated data symbols, d_n, is referred to as an OFDM symbol. The complex envelope of an OFDM symbol with pulse shaping $\phi(t)$ has the form

$$x(t) = e^{j2\pi f_0 t} \sum_{l=0}^{L-1} d_l \, \phi(t) e^{j2\pi l \Delta_s t} \tag{2.12}$$

with the L subcarrier frequencies located at

$$f_l = f_0 + l\Delta_s \quad . \tag{2.13}$$

The normalized power density spectrum of an OFDM symbol with 16 subcarriers versus the normalized frequency is depicted as a solid curve in Fig. 5. The L symbols are transmitted with equal power. The dotted curve illustrates the power density spectrum of the first modulated subcarrier and indicates the construction of the overall power density spectrum as the sum of L individual power density spectra shifted by Δ_s.

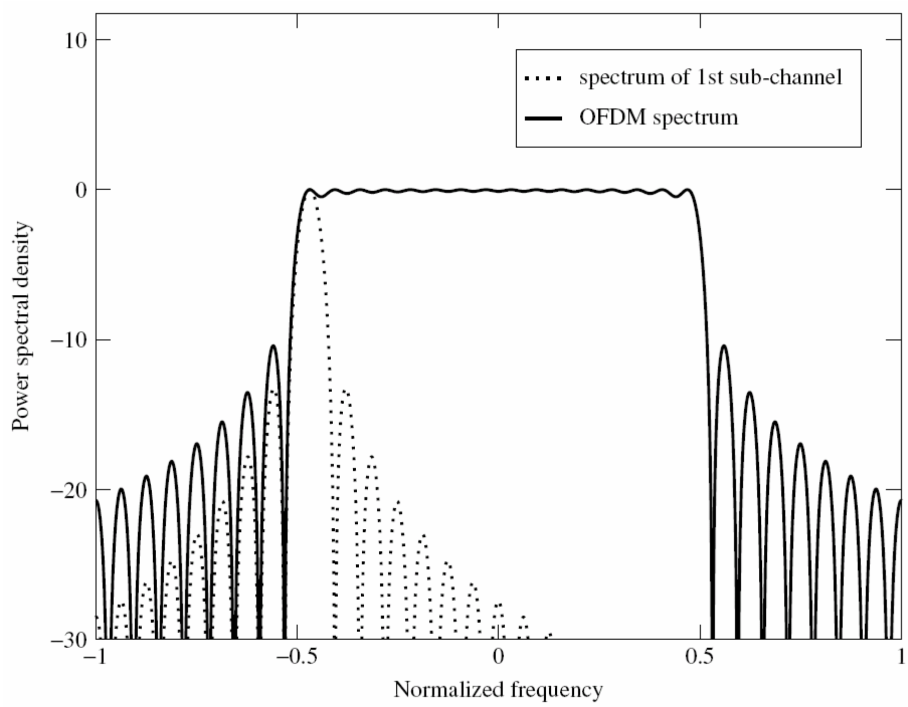

Figure 5: OFDM symbol with 16 sub-carriers.

One of the key advantages of using OFDM modulation is that multicarrier modulation can be implemented efficiently in the discrete domain by using an IDFT or IFFT. When sampling $x(t)$ with rate $1/T_d$, the samples become

$$x_m = \sum_{l=0}^{L-1} d_l \, e^{j2\pi \frac{lm}{L}} \tag{2.14}$$

which resembles the familiar IFFT formulation.

2.3.2. OFDM Transmission

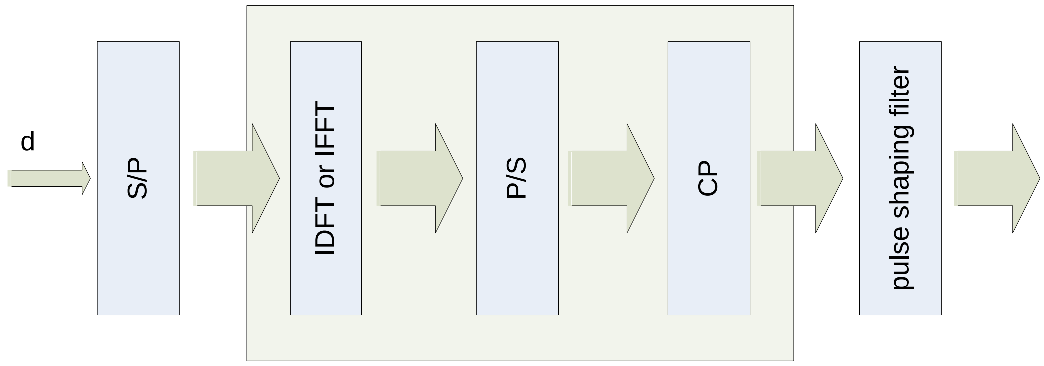

Figure 6: OFDM signal generation and transmission.

The diagram of a multicarrier modulator employing OFDM is illustrated in Fig. 6. When the number of subcarriers increases, the OFDM symbol duration T_s becomes large compared with the duration τ_0 of the channel impulse response, and the amount of ISI is therefore reduced. However, to completely avoid the effects of ISI and maintain the orthogonality between the signals on the subcarriers, i.e., to avoid ICI, a guard interval of duration

$$T_g \geq \tau_0 \tag{2.15}$$

has to be inserted between adjacent OFDM symbols. The guard interval is a cyclic extension of each OFDM symbol and is obtained by extending the duration of an OFDM symbol to

$$T_s' = T_g + T_s \quad . \tag{2.16}$$

The discrete length of the guard interval has to be

$$L_g \geq \frac{\tau_0}{T_d} \tag{2.17}$$

samples in order to avoid ISI. The new sampled sequence with cyclic extended guard interval results in

$$x_n = \sum_{l=0}^{L-1} d_l \, e^{j2\pi\frac{ln}{L}} \tag{2.18}$$

where $n = -L_g, \ldots, 0, \ldots L_s - 1$. This sequence is passed through a D/A converter, RF up-converter and transmitted to the channel.

2.3.3. OFDM Receiver

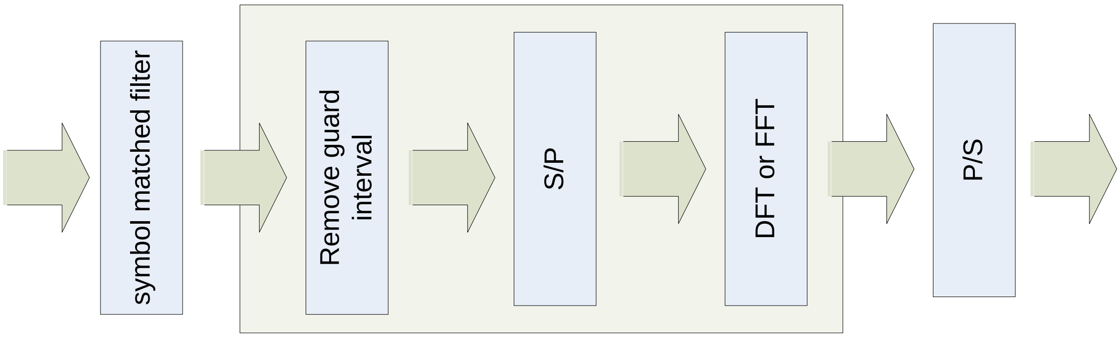

Figure 7: OFDM receiver structure.

The basic structure of an OFDM receiver is shown in Fig. 7. The output of the channel, after RF down conversion, is the received signal waveform $y(t)$ obtained from convolution of $x(t)$ with the channel impulse response $h(t)$ plus AWGN $n(t)$:

$$y(t) = \int_{-\infty}^{+\infty} x(t-\tau)\,h(\tau)\,d\tau + n(t) \; . \tag{2.19}$$

The received signal $y(t)$ is passed through an A/D converter and sampled at $f_d = 1/T_d$ with output y_m, $m = -L_g, \ldots, 0, \ldots, L_c - 1$. Since ISI is only present in the first L_g samples of the received sequences, these L_g samples are removed before OFDM demodulation. The remaining L_s ISI-free samples are OFDM demodulated by DFT or FFT to produce

$$r_n = \sum_{l=0}^{L-1} y_m e^{-j2\pi\frac{nl}{L}} \quad . \tag{2.20}$$

Since ICI can be avoided due to the guard interval and the fading on each subchannel is flat, or frequency non-selective, the received signal is

$$r = h \cdot d + n \, . \tag{2.21}$$

When ISI and ICI can be neglected, the OFDM system can be viewed as a discrete time and frequency transmission system. The information transmitted within these OFDM symbols belongs together and is referred to as an OFDM frame. So the matrix-vector notation of the transmission of one frame can be represented by

$$\mathbf{r} = \mathbf{H}\mathbf{d} + \mathbf{n} \tag{2.21}$$

where

$$\mathbf{r} = \begin{bmatrix} r_0 & r_1 & \ldots & r_{L-1} \end{bmatrix}^{\mathrm{T}} \tag{2.22}$$

$$\mathbf{H} = \begin{bmatrix} h_{0,0} & & & \\ & h_{1,1} & & \\ & & \ddots & \\ & & & h_{L-1,L-1} \end{bmatrix} \tag{2.23}$$

$$\mathbf{d} = \begin{bmatrix} d_0 & d_1 & \dots & d_{L-1} \end{bmatrix}^{\mathrm{T}} \quad . \tag{2.24}$$

2.4. SUMMARY

CDMA technology spreads the narrow-band signal into a wide signal spectrum and provides additional processing. It has many advantages over a conventional narrowband system, including universal frequency spectrum reuse, anti-jam capability, etc. However, it suffers from the near-far problem, where a near-by or strong signal source can block the reception a far-away or weak signal source. Multiuser detection is the receiver design approach for solving the near-far problem. Multiuser detection is a joint detection technology which exploits both desired signal and interfering signal structures and detects desired signals by considering interfering signal information. On the other hand, OFDM system takes a different approach to transmit signals. It converts the available signal spectrum into many orthogonal narrow-band carriers using the FFT and each carrier transmits one symbol at the same time. Since signal symbols are modulated and transmitted in narrowband with a long symbol period and there is additional time-domain guard padding available. It is known to have strong resistance to multipath effects and simplified receiver structure.

Chapter 3

MC-CDMA System

We first consider a K-user MC-CDMA single cell system, where K users are individually sending and receiving frequency-domain spread OFDM signals to and from the same base station. In the forward links of this MC-CDMA system, signals for different users synchronously arrive at each user's receiver. In the reverse links, because of the different propagation distances, these signal symbols asynchronously arrive at the base station receiver even though they are perfectly synchronized to be transmitted when they are sent by each user. In addition, there is an unknown number of asynchronous inter-cell interference from out-of-cell users to both the base station receiver and mobile station receivers. Research has shown that interfering signals from unknown out-of-cell users may account to about 40% of interference at base station or mobile station receiver [40].

In this chapter, a group-based MC-CDMA system model, in which it is assumed that the receiver may be interested in detecting signals from more than one spread signal sources, is presented. Both forward links, where signals are sent from bases station to mobile stations, and reverse links, where signals are independently sent from each mobile station to base station, are discussed. Due to the synchronous propagation nature of MC-CDMA system, it is shown that the presented MC-CDMA vector channel models for forward links and reverse links may be presented using a synchronous MC-CDMA channel model if the cyclic padding is long enough. The discussion results presented here will become the foundation for designing MC-CDMA multiuser detection schemes in the next chapter.

3.1. Signal Structure

A basic MC-CDMA signal $\mathbf{s}_k(t)$ for user k is generated by a serial concatenation of classical DS-CDMA and OFDM. Each data symbol $b_k[n]$ of user k is spread using its own spreading sequence $\mathbf{c}_k$ in the frequency domain instead of the time domain. This means that each chip of the direct sequence spread data symbol $b_k[n]\mathbf{c}_k$ is mapped onto a different subcarrier. Thus, with MC-CDMA, the chips of a spread data symbol are transmitted in parallel through different subcarriers, in contrast to a serial transmission with DS-CDMA.

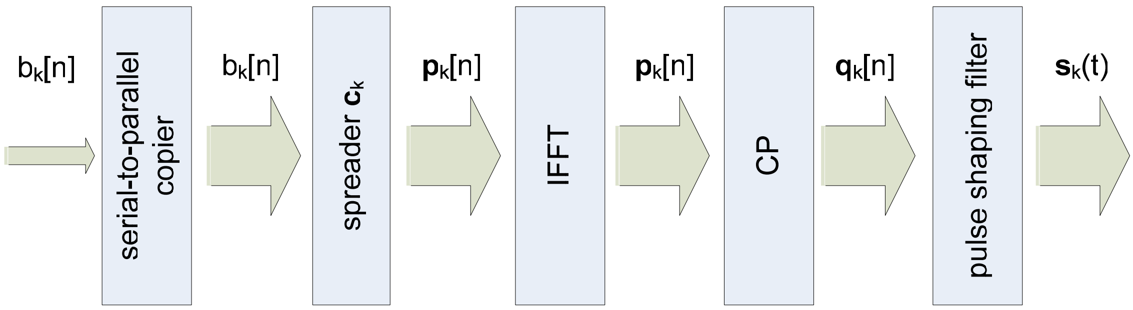

Figure 8: MC-CDMA signal generation diagram.

Fig. 8 shows multicarrier frequency domain spreading of one complex-valued data symbol $b_k[n]$ assigned to user k. The data symbol $b_k[n]$ goes through serial-to-parallel converter, spreading, IFFT, cyclic prefix and is sent out after modulation. The serial data symbol rate is

$$f_b = \frac{1}{T_b}. \tag{3.1}$$

For brevity, and without loss of generality, the MC-CDMA signal generation is described as a single data symbol per user as far as possible, such that the data symbol index can be omitted. At the transmitter, the complex-valued data symbol b_k is multiplied with the user's specified spreading code

$$\mathbf{c}_k = \begin{bmatrix} c_{k1} \\ c_{k2} \\ \vdots \\ c_{kL_c} \end{bmatrix} . \tag{3.2}$$

The frequency-domain spreading code $\mathbf{c}_k$ has length L_c; therefore the chip rate of $\mathbf{c}_k$ is

$$\begin{aligned} f_c &= \frac{1}{T_c} \\ &= \frac{L_c}{T_b} \end{aligned} . \tag{3.3}$$

The complex-valued sequence obtained after spreading is given in vector notation by

$$\mathbf{p}_k = \begin{bmatrix} p_{k1}[n] \\ p_{k2}[n] \\ \vdots \\ p_{kL_c}[n] \end{bmatrix} = b_k \mathbf{c}_k . \tag{3.4}$$

The resulting L_c-length sequence $\mathbf{p}_k$ is subject to the IFFT producing the L_c-length data sequences

$$\begin{aligned} \tilde{\mathbf{p}}_k &= \mathbf{W}_{\text{IFFT}} \mathbf{p}_k \\ &= b_k \mathbf{W}_{\text{IFFT}} \mathbf{c}_k \end{aligned} \tag{3.5}$$

with

$$\left[\mathbf{W}_{\text{IFFT}} \right]_{mn} = \frac{\sqrt{L_c}}{L_c} e^{j 2\pi \frac{mn}{L_c}} \tag{3.6}$$

for m , $n \in \{0 \quad 1 \quad \ldots \quad L_c - 1\}$. After the IFFT, a cyclic prefix of length $L_{cp} \ll L_c$, consisting of a replica of the first L_{cp} elements of $\tilde{\mathbf{p}}_k$, is appended at the end of $\tilde{\mathbf{p}}_k$ to obtain the L -length vector

$$\begin{aligned}\mathbf{q}_k &= \mathbf{T}_{\mathrm{cp}}\tilde{\mathbf{p}}_k \\ &= \begin{bmatrix} & \mathbf{I}_{L_c} & \\ \mathbf{I}_{L_{cp}} & & \mathbf{0} \end{bmatrix}\tilde{\mathbf{p}}_k\end{aligned} \tag{3.7}$$

where $L = L_{\mathrm{c}} + L_{\mathrm{cp}}$. In many MC-CDMA modulation cases, $L_{cp} = 0$.

A multicarrier spread spectrum signal

$$\begin{aligned}\mathbf{s}_k(t) &= \begin{bmatrix} s_{k1}(t) \\ s_{k2}(t) \\ \vdots \\ s_{kL}(t) \end{bmatrix} \\ &= \begin{bmatrix} q_{k1}[n]\phi(t - nT_s) \\ q_{k2}[n]\phi(t - nT_s) \\ \vdots \\ q_{kL}[n]\phi(t - nT_s) \end{bmatrix} \\ &= \mathbf{q}_k\phi(t - nT_s)\end{aligned} \tag{3.8}$$

is obtained by modulating the components $q_{kn}[n]$, n = 1, . . . , L in parallel onto L subcarriers. Therefore with multicarrier spread spectrum, each data symbol is spread over L subcarriers and L_c consecutive data symbols are IFFT-ed and cyclically prefixed into one OFDM symbol. The OFDM symbol duration with multicarrier spread spectrum including a guard interval results in

$$T_{\mathrm{s}} = T_{\mathrm{g}} + LT_{\mathrm{c}} \quad . \tag{3.9}$$

In this case, L_c data symbols per user are transmitted in one OFDM symbol.

3.2. Reverse Link

We first consider reverse-link transmission and there are K in-cell active users in the cell and K' out-of-cell active users in the system. The signals from these $K+K'$ users are individually propagating through different paths to the base station. At the base station, the baseband representation of the received signal due to the k^{th} user is given by

$$\begin{aligned}
\mathbf{r}_k(t) &= \begin{bmatrix} r_{k1}(t) \\ r_{k2}(t) \\ \vdots \\ r_{kL}(t) \end{bmatrix} \\
&= \mathbf{h}_k(t) \otimes \mathbf{s}_k(t) \\
&= \begin{bmatrix} h_{k1}(t) \otimes s_{k1}(t) \\ h_{k2}(t) \otimes s_{k2}(t) \\ \vdots \\ h_{kL}(t) \otimes s_{kL}(t) \end{bmatrix} \\
&= \begin{bmatrix} \sum_{n=1}^{N_k} \varphi_{k1}(t-\tau_{kn}) \otimes s_{k1}(t) \\ \sum_{n=1}^{N_k} \varphi_{k2}(t-\tau_{kn}) \otimes s_{k2}(t) \\ \vdots \\ \sum_{n=1}^{N_k} \varphi_{kL}(t-\tau_{kn}) \otimes s_{kL}(t) \end{bmatrix} \\
&= \sum_{n=1}^{N_k} \boldsymbol{\varphi}_k(t-\tau_{kn}) \otimes \mathbf{s}_k(t)
\end{aligned} \tag{3.10}$$

where

$$\mathbf{h}_k(t) = \begin{bmatrix} h_{k1}(t) \\ h_{k2}(t) \\ \vdots \\ h_{kL}(t) \end{bmatrix} \tag{3.11}$$

is the total multipath channel response vector for user k, with $h_{kl}(t)$ being the response for the l^{th} subcarrier of user k. This model assumes that there are a total of N_k propagation paths for user k with the n^{th} channel path response vector

$$\boldsymbol{\varphi}_k(t-\tau_{kn}) = \begin{bmatrix} \varphi_{k1}(t-\tau_{kn}) \\ \varphi_{k2}(t-\tau_{kn}) \\ \vdots \\ \varphi_{kL}(t-\tau_{kn}) \end{bmatrix} \tag{3.12}$$

and delay τ_{kn}, which is the same for each subcarrier in the n^{th} channel path.

The baseband signal received at the base station, therefore, is

$$\mathbf{r}(t) = \sum_{k=1}^{K+K'} \mathbf{r}_k(t) + \mathbf{n}(t). \tag{3.13}$$

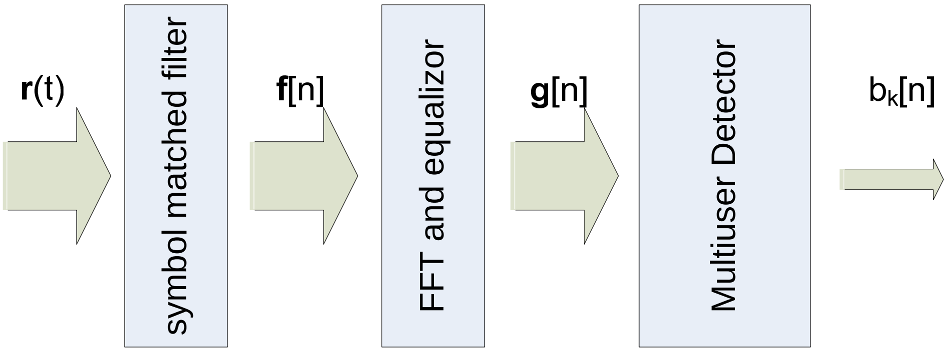

Figure 9: A basic MC-CDMA receiver for user *k*.

The structure of a basic MC-CDMA receiver with multiuser detection for user k is shown in Fig. 9. The received signal is synchronized, passed through the symbol match filter (SMF) and sampled at the symbol rate T_s into $\mathbf{f}$ of length L. The vector of the output samples of the SMF is then FFT-ed and equalized into $\mathbf{g}$ of length L_c. Because of channel distortion, the multipath effects and unknown MAI, $\mathbf{g}$ can be taken to consist of the desired component $A_k \mathbf{c}_k b_k$, the MAI $\mathbf{m}$ and the AWGN component $\mathbf{n}$. The MAI $\mathbf{m}$ can be further divided into two groups: the known MAI from K_1 known users with the spreading sequence $\mathbf{c}_i$ for user i, i=1, …, K_1, and unknown MAI from K_2 unknown users with the spreading sequence $\mathbf{c}_j$ for user j at the base station, j=1, …, K_2. $K_1 + K_2 + 1 = K + K'$ is the total number of signals received at the base station. The vector $\mathbf{g}$ can then be expressed by

$$\begin{aligned}
\mathbf{g} &= A_k \mathbf{c}_k b_k + \mathbf{m} + \mathbf{n} \\
&= A_k \mathbf{c}_k b_k + \sum_{\substack{i=1 \\ i \neq k}}^{K_1} A_i \mathbf{c}_i b_i + \sum_{j=1}^{K_2} A_j \mathbf{c}_j b_j + \mathbf{n} \qquad (3.14) \\
&= \mathbf{CAb} + \mathbf{n}
\end{aligned}$$

where

$$\mathbf{C} = \begin{bmatrix} \mathbf{C}_k & \mathbf{C}' \end{bmatrix} \tag{3.15}$$

is the spreading sequence matrix received by user k at the base station. It consists of user k's spreading sequence $\mathbf{c}_k$ and the other $K + K' - 1$ known spreading sequence $\mathbf{c}_i$, which can be collectively represented by $\mathbf{C}_k$, and the unknown spreading sequence matrix $\mathbf{C}'$. The matrix

$$\mathbf{A} = \operatorname{diag}\left\{ \begin{bmatrix} \mathbf{A}_k & \mathbf{A}' \end{bmatrix} \right\} \tag{3.16}$$

is the received amplitude diagonal matrix: $\mathbf{A}_k$ is the diagonal matrix which consists of the amplitudes of user k and the other $K - 1$ known users' amplitudes, and $\mathbf{A}'$ is the amplitude diagonal matrix for all other unknown K' interfering users. At last, $\mathbf{g}$ is passed through the multiuser detector and the final output becomes $\hat{b}_k$, the detected information for user k at time $t = nT_b$.

3.3. Forward Link

In the forward link, the base station simultaneously sends signals to all active K in-cell users. The signals for user k propagate through the same paths as the signals for other users. On the other hand, especially at the cell boundaries, there are also K' out-of-cell unknown asynchronously emitting interference signals into the cell. Therefore, the base-band representation of the received signal at the k^{th} user's receiver is generally given by

$$\begin{aligned} \mathbf{r}(t) &= \begin{bmatrix} r_1(t) \\ r_2(t) \\ \vdots \\ r_L(t) \end{bmatrix} + \mathbf{n}(t) \\ &= \sum_{i=1}^{K} \mathbf{h}(t) \otimes \mathbf{s}_i(t) + \sum_{j=1}^{K'} \mathbf{h}'_j(t) \otimes \mathbf{s}_j(t) + \mathbf{n}(t) \end{aligned} \tag{3.17}$$

which can be further expanded into

$$\mathbf{r}(t)=\begin{bmatrix}\sum_{i=1}^{K}h_1(t)\otimes s_{i1}(t)+\sum_{j=1}^{K'}h'_{j1}(t)\otimes s_{j1}(t)\\ \sum_{i=1}^{K}h_2(t)\otimes s_{i2}(t)+\sum_{j=1}^{K'}h'_{j2}(t)\otimes s_{j2}(t)\\ \vdots\\ \sum_{i=1}^{K}h_L(t)\otimes s_{iL}(t)+\sum_{j=1}^{K'}h'_{jL}(t)\otimes s_{jL}(t)\end{bmatrix}+\mathbf{n}(t)$$

$$=\begin{bmatrix}\sum_{i=1}^{K}\sum_{n=1}^{N}\varphi_1(t-\tau_n)\otimes s_{i1}(t)+\sum_{j=1}^{K'}\sum_{n=1}^{N_j}\varphi_{j1}(t-\tau_{jn})\otimes s_{j1}(t)\\ \sum_{i=1}^{K}\sum_{n=1}^{N}\varphi_2(t-\tau_n)\otimes s_{i2}(t)+\sum_{j=1}^{K'}\sum_{n=1}^{N_j}\varphi_{j2}(t-\tau_{jn})\otimes s_{j2}(t)\\ \vdots\\ \sum_{i=1}^{K}\sum_{n=1}^{N}\varphi_L(t-\tau_n)\otimes s_{iL}(t)+\sum_{j=1}^{K'}\sum_{n=1}^{N_j}\varphi_{jL}(t-\tau_{jn})\otimes s_{jL}(t)\end{bmatrix}+\mathbf{n}(t) \tag{3.18}$$

where

$$\mathbf{h}(t)=\begin{bmatrix}h_1(t)\\ h_2(t)\\ \vdots\\ h_L(t)\end{bmatrix} \tag{3.19}$$

is the total multipath channel response vector from the base station to user k with $h_l(t)$ is the response for the l^{th} subcarrier of user k. This model assumes that there are a total of N propagation paths with the n^{th} channel path response $\varphi_n(t)$ and the delay τ_n.

The received signal at user k is synchronized, passed through the symbol matched filter and sampled at the symbol rate T_s to form $\mathbf{f}$. $\mathbf{f}$ is FFT-ed and equalized to form $\mathbf{g}$. Because of channel distortion, the multipath effects and unknown MAI, the vector $\mathbf{g}$ can then be expressed by

$$
\begin{aligned}
\mathbf{g} &= A_k \mathbf{c}_k b_k + \mathbf{m} + \mathbf{n} \\
&= A_k \mathbf{c}_k b_k + \sum_{\substack{i=1 \\ i \neq k}}^{K} A_i \mathbf{c}_i b_i + \sum_{j=1}^{K'} A_j \mathbf{c}_j b_j + \mathbf{n} \\
&= \mathbf{CAb} + \mathbf{n}
\end{aligned}
\tag{3.20}
$$

where $\mathbf{m}$ represents MAI and can be written as

$$
\mathbf{m} = \sum_{\substack{i=1 \\ i \neq k}}^{K} A_i \mathbf{c}_i b_i + \sum_{j=1}^{K'} A_j \mathbf{c}_j b_j \tag{3.21}
$$

where

$$
\mathbf{C} = \begin{bmatrix} \mathbf{c}_k & \mathbf{C}' \end{bmatrix} \tag{3.22}
$$

is the spreading sequence matrix received by user k with $\mathbf{C}'$ for the other $K + K' - 1$ unknown users. The matrix

$$
\mathbf{A} = \operatorname{diag}\left\{ \begin{bmatrix} A_k & \mathbf{A}' \end{bmatrix} \right\} \tag{3.23}
$$

is the received amplitude diagonal matrix where $\mathbf{A}'$ is the amplitude diagonal matrix for other $K + K' - 1$ unknown users. $\mathbf{g}$ is sent to the multiuser detector and the final output becomes $\hat{b}_k$ for user k at the time nT_b .

3.4. Detection Strategies

From the previous discussion, obviously the forward link can be taken as a special case of the reverse link of MC-CDMA with the equal propagation delays. Hence, the MC-CDMA forward link or reverse link can be generally represented by

$$
\begin{aligned}
\mathbf{g} &= \sum_{k=1}^{K''=K+K'} A_k \mathbf{c}_k b_k + \mathbf{n} \\
&= \sum_{k=1}^{K} A_k \mathbf{c}_k b_k + \sum_{k'=1}^{K'} A_{k'} \mathbf{c}_{k'} b_{k'} + \mathbf{n} \\
&= \mathbf{CAb} + \mathbf{n}
\end{aligned}
\tag{3.24}
$$

where $\mathbf{c}_k$, $k = 1, 2, \cdots K$, denotes the K known spreading sequences at the receiver and $\mathbf{c}_{k'}$, $k' = 1, 2, \cdots K'$, denotes the spreading sequences for the K' unknown interfering signals. Now the multiuser detection question is how to efficiently detect the K information bits for the K desired users.

There are several detection strategies for MC-CDMA systems. For synchronous downlink transmissions, due to the ease of channel estimation, diversity-combining schemes can be implemented at the receiver, such as orthogonality restoring combining (ORC), controlled equalization (CE), equal gain combining (EGC), maximum ratio combining (MRC) and minimal mean square error combining (MMSEC) [1, 4, 9, 10]. For asynchronous uplink transmissions, because of the loss of orthogonality among users, more advanced detection strategies, such as multiuser detection, are needed in order to maintain quality of service.

3.5. SUMMARY

The signal structure of a MC-CDMA system is discussed in this chapter. MC-CDMA can be taken as a concatenation of DS-CDMA and OFDM. Both the forward link and reverse link of MC-CDMA system are discussed and compared. The forward link of MC-CDMA is a synchronous signal structure where all user signals arrive at receiver with same delays, while the reverse link is asynchronous structure. Besides this, several conventional detection strategies for MC-CDMA are also presented.

Chapter 4

Semiblind Multiuser Detection for MC-CDMA

It is widely known that the large gap in performance and complexity between the conventional single-user matched filter and optimum multiuser detectors encourages the search for other MUDs that exhibit good performance/complexity tradeoffs. Optimal and conventional multiuser detectors can achieve good performance but require too much prior knowledge of other users which makes them unfit for practical applications. Alternatively, blind multiuser detectors only require the spreading sequence and timing of the desired user(s); they can achieve reasonable performance which may be close to that of conventional detectors. However, blind multiuser detectors usually need some additional estimation, training or search procedures before detecting desired user(s)' data [16, 29]; this normally requires too much additional computation and time for practical implementation. Here, we propose an alternative approach, semiblind multiuser detection, in which the amplitudes of desired user(s) need to be known beforehand besides spreading sequences and timing, but no information of other users is required.

It is known that the conventional decorrelating detector is a simple and natural strategy for multiuser detection. It is based on the classic least squares estimation scheme. The derivation of the asymptotic efficiency of the decorrelating detector for synchronous channels and the proof of its optimum near-far resistance property were done by Verdu in the case of nonsingular covariance matrices [29]. In this chapter, we first propose a semiblind multiuser detection framework. With this framework and the same least-square estimation principle for conventional decorrelating detection, we develop several least-square-based semiblind multiuser detectors for MC-CDMA, including least-square (LS) semiblind multiuser detector, total least-square (TLS) semiblind multiuser detector and mixed LS/TLS semiblind multiuser detector. The detectors are simple and direct without using any additional estimation procedure as many other blind or semiblind schemes do. In the following, the semiblind multiuser detection framework and algorithms will be discussed in detail.

4.1. Semiblind Multiuser Detection Framework

It is assumed that there are a total of $K'' = K + K'$ user signals received by the multiuser receiver. Among these K'' user signals, the first K signals are from desired users while the other K' signals are interference. Without loss of generality, we assume that only bits sent by the first K known users, whose spreading sequences, amplitudes and timing are already known, are considered in the following description. A new $L \times P$ blind spreading sequence matrix $\mathbf{S}$ is constructed to take place of the original spreading sequence matrix $\mathbf{C}$ in (3.24) by

$$\begin{aligned}
\mathbf{S} &= \begin{bmatrix} \mathbf{s}_1 & \mathbf{s}_2 & \cdots & \mathbf{s}_P \end{bmatrix} \\
&= \begin{bmatrix} A_1\mathbf{c}_1 & \cdots & A_K\mathbf{c}_K & \mathbf{g}_1 & \cdots & \mathbf{g}_{P-K} \end{bmatrix} \\
&= \mathbf{C}\begin{bmatrix} \mathbf{AE} & \mathbf{Ab}_1 & \cdots & \mathbf{Ab}_{P-K} \end{bmatrix} + \mathbf{N} \\
&= \mathbf{CA}\begin{bmatrix} \mathbf{E} & \mathbf{D} \end{bmatrix} + \mathbf{N} \\
&= \mathbf{CAB} + \mathbf{N}
\end{aligned} \tag{4.1}$$

where $\mathbf{g}_p$, p =1, …, $P-K$, are $P-K$ previously received signal vectors and $\mathbf{b}_p$ are the corresponding information vectors, $P > K''$,

$$\mathbf{E} = \begin{bmatrix} \mathbf{I}_K & \mathbf{0} \end{bmatrix}^{\mathrm{T}}, \tag{4.2}$$

$$\begin{aligned}
\mathbf{D} &= \begin{bmatrix} \mathbf{b}_1 & \mathbf{b}_2 & \cdots & \mathbf{b}_{P-K} \end{bmatrix} \\
&= \begin{bmatrix} \bar{\mathbf{D}} \\ \tilde{\mathbf{D}} \end{bmatrix}
\end{aligned} \tag{4.3}$$

is the information matrix for $\mathbf{g}_p$ with $\bar{\mathbf{D}}$ being the $K \times (P-K)$ known data sub-matrix which denotes the information previously detected by user 1. Also,

$$\begin{aligned}\mathbf{B} &= [\mathbf{E} \quad \mathbf{D}] \\ &= \begin{bmatrix} \mathbf{I} & \bar{\mathbf{D}} \\ \mathbf{0} & \tilde{\mathbf{D}} \end{bmatrix}.\end{aligned} \tag{4.4}$$

Since the spreading matrix $\mathbf{S}$ in (4.1) is constructed with K known users' spreading sequences, $\mathbf{c}_i$, and amplitude A_k and $P-K$ previously received signal vector $\mathbf{g}_p$ and there is no information regarding other unknown users being involved, $\mathbf{S}$ is called a semiblind spreading matrix. Compared to the original spreading matrix $\mathbf{C}$, there may be a different blind signature matrix for each user. Even for the same user, the semiblind signature matrix is not fixed and it can change from time to time.

After defining the semiblind spreading matrix, the relationship between the current FFT-ed and equalized signal vector $\mathbf{g}$ and $\mathbf{S}$ can be written as

$$\begin{aligned}\mathbf{g} &= \mathbf{CAb} + \mathbf{n} \\ &= (\mathbf{S} - \mathbf{N})\mathbf{B}^{+}\mathbf{Ab} + \mathbf{n} \\ &= \mathbf{SB}^{+}\mathbf{Ab} - \mathbf{NB}^{+}\mathbf{Ab} + \mathbf{n} \\ &= \mathbf{Sd} + \tilde{\mathbf{n}}\end{aligned} \tag{4.5}$$

where $\mathbf{B}^{+}$ is the pseudo-inverse of $\mathbf{B}$, $\mathbf{d}$ denotes the new $P \times 1$ detection vector defined by

$$\mathbf{d} = \mathbf{B}^{+}\mathbf{Ab} \tag{4.6}$$

and $\tilde{\mathbf{n}}$ is the new blind noise vector defined by

$$\tilde{\mathbf{n}} = -\mathbf{NB}^{+}\mathbf{Ab} + \mathbf{n}\,. \tag{4.7}$$

Since we know $\mathbf{d} = \mathbf{B}^{+}\mathbf{A}\mathbf{b}$, then

$$\begin{aligned}\mathbf{d} &= \begin{bmatrix} \bar{\mathbf{d}} \\ \tilde{\mathbf{d}} \end{bmatrix} \\ &= \begin{bmatrix} \mathbf{I} & \bar{\mathbf{D}} \\ \mathbf{0} & \tilde{\mathbf{D}} \end{bmatrix}^{+} \begin{bmatrix} \mathbf{A}_1 & \\ & \tilde{\mathbf{A}} \end{bmatrix} \begin{bmatrix} \mathbf{b}_1 \\ \tilde{\mathbf{b}} \end{bmatrix}\end{aligned} \quad . \tag{4.8}$$

Furthermore,

$$\begin{bmatrix} \mathbf{A}_1 & \\ & \tilde{\mathbf{A}} \end{bmatrix} \begin{bmatrix} \mathbf{b}_1 \\ \tilde{\mathbf{b}} \end{bmatrix} = \begin{bmatrix} \mathbf{I} & \bar{\mathbf{D}} \\ \mathbf{0} & \tilde{\mathbf{D}} \end{bmatrix} \begin{bmatrix} \bar{\mathbf{d}} \\ \tilde{\mathbf{d}} \end{bmatrix} \quad . \tag{4.9}$$

Hence, if $\mathbf{d}$ can be known, $\mathbf{A}_1\mathbf{b}_1$ can be estimated by

$$\begin{aligned}\mathbf{A}_1\mathbf{b}_1 &= \mathbf{F}\mathbf{d} \\ &= \begin{bmatrix} \mathbf{I} & \bar{\mathbf{D}} \end{bmatrix} \begin{bmatrix} \bar{\mathbf{d}} \\ \tilde{\mathbf{d}} \end{bmatrix} \\ &= \bar{\mathbf{d}} + \bar{\mathbf{D}}\tilde{\mathbf{d}}\end{aligned} \quad . \tag{4.10}$$

As the new semiblind spreading matrix $\mathbf{S}$ is defined in (4.1), the conventional multiuser detection model becomes (4.5). Comparing (3.24) with (4.5), the major difference is that the signal bit vector $\mathbf{b}$ is replaced by the detection vector $\mathbf{d}$ and the original AWGN vector $\mathbf{n}$ is replaced by the new noise vector $\tilde{\mathbf{n}}$. Fortunately, it is still possible to determine $\mathbf{b}_1$ with the detection vector $\mathbf{d}$ and the previously detected data matrix $\bar{\mathbf{D}}$. A comparison between the conventional model and the proposed model is shown in Table 1. The next question is how to estimate the detection vector $\mathbf{d}$ as efficiently as possible.

Table 1: Comparison between the conventional model and proposed model.

	Conventional Model	Proposed Semiblind Model
Dedicated/Common	Common, fixed	Dedicated, not fixed
Input Vector	$\mathbf{b}$	$\mathbf{D}$
Amplitude Matrix	$\mathbf{A}$	N/A
Spreading Matrix	$\mathbf{C}$	$\mathbf{S}$
Noise Vector	$\mathbf{n}$	$\tilde{\mathbf{n}}$
Output	$\mathbf{g}$	$\mathbf{G}$

With (4.5), if there is no noise in the semiblind signatures and received signal, the vector $\mathbf{g}$ can be taken as a linear combination of the columns of the blind spreading matrix $\mathbf{S}$. With noise present, the estimation of $\mathbf{d}$ becomes a typical subspace fitting problem which can be expressed by

$$\hat{\mathbf{d}} = \arg\min_{\mathbf{x}\in\mathbb{R}^P} \|\mathbf{Sx}-\mathbf{g}\|_p \quad . \tag{4.11}$$

Obviously different norms render different optimum solutions. The general p-norm is complicated by the fact that the function

$$\mathrm{f}(\mathbf{x}) = \|\mathbf{Sx}-\mathbf{g}\|_p \tag{4.12}$$

is not easy to differentiate. In contrast to general p-norm minimization, the 2-norm problem is more tractable because

1) $\mathrm{f}(\mathbf{x})$ is a differentiable function of $\mathbf{x}$.
2) The 2-norm is preserved under orthogonal transformations.

There are the following three schemes to estimate the detection vector $\mathbf{d}$ based on the least squares criteria.

4.2. Least-Squares Estimation

At first, the measurement of $\mathbf{S}$ is assumed to be free of error and noise. All errors and noise are assumed to be confined to the received vector $\mathbf{g}$. This leads to the following least square (LS) estimation problem:

$$\mathbf{d}_{\mathrm{LS}} = \arg\min_{\mathbf{x}\in\mathbb{R}^{P\times 1}} \left\|\mathbf{g} - \mathbf{S}\mathbf{x}\right\|_2 \ . \tag{4.13}$$

The LS solution is:

$$\begin{aligned}\mathbf{d}_{\mathrm{LS}} &= \mathbf{S}^{-1}\mathbf{g} \\ &= \mathbf{d} + \mathbf{S}^{-1}\overline{\mathbf{n}}\end{aligned} \ . \tag{4.14}$$

When $P \le L$ and $\operatorname{rank}\{\mathbf{S}\} < P$, (4.13) becomes a rank-deficient LS problem and there is an infinite number of solutions. Here, we use singular value decomposition technique to obtain the minimum 2-norm solution according to the following [27, 35].

Suppose $\mathbf{U}^{\mathrm{T}}\mathbf{S}\mathbf{V} = \mathbf{\Lambda}$ is the SVD of the semiblind spreading matrix $\mathbf{S}\in\mathbb{R}^{L\times P}$ with $r = \operatorname{rank}\{\mathbf{S}\}$. If $\mathbf{U} = \begin{bmatrix}\mathbf{u}_1 & \mathbf{u}_2 & \ldots & \mathbf{u}_L\end{bmatrix}$ and $\mathbf{V} = \begin{bmatrix}\mathbf{v}_1 & \mathbf{v}_2 & \ldots & \mathbf{v}_P\end{bmatrix}$ are column partitioned, then

$$\begin{aligned}\arg\min\left\|\mathbf{S}\mathbf{x} - \mathbf{g}\right\|_2^2 &= \arg\min\left\|\left(\mathbf{U}^{\mathrm{T}}\mathbf{S}\mathbf{V}\right)\left(\mathbf{V}^{\mathrm{T}}\mathbf{x}\right) - \mathbf{U}^{\mathrm{T}}\mathbf{g}\right\|_2^2 \\ &= \arg\min\left\|\mathbf{\Lambda}\left(\mathbf{V}^{\mathrm{T}}\mathbf{x}\right) - \mathbf{U}^{\mathrm{T}}\mathbf{g}\right\|_2^2 \\ &= \arg\min\left[\sum_{i=1}^{r}\left(\sigma_i\,\alpha_i - \mathbf{u}_i^{\mathrm{T}}\mathbf{g}\right)^2 + \sum_{j=r+1}^{P}\left(\mathbf{u}_j^{\mathrm{T}}\mathbf{g}\right)^2\right]\end{aligned} \tag{4.15}$$

where $\boldsymbol{\Lambda} = \mathrm{diag}\left\{\begin{bmatrix}\sigma_1 & \cdots & \sigma_r & 0 & \cdots & 0\end{bmatrix}\right\}$ consisting of singular values σ_i and $\alpha_i = \mathbf{v}_i^{\mathrm{T}}\mathbf{g}$. Hence, the LS solution is

$$\mathbf{d}_{\mathrm{LS}} = \sum_{i=1}^{r} \frac{\mathbf{u}_i^{\mathrm{T}}\mathbf{g}}{\sigma_{\mathrm{i}}} \mathbf{v}_i \quad . \tag{4.16}$$

The estimation error is

$$\varepsilon_{\mathrm{LS}}^2 = \left\|\mathbf{S}\mathbf{d}_{\mathrm{LS}} - \mathbf{g}\right\|_2^2 = \sum_{j=r+1}^{P} \left(\mathbf{u}_j^{\mathrm{T}}\mathbf{g}\right)^2 . \tag{4.17}$$

Finally, $\mathbf{b}_1$ can be estimated by

$$\begin{aligned}\mathbf{b}_{1\mathrm{LS}} &= \mathrm{sgn}\left\{\mathbf{F}\mathbf{d}_{\mathrm{LS}}\right\} \\ &= \mathrm{sgn}\left\{\mathbf{F}\left(\sum_{i=1}^{r} \frac{\mathbf{u}_i^{\mathrm{T}}\mathbf{g}}{\sigma_{\mathrm{i}}} \mathbf{v}_i\right)\right\}\end{aligned} \quad . \tag{4.18}$$

4.3. Total Least Squares (TLS) Estimation

In (4.13), it assumed that the blind spreading matrix $\mathbf{S}$ is error-free. However, this assumption obviously is not true by its definition in (4.1) where the noise or error matrix $\mathbf{N}$ is present. Hence, $\mathbf{g}$ can be expressed by

$$\begin{aligned}\mathbf{g} &= \left(\mathbf{S} - \mathbf{N}\right)\mathbf{d} + \mathbf{n} \\ &= \hat{\mathbf{S}}\mathbf{d} + \mathbf{n}\end{aligned} \tag{4.19}$$

where $\hat{\mathbf{S}} = \mathbf{S} - \mathbf{N} = \mathbf{CAB}$. Now, the estimation of $\mathbf{d}$ becomes the following TLS problem:

$$\mathbf{d}_{\mathrm{TLS}} = \arg\min_{\mathbf{Y},\mathbf{x}} \left\| [\mathbf{S} \quad \mathbf{g}] - [\mathbf{Y} \quad \mathbf{Yx}] \right\|_2 \ . \tag{4.20}$$

Additionally, (4.13) can be equivalently written as

$$[\mathbf{S} \quad \mathbf{g}] \begin{bmatrix} \mathbf{d} \\ -1 \end{bmatrix} = -\mathbf{n} \, . \tag{4.21}$$

The TLS solution of (4.21) is found from the SVD of the augmented data matrix

$$\begin{aligned} \mathbf{Z} &= [\mathbf{S} \quad \mathbf{g}] \\ &= \mathbf{U}\boldsymbol{\Lambda}\mathbf{V}^{\mathrm{T}} \\ &= [\mathbf{U}_1 \quad \mathbf{U}_2] \begin{bmatrix} \boldsymbol{\Lambda}_1 & \\ & \boldsymbol{\Lambda}_2 \end{bmatrix} \begin{bmatrix} \mathbf{V}_{11}^{\mathrm{T}} & \mathbf{V}_{21}^{\mathrm{T}} \\ \mathbf{V}_{12}^{\mathrm{T}} & \mathbf{V}_{22}^{\mathrm{T}} \end{bmatrix} \end{aligned} \ . \tag{4.22}$$

We assume that the $K''+1$ singular values are sorted and the remaining smallest $K''-r$ singular values are equal, i.e.,

$$\boldsymbol{\Lambda}_1 = \mathrm{diag}\left\{ \begin{bmatrix} \sigma_1 & \sigma_2 & \cdots & \sigma_{\mathrm{r}} \end{bmatrix} \right\} \ , \tag{4.23}$$

$$\boldsymbol{\Lambda}_2 = \mathrm{diag}\left\{ \begin{bmatrix} \sigma_{\mathrm{r}+1} & \sigma_{\mathrm{r}+2} & \cdots & \sigma_{\mathrm{K}''+1} \end{bmatrix} \right\} \ , \tag{4.24}$$

$$\mathbf{V} = \begin{array}{c} \begin{bmatrix} \mathbf{V}_{11} & \mathbf{V}_{12} \\ \mathbf{V}_{21} & \mathbf{V}_{22} \end{bmatrix} \begin{array}{l} \}K'' \\ \}1 \end{array} \\ \underbrace{\qquad}_{r} \underbrace{\qquad}_{K''+1-r} \qquad \end{array} \tag{4.25}$$

and $\sigma_1 \geq \sigma_2 \geq \cdots \geq \sigma_r \geq \sigma_{r+1} = \cdots = \sigma_{K''+1}$. Then, the minimum norm TLS solution is

$$
\begin{aligned}
\mathbf{d}_{1\mathrm{TLS}} &= -\mathbf{V}_{12}\mathbf{V}_{22}^{+} \\
&= -\mathbf{V}_{12}\mathbf{V}_{22}^{\mathrm{T}}\left(\mathbf{V}_{22}\mathbf{V}_{22}^{\mathrm{T}}\right)^{-1}
\end{aligned}
\tag{4.26}
$$

or

$$
\begin{aligned}
\mathbf{d}_{1\mathrm{TLS}} &= \left(\mathbf{V}_{11}^{\mathrm{T}}\right)^{+}\mathbf{V}_{21}^{\mathrm{T}} \\
&= \mathbf{V}_{11}\mathbf{V}_{21}^{\mathrm{T}}\left(\mathbf{I}-\mathbf{V}_{21}\mathbf{V}_{21}^{\mathrm{T}}\right)^{-1} \quad . \\
&= \left(\mathbf{S}^{\mathrm{T}}\mathbf{S}-\sigma_{r+1}^{2}\mathbf{I}\right)^{-1}\mathbf{S}^{\mathbf{T}}\mathbf{g}
\end{aligned}
\tag{4.27}
$$

4.4. Mixed TLS/LS Estimation

In the LS problem of (4.13), it assumed the blind spreading matrix $\mathbf{S}$ is error-free. Again, this assumption is not completely accurate. In the TLS problem of (4.20), it assumed that in each column of the semiblind spreading matrix $\mathbf{S}$, some noise or error exists. This assumption is also incomplete. Though there is a noise or error matrix $\mathbf{N}$ in $\mathbf{S}$ from (4.1), its first K columns are exactly known to be noise-free or error-free. Hence, to maximize the estimation accuracy of the detection vector $\mathbf{d}$, it is natural to require that the corresponding columns of $\mathbf{S}$ be unperturbed since they are known exactly. The problem of estimating the detection vector $\mathbf{d}$ can then be transformed into the following MLS problem by considering (4.13) and (4.20):

$$
\mathbf{d}_{\mathrm{MLS}} = \arg\min_{\tilde{\mathbf{Y}},\mathbf{x}} \left\| \begin{bmatrix}\mathbf{S} & \mathbf{g}\end{bmatrix} - \begin{bmatrix}\begin{bmatrix}\mathbf{C}_1 & \tilde{\mathbf{Y}}\end{bmatrix} & \mathbf{Y}\mathbf{x}\end{bmatrix} \right\|_2 \quad . \tag{4.28}
$$

Consider the MLS problem in (4.21) and perform the Householder transformation $\mathbf{Q}$, defined by

$$
\begin{aligned}
\mathbf{Q} &= \prod_{k=1}^{K}\mathbf{Q}_k \\
&= \prod_{k=1}^{K}\left(1-2\mathbf{s}_k\mathbf{s}_k^{\mathrm{T}}\right)
\end{aligned}
$$

on the matrix $\begin{bmatrix}\mathbf{S} & \mathbf{g}\end{bmatrix}$ so that

$$\mathbf{Q}^{\mathrm{H}}\begin{bmatrix}\mathbf{S} & \mathbf{g}\end{bmatrix}=\begin{bmatrix}\mathbf{R}_{11} & \mathbf{R}_{12} & \mathbf{r}_{1g}\\ \mathbf{0} & \mathbf{R}_{22} & \mathbf{r}_{2g}\end{bmatrix} \tag{4.29}$$

where $\mathbf{R}_{12}$ is a $K''\times(P-K'')$ matrix, $\mathbf{R}_{22}$ is a $(L-K'')\times(P-K'')$ matrix, $\mathbf{r}_{1g}$ is a $K''\times 1$ vector and $\mathbf{r}_{2g}$ is a $(L-K'')\times 1$ vector.

Denote σ'' as the smallest singular value of $\mathbf{R}_{22}$ and σ' as the smallest singular value of $\begin{bmatrix}\mathbf{R}_{22} & \mathbf{r}_{2g}\end{bmatrix}$. If $\sigma''>\sigma'$, then the MLS solution uniquely exists and is given by

$$\mathbf{d}_{\mathrm{MLS}}=\left(\mathbf{S}^{\mathrm{T}}\mathbf{S}-\sigma'^{2}\begin{bmatrix}\mathbf{0} & \\ & \mathbf{I}\end{bmatrix}\right)^{-1}\mathbf{S}^{\mathrm{T}}\mathbf{g} \quad . \tag{4.30}$$

4.5. SUMMARY

Several least-square-based semiblind MC-CDMA multiuser detectors were presented after a new semiblind signal signature matrix was given. With different assumptions regarding the proposed semiblind signal signature matrix, least-squares, total-least-squares and mixed-least-squares detectors are constructed, individually. Compared to conventional multiuser detection approaches, only the desired users' signal amplitudes, timing and signal signatures are required. Compared with existing semiblind or blind multiuser detectors, there is no additional channel estimation necessary. Therefore, the detection complexity is greatly simplified.

Chapter 5

Performance Analysis

In this chapter, we investigate the performance of all the proposed algorithms. Similar to the conventional decorrelating detector, we give a geometric explanation of the proposed LS semiblind multiuser detector since both the TLS and MLS semiblind multiuser detectors can be taken as variation of the LS semiblind detection. We know that the noise term in conventional decorrelating detection plays an important role in the performance; we therefore analyze the covariance matrix of the new noise item in (4.7). It shows that the performance of the proposed least-square semiblind detector may be improved by increasing P .

5.1. Comparison with Decorrelating Detection

Compared with the conventional multiuser detection model, the proposed semiblind multiuser detection model actually uses a specially designed semiblind spreading sequence matrix $\mathbf{S}$ instead of the original spreading sequence matrix $\mathbf{C}$. In order to keep the original transmitted signal $\mathbf{g}$, it is assumed that the detection vector $\mathbf{d}$ is spread by $\mathbf{S}$ while the original data $\mathbf{b}$ is spread by $\mathbf{C}$. This can be illustrated by Fig. 10.

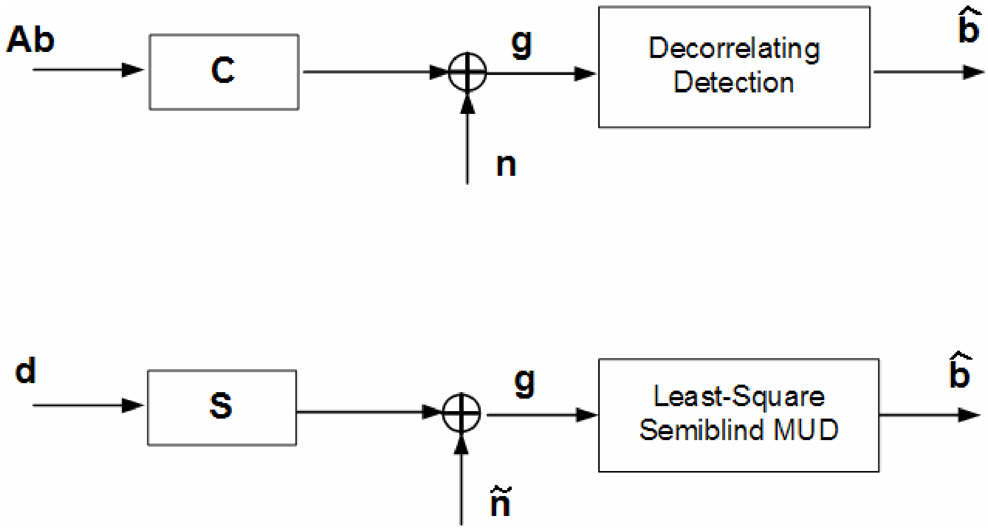

Figure 10: Comparison between the conventional decorrelating detector and the proposed least squares detector.

The bit-error rate of the conventional decorrelating detector is [29]

$$P_{ek}^{\mathrm{DD}} = Q\left(\frac{\mathrm{A}_{\mathrm{k}}}{\sigma\left[\mathbf{R}^{-1}\right]_{kk}}\right) \tag{5.1}$$

where $\left[\mathbf{R}^{-1}\right]_{kk}$ denotes the k^{th} diagonal component of $\mathbf{R}^{-1} = \left(\mathbf{C}^{\mathrm{T}}\mathbf{C}\right)^{-1}$. The proposed least-squares semiblind multiuser detection can be expressed by

$$\begin{aligned} \mathbf{b}_{1\mathrm{LS}} &= \operatorname{sgn}\left\{\mathbf{F}\mathbf{d}_{\mathrm{LS}}\right\} \\ &= \operatorname{sgn}\left\{\mathbf{F}\left(\sum_{i=1}^{r}\frac{\mathbf{u}_i^{\mathrm{T}}\mathbf{g}}{\sigma_{\mathrm{i}}}\mathbf{v}_i\right)\right\}. \end{aligned} \tag{5.2}$$

If there is no noise in the semiblind spreading matrix $\mathbf{S}$, the least-squares semiblind multiuser detection can be expressed by

$$\mathbf{b}_{1\mathrm{LS}} = \operatorname{sgn}\left\{\mathbf{b} + \mathbf{S}^{-1}\mathbf{n}\right\}. \tag{5.3}$$

Thus, the bit-error rate of the lease-squares semi-blind spreading is

$$P_{e1}^{\mathrm{LS}} = Q\left(\frac{\mathrm{A}_1}{\sigma \sum_{i=1}^{L} \left[\mathbf{S}^{+} \right]_{1i}} \right) . \tag{5.3}$$

Obviously the conventional decorrelating detector can be taken as a special case of least-squares semiblind detector with $\mathbf{B} = \mathbf{I}$ in (4.1) and (4.18).

5.2. Geometric Explanations

In decorrelating detection, the data $\mathbf{Ab}$ is spread by $\mathbf{C}$ and estimated by

$$\begin{aligned} \hat{\mathbf{A}}\hat{\mathbf{b}} &= \mathbf{W}_{\mathrm{DD}}^{\mathrm{T}}\mathbf{g} \\ &= \mathbf{C}^{+}\mathbf{g}. \end{aligned} \tag{5.4}$$

It is easy to see that the least-square semiblind estimate of $\mathbf{d}$ is given by

$$\begin{aligned} \hat{\mathbf{d}} &= \mathbf{W}_{\mathrm{d,\,LS}}^{\mathrm{T}}\mathbf{g} \\ &= \mathbf{S}^{+}\mathbf{g}. \end{aligned} \tag{5.5}$$

Clearly, this is similar to the conventional decorrelating detector. Hence, each element of $\hat{\mathbf{d}}$ can be taken as the projection of the received vector $\mathbf{g}$ onto the subspace, which is orthogonal to the codes of the other users [29, 39]. Furthermore, in [38], it is shown that conventional decorrelating detector is ***the oblique projection*** of the desired user's signature vector onto the space spanned by other users' signature vectors along the complement of this space. For example, the first column of $\mathbf{w}_{\mathrm{1d,\,LS}}$ for the first user is

$$
\begin{aligned}
\mathbf{w}_{1\mathrm{d,\,LS}} &= \mathbf{P}_1^{O}\mathbf{s}_1 \\
&= \frac{1}{\mathbf{s}_1^{\mathrm{T}}\mathbf{P}_1^{\perp}\mathbf{s}_1}\mathbf{P}_1^{\perp}\mathbf{s}_1
\end{aligned}
\tag{5.6}
$$

where $\mathbf{P}_1^{O}$ denotes the oblique projection of $\mathbf{s}_1$, and $\mathbf{P}_1^{\perp}$ is the orthogonal projection for user 1 onto the orthogonal complement of the subspace spanned by the other users' signature vectors. This is shown in Fig. 11.

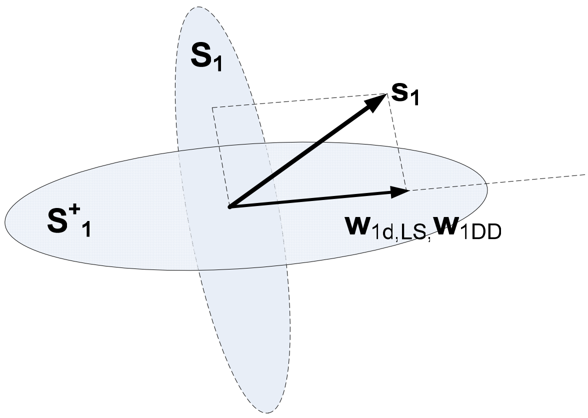

Figure 11: The geometric explanation of the LS estimation of $\mathbf{d}$ and the decorrelating detector.

5.2. Analysis of the New Noise Vector $\tilde{\mathbf{n}}$

When the conventional multiuser models in (2.11) and (2.19) are replaced by (4.5), the original AWGN vector $\mathbf{n}$ becomes $\tilde{\mathbf{n}}$ defined by (4.7). Certainly $\tilde{\mathbf{n}}$ is still an AWGN vector since the whole system model is a linear system based on an AWGN channel. It is important to know how this change affects the performance of the proposed semiblind multiuser detection algorithms. To answer this question, the first thing to know is the covariance matrix $\mathbf{C}_{\tilde{\mathbf{n}}}$ of $\tilde{\mathbf{n}}$ is, which is defined by

$$\begin{aligned}\mathbf{C}_{\tilde{\mathbf{n}}} &= \mathrm{E}\left\{\tilde{\mathbf{n}}\tilde{\mathbf{n}}^{\mathrm{T}}\right\} \\ &= \mathrm{E}\left\{\left(-\mathbf{N}\mathbf{B}^{+}\mathbf{A}\mathbf{b}+\mathbf{n}\right)\left(-\mathbf{N}\mathbf{B}^{+}\mathbf{A}\mathbf{b}+\mathbf{n}\right)^{\mathrm{T}}\right\} \\ &= \mathrm{E}\left\{\mathbf{N}\mathbf{B}^{+}\mathbf{A}^{2}\mathbf{B}^{+\mathrm{T}}\mathbf{N}^{\mathrm{T}}\right\}+\sigma^{2}\mathbf{I}\end{aligned} \quad . \tag{5.7}$$

Though the power spectrum density of $\mathbf{B}$ may be determined, the PDF of $\mathbf{B}^{+}$ is largely unknown. So it is difficult to decide the closed-form solution of $\mathbf{C}_{\tilde{\mathbf{n}}}$. On the other hand, with Girko's law, when $\alpha=(P-K)/(P-K'')$ is fixed with $K,P\to\infty$, the diagonal element of $\frac{1}{P-K''}\tilde{\mathbf{d}}\tilde{\mathbf{d}}^{\mathrm{T}}$ may be approximated by

$$\lim \tfrac{1}{P-K''}\left[\tilde{\mathbf{d}}\tilde{\mathbf{d}}^{\mathrm{T}}\right]_{ii}^{-1}=1-\alpha . \tag{5.8}$$

Then,

$$
\begin{aligned}
\mathbf{C}_{\tilde{\mathbf{n}}} &= \mathrm{E}\left\{ \mathbf{N} \begin{bmatrix} \mathbf{I} & \bar{\mathbf{D}} \\ \mathbf{0} & \tilde{\mathbf{D}} \end{bmatrix}^{+} \begin{bmatrix} \mathbf{A}_1 & \\ & \tilde{\mathbf{A}} \end{bmatrix} \begin{bmatrix} \mathbf{b}_1 \\ \tilde{\mathbf{b}} \end{bmatrix} \begin{bmatrix} \mathbf{b}_1 \\ \tilde{\mathbf{b}} \end{bmatrix}^{\mathrm{T}} \begin{bmatrix} \mathbf{A}_1 & \\ & \tilde{\mathbf{A}} \end{bmatrix}^{\mathrm{T}} \begin{bmatrix} \mathbf{I} & \bar{\mathbf{D}} \\ \mathbf{0} & \tilde{\mathbf{D}} \end{bmatrix}^{+\mathrm{T}} \mathbf{N}^{\mathrm{T}} \right\} + \sigma^2 \mathbf{I} \\
&= \mathrm{E}\left\{ \mathbf{N} \begin{bmatrix} \frac{2P-K-K''}{P-K} \mathbf{A}_1^2 & \\ & \frac{1}{P-K} \mathbf{I} \end{bmatrix} \mathbf{N}^{\mathrm{T}} \right\} + \sigma^2 \mathbf{I} \\
&= \left(1 + \tfrac{P-K''}{P-K}\right) \sigma^2 \mathbf{I}
\end{aligned} \qquad . \qquad (5.9)
$$

Hence, we can see that

$$\lim_{P \to \infty} \mathbf{C}_{\tilde{\mathbf{n}}} = 2\mathbf{C}_{\mathbf{n}} . \qquad (5.10)$$

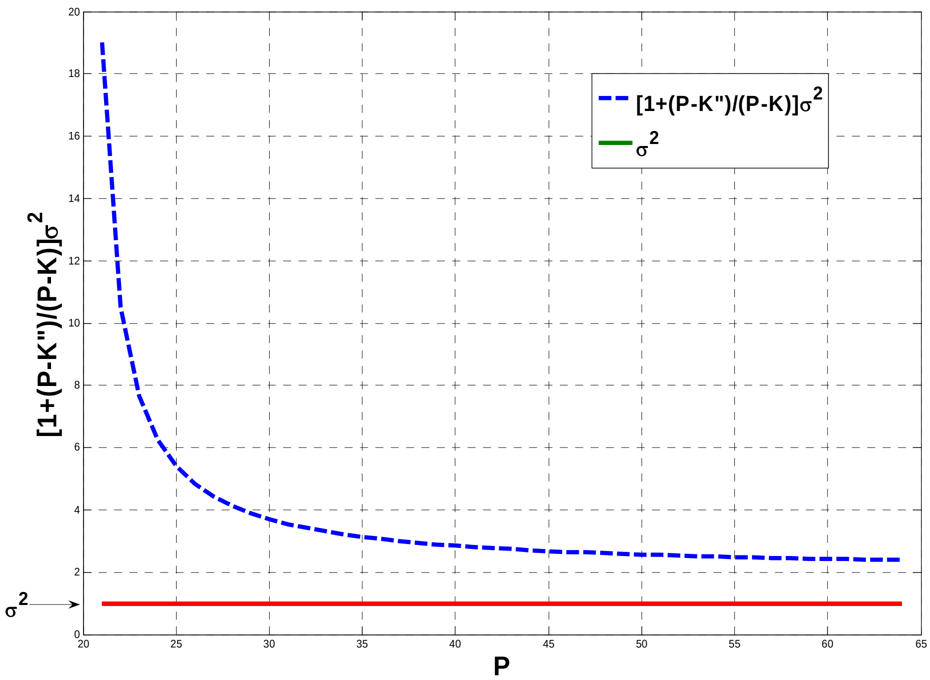

Figure 12: The relationship between $\tilde{\sigma}^2 = \left(1 + \frac{P-K''}{P-K}\right)\sigma^2$ and σ^2, where σ^2 denotes normalized noise power.

This relationship in (5.10) is shown in Fig. 12. The performance of the proposed semiblind multiuser detectors improve with increasing P, the column size of semiblind spreading matrix $\mathbf{S}$.

5.3. SUMMARY

The proposed semiblind multiuser detectors are analyzed and compared with conventional least-squares detector, the decorrelating detector. At first, the bit-error rate of the proposed least-square detector is derived. Then, the geometric explanation of the proposed detector is discussed. Finally, the noise enhancement of proposed detectors is illustrated.

Chapter 6

Computer Simulations

In this chapter, various computer simulation results are presented to demonstrate the performance of the proposed algorithms. The proposed algorithms are compared with the conventional decorrelating detector and also with single-user matched filter.

In the computer simulations, it is assumed that there are K'' active users in the cell. The spreading sequences are random sequences with spreading gain $L = 64$. The group size K, which means that the receiver actually knows K spreading sequences beforehand. We assumes all users in the same group have equal received power and the BER is accounted with averaging all the users in the group. Since MC-CDMA essentially is a synchronous CDMA system, the multiuser receiver knows the timing of received signals, the desired spreading sequences and the amplitudes of the desired users. In order to demonstrate the performance of the proposed MC-CDMA multiuser detection schemes for fading channel, we use nonorthogonal random spreading sequences instead of orthogonal spreading sequence such as Walsh sequences.

6.1. BER Performance versus SNR

We exam the BER performances of the proposed least-squares, total least-squares and mixed TLS/LS semiblind algorithms by changing the power of the additive background white channel noise. The results are presented in Figs. 13 to 18. The data $\overline{\mathbf{D}}$ associated with the semiblind matrix $\mathbf{S}$ is assume to be perfectly known in Figs. 13 to 15 and is estimated in Fig. 16 to 18. The results show that the performance of single-user matched filter and decorrelating detector are best when near-far ratio is very low. In those scenarios, the performance of least-square semiblind detector is close to the decorrelating detector and the single-user matched filter especially when SNR is greater than 15dB. With increasing near-far ratio, the single-user matched filter obviously experiences the near-far problem while the performance of the conventional decorrelating detector basically remains unchanged. The performance of the proposed least-squares semiblind multiuser detector becomes better than that of the single-user matched filter for large near-far ratios. However, Figs. 14, 15, 17 and 18 show that the performances of the proposed total least-square and mixed TLS/LS detectors are not as good as the least-squares approach, though both of them try to estimate noise and mitigate noise effects. Since their noise estimation is not accurate, it actually brings possible bias and noise-enhancement effects in the detection of the signals.

6.2. BER Performance against Near-Far Ratio

The near-far resistant performances of the proposed detectors are examined here. In the system, we assume there are only two active users but only one user's spreading sequence is known. The spreading sequence covariance matrix $\mathbf{R}$ is given as

$$\mathbf{R} = \mathbf{C}^{\mathrm{T}}\mathbf{C} = \begin{bmatrix} 1 & \rho \\ \rho & 1 \end{bmatrix}. \tag{6.1}$$

The results are shown in Figs. 19 and 20 which show that the performances of the proposed least-squares detector are much closer to that of the decorrelating detector than the single-user matched filter does. When the near-far ratio is low, the performance of the proposed least-square detector, single-user matched filter and decorrelating detector are close to each other. When the near-far ratio is high, the performance of single-user matched filter worsens while the performances of the proposed least-squares detector and decorrelating detector remain close to each other and do not change much. Totally, the BERs of the least-square, total least-square and mixed LS/TLS semiblind detectors are pretty flat, compared with the single-user matched filter. However, the performances of the proposed TLS and MLS detectors are still poor. Their BERs are higher than those of the least-square semiblind detector nearly all the time.

6.3. BER Performance against Amplitude Errors

Here we exam the performance of the proposed semiblind detectors by intentionally adding some noise/errors on the known amplitude to see how their resistance to this kind of error is. The performances of the proposed semiblind detectors with changing amplitude errors and the background white Gaussian channel noise are presented in Fig. 21, 22 and 23. Fig. 21 shows that the performance of the proposed least-square semiblind multiuser detector does not change much with increasing amplitude errors when the SNR is greater than 5 or 6 dB. It is similar to the property of the conventional decorrelating detector, the performance of which is independent of the amplitudes of received signals. This is an important property. This means that if we develop the proposed semiblind algorithms to be blind algorithms with adding additional amplitude estimation, it is possible that the performance of this kind of blind algorithms will not be heavily dependent on amplitude estimators when SNR is large. However, the performances of the proposed TLS and MLS semiblind detectors are not satisfying: their BERs are not stable as shown in Figs. 22 and 23.

6.4. BER Performance with Changing *P*

The performances of the proposed semiblind detectors are examined with changing the size of the semiblind spreading matrix $\mathbf{S}$. Though the theoretical analysis of the new noise vector $\tilde{\mathbf{n}}$ shows that the new noise covariance value becomes smaller with increasing P which is the size of semiblind signature matrix, the actual performance of the proposed semiblind detectors do not become better with a larger P. This is shown in Figs. 24, 25 and 26 When observing the simulation results, we find that the condition number of $\mathbf{S}^{\mathrm{T}}\mathbf{S}$ becomes larger when P is increased.

This maybe is one of the reasons that make the actually results different from the theoretical analysis.

6.5. SUMMARY

The computer simulations of the proposed semiblind multiuser detectors are presented and compared with conventional detectors. At first, we examine the BER of the proposed detectors by changing the SNR. It is surprising to see that the proposed least-square detector outperforms other detectors, even it does not take advantage any noise estimation. This may be because the noise estimation in other detectors is not accurate enough. Then we can check the near-far resistance of the proposed detectors. It shows that the near-far resistance of the proposed least-squares detector is better than that of the single-user matched filter but not as good as the conventional decorrelating detector when the interference is strong. After this, we check the performance of the proposed multiuser detectors by adding noise on the known amplitude to illustrate estimation effects. Finally, we exam their performance by changing the group size.

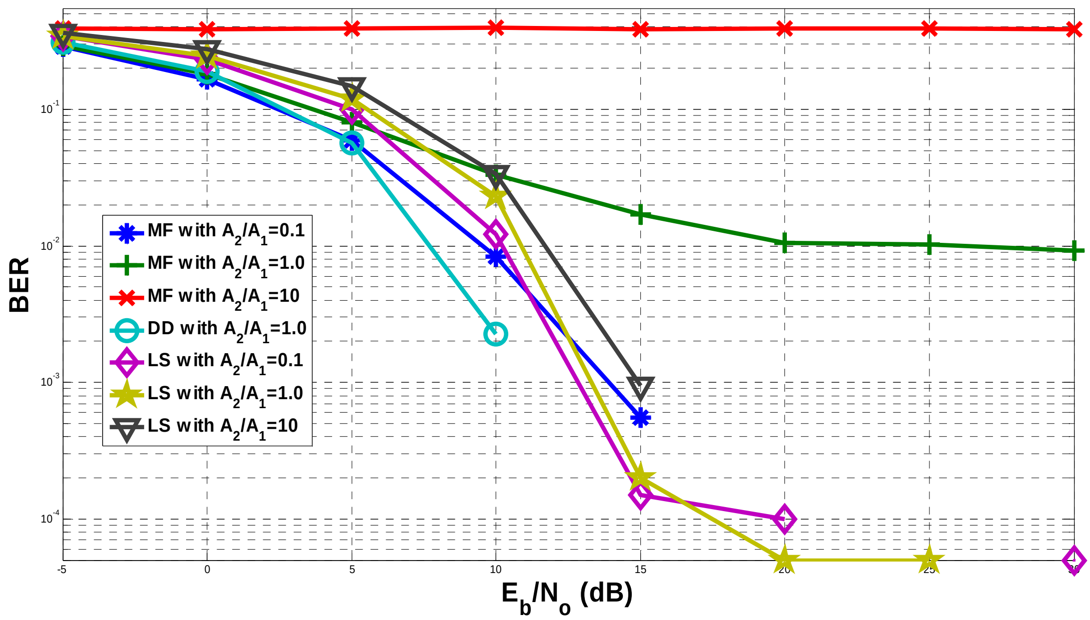

Figure 13. The BER performance comparison of conventional detectors and the proposed least-square detector with known $\overline{\mathbf{D}}$. K =8, K'' =16, P = 32, L =64.

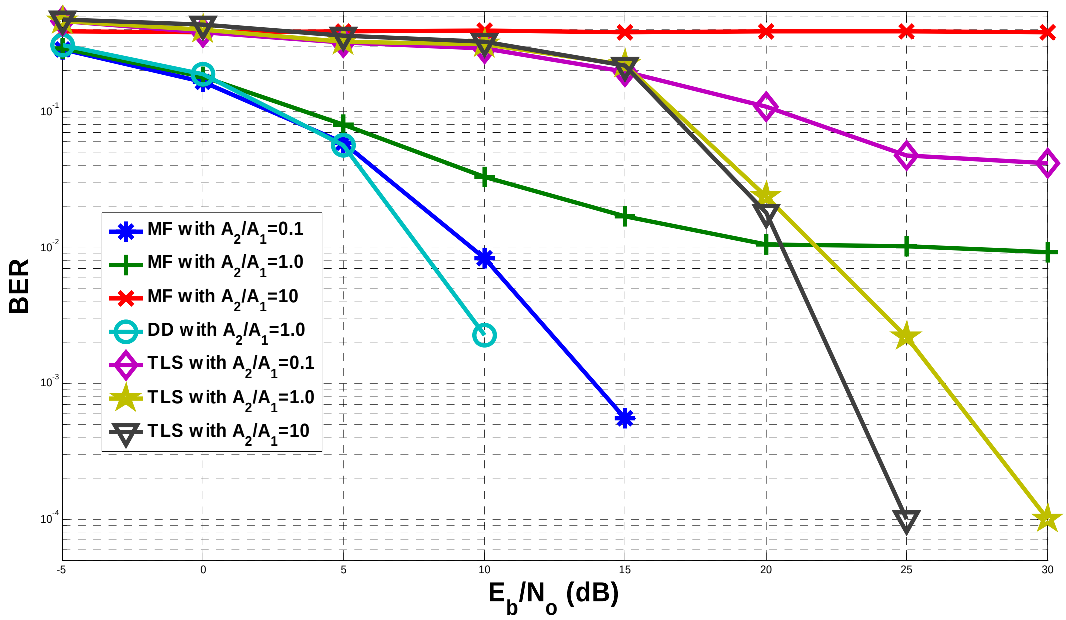

Figure 14. The BER performance comparison of conventional detectors and the proposed total least-square detector with known $\overline{\mathbf{D}}$. K =8, K'' =16, P = 32, L =64.

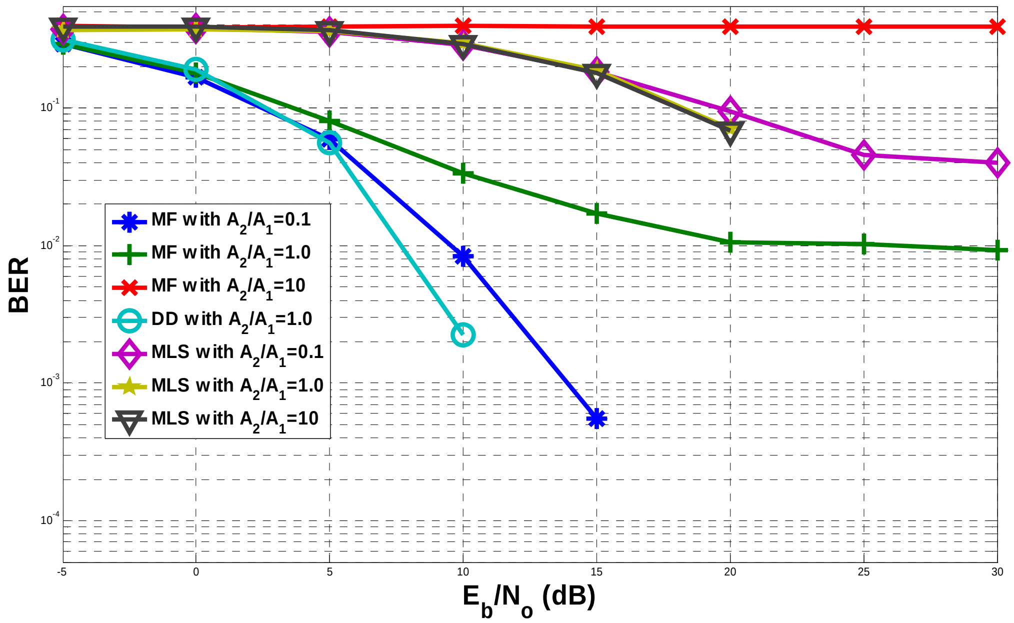

Figure 15. The BER performance comparison of conventional detectors and the proposed mixed least-square detector with known $\overline{\mathbf{D}}$. K =8, K'' =16, P = 32, L =64.

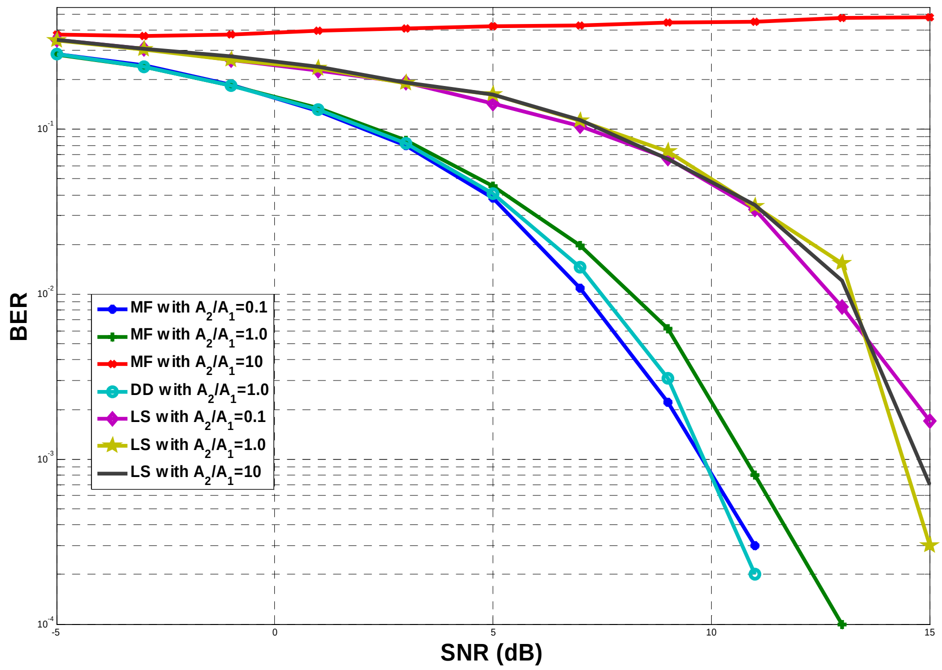

Figure 16. The BER performance comparison of conventional detectors and the proposed least-square detector with estimated $\overline{\mathbf{D}}$. K =8, K'' =16, P = 32, L =64.

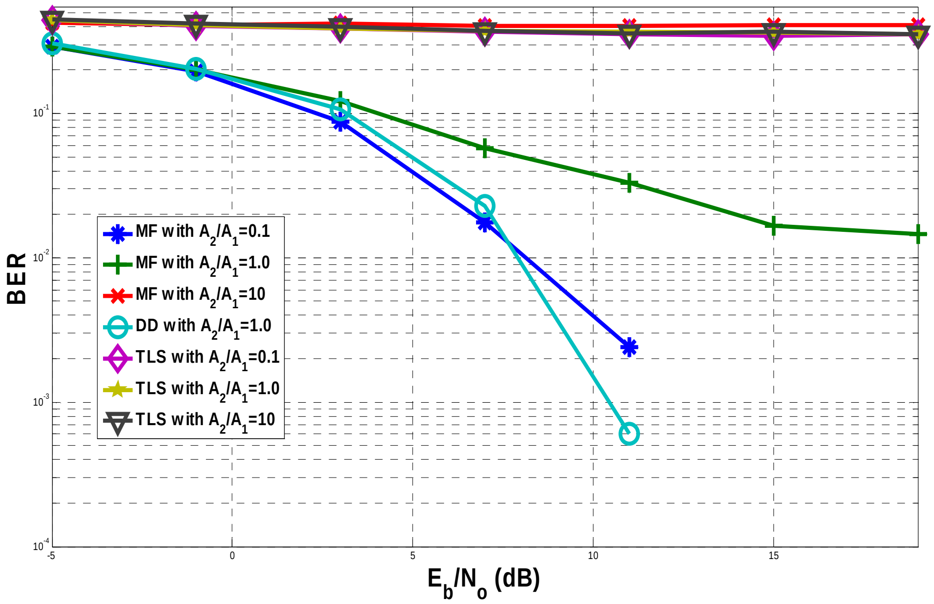

Figure 17. The BER performance comparison of conventional detectors and the proposed total least-square detector with estimated $\overline{\mathbf{D}}$. K =8, K'' =16, P = 32, L =64.

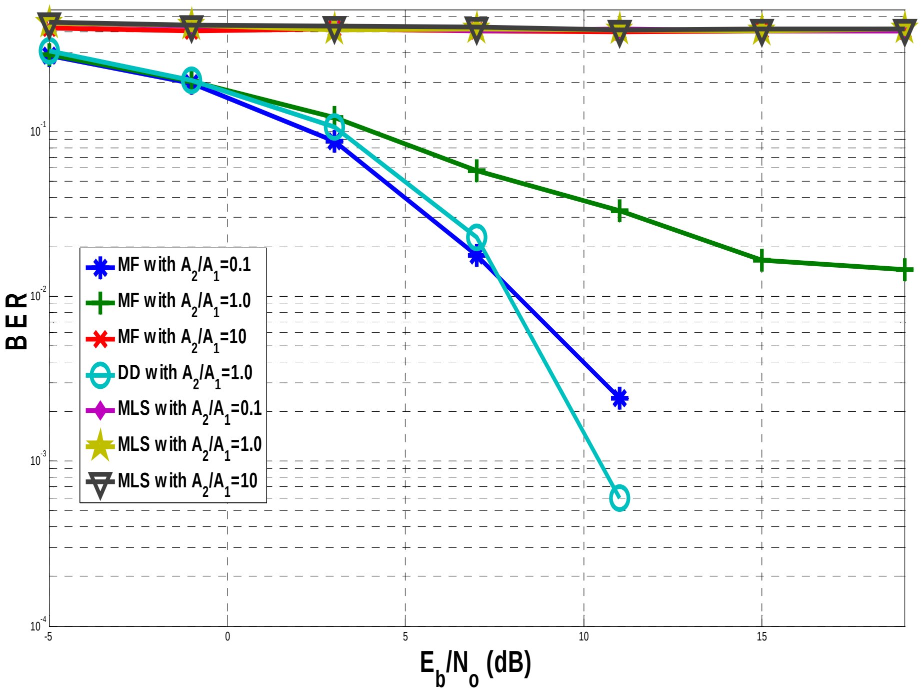

Figure 18. The BER performance comparison of conventional detectors and the proposed mixed least-square detector with estimated $\overline{\mathbf{D}}$. K =8, K'' =16, P = 32, L =64.

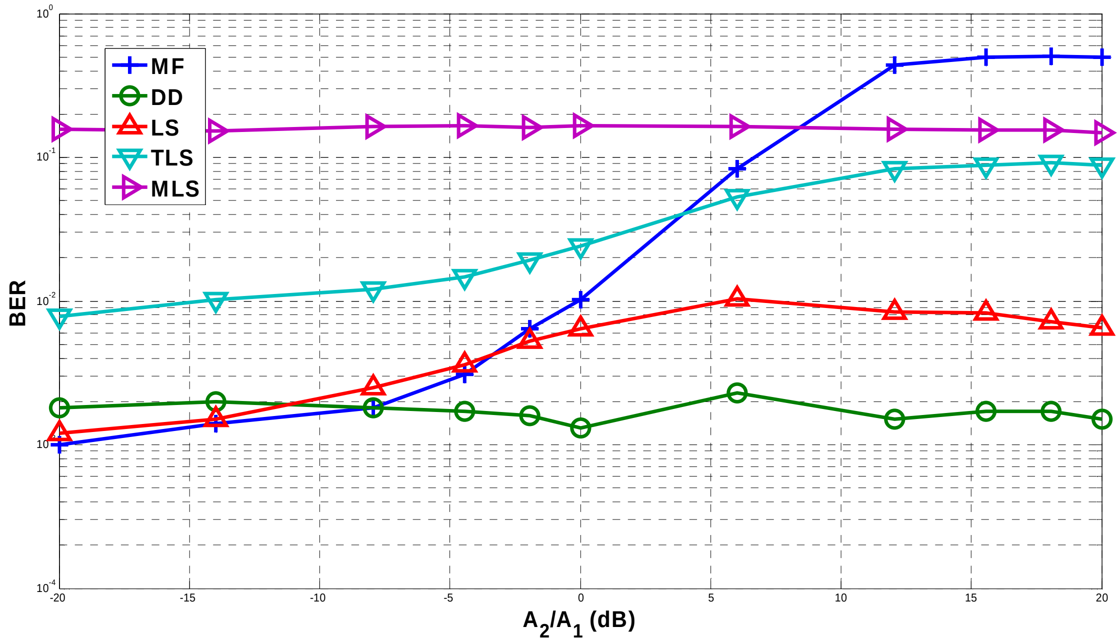

Figure 19. The NFR comparison of conventional detectors and the proposed detectors with known $\overline{\mathbf{D}}$. $K=1$, $K''=2$, $P=4$, $L=64$, $\rho=0.344$. SNR=10dB.

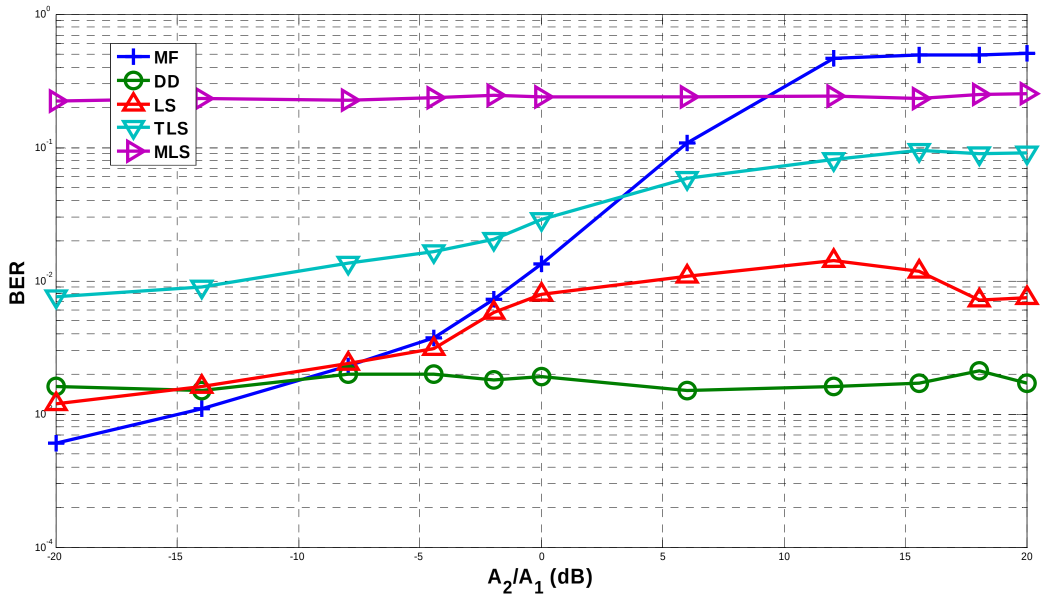

Figure 20. The NFR comparison of conventional detectors and the proposed detectors with estimated $\overline{\mathbf{D}}$. $K=1$, $K''=2$, $P=4$, $L=64$, $\rho=0.375$. SNR=10dB.

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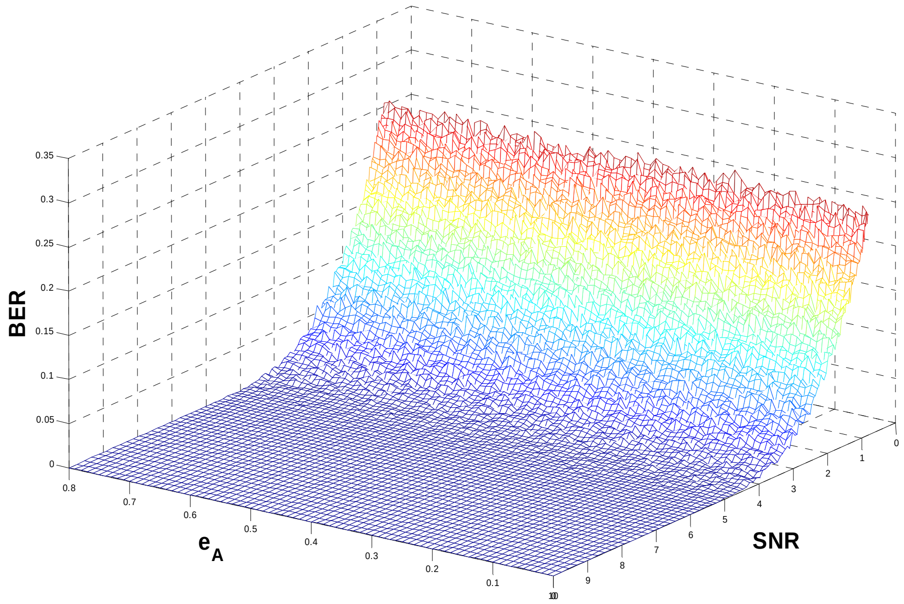

Figure 21. The BER performance of the proposed least-square semiblind detector with amplitude errors. $e_{\mathbf{A}} = \left\| \Delta \mathbf{A}_1 / \mathbf{A}_1 \right\|$.

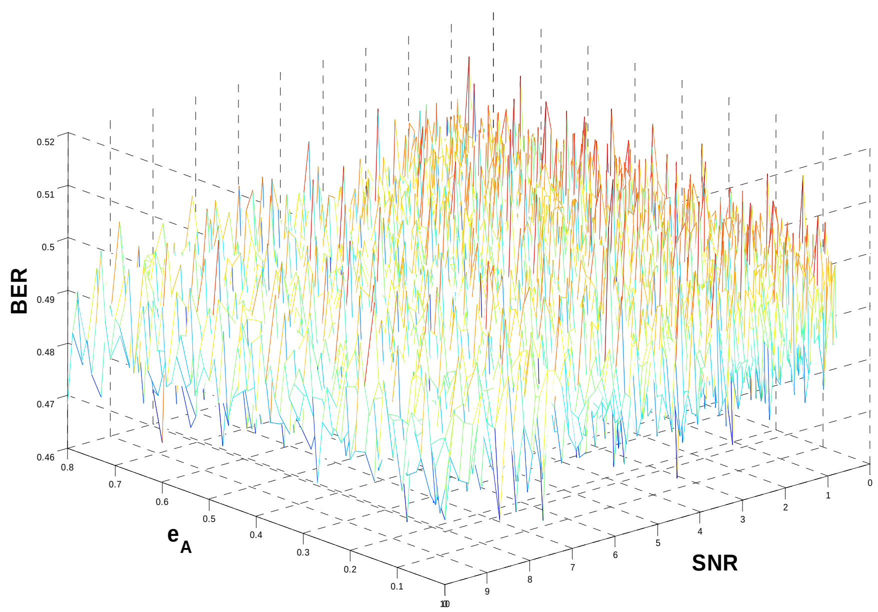

Figure 22. The BER performance of the proposed total least-square semiblind detector with amplitude errors. $e_{\mathbf{A}} = \left\| \Delta \mathbf{A}_1 / \mathbf{A}_1 \right\|$.

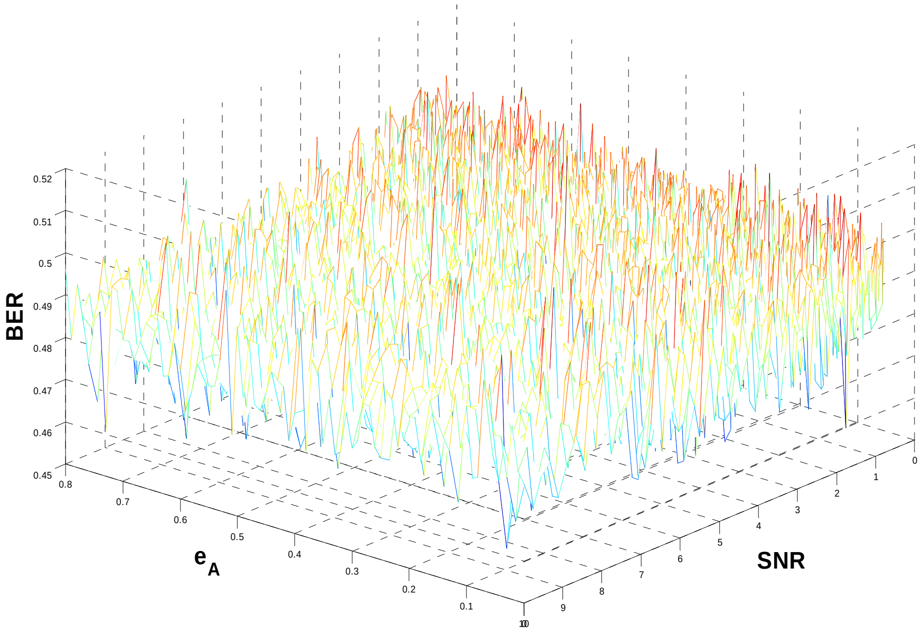

Figure 23. The BER performance of the proposed mixed TLS/LS semiblind detector with amplitude errors. $e_{\mathbf{A}} = \left\| \Delta\mathbf{A}_1 / \mathbf{A}_1 \right\|$.

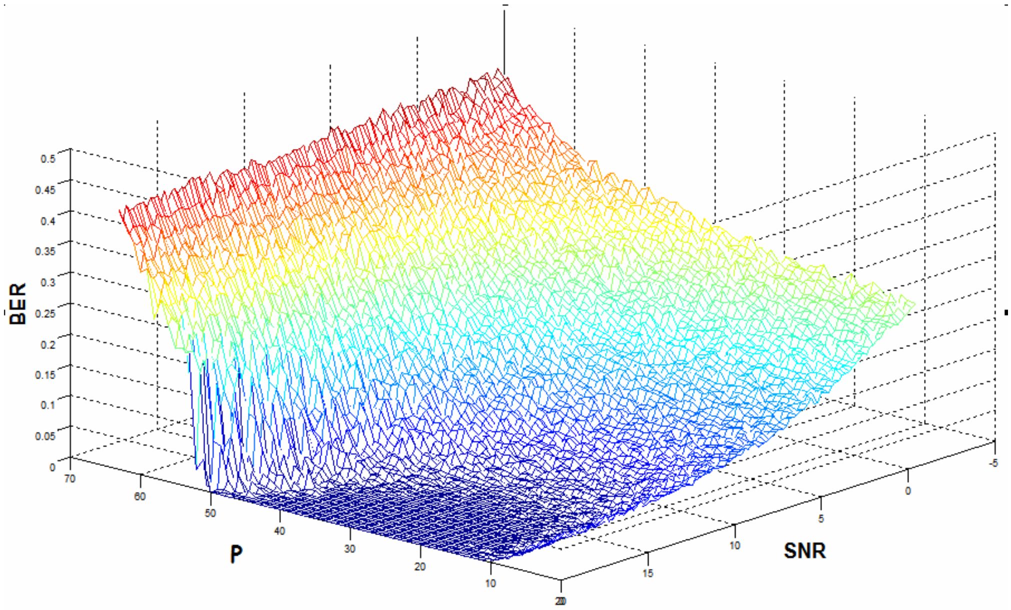

Figure 24. The performance of the proposed least-square semiblind detector with changing *P*.

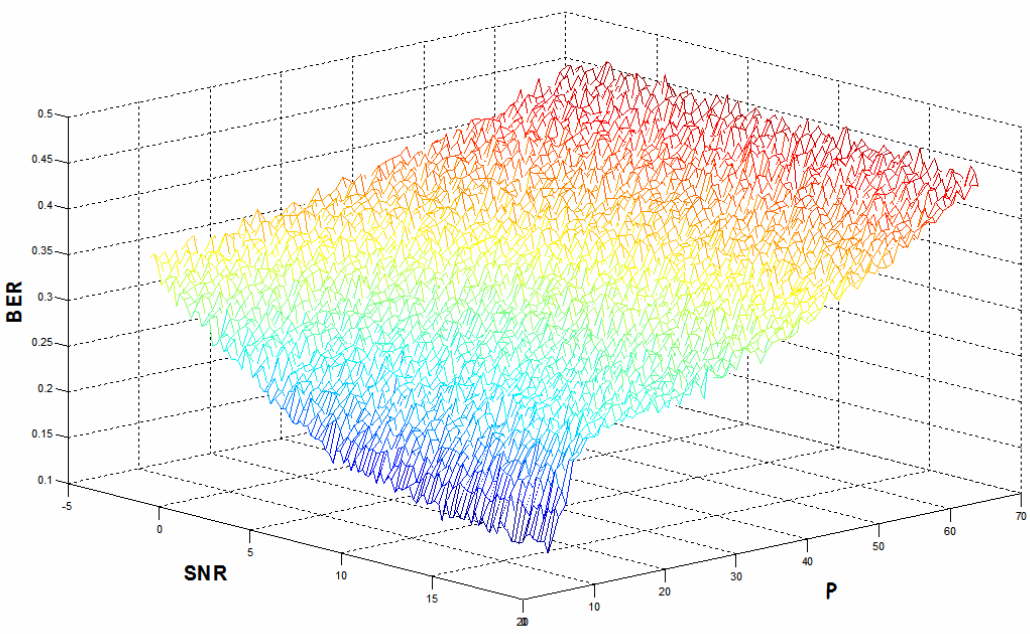

Figure 25. The BER performance of the proposed total least-square semiblind detector with changing *P*.

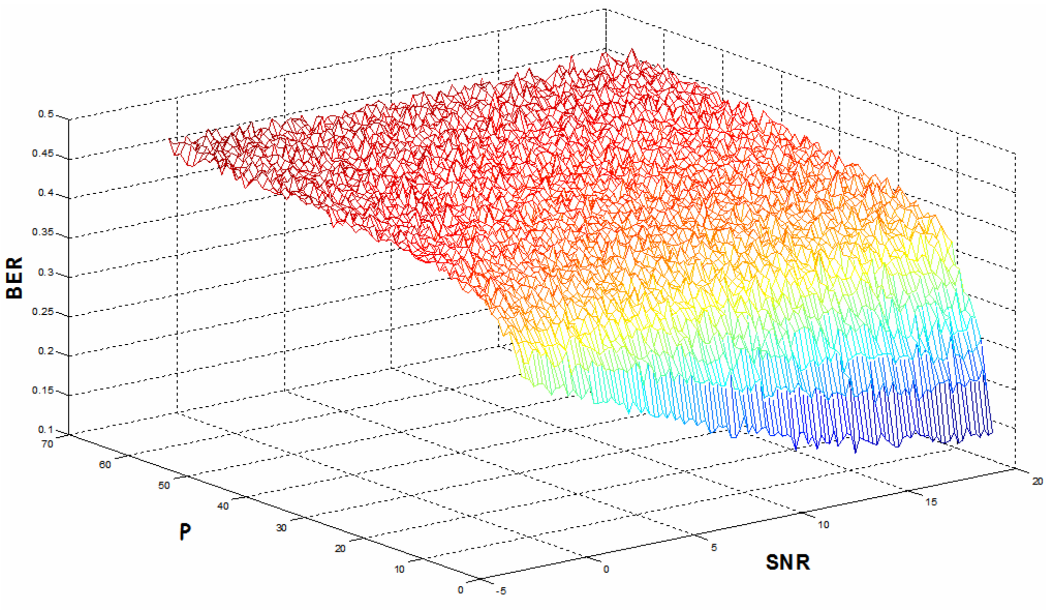

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Figure 26. The BER performance of the proposed mixed TLS/LS semiblind detector with changing *P*.

Chapter 7

Conclusions and Future Work

In this thesis, we proposed a group-based semiblind multiuser detection framework for MC-CDMA system, which essentially is a synchronous CDMA system. Instead of detecting the data for only one desired user or channel, the information data for multiple users and channels are detected at the same time. Compared to other semiblind multiuser detection schemes, a semiblind spreading matrix and a new detection model are proposed. Based on this detection model, the least-square multiuser detection approaches, including least squares estimation, total least squares estimation and mixed LS/TLS estimation, are proposed. In these approaches, besides the spreading sequences and timing of desired users, the amplitudes of desired users are also required so that they are all called semiblind detection. Compared with other blind or semiblind multiuser detection algorithms, the proposed ones are simple and direct without additional estimation procedure for discovering other unknown users' interference structure. Computer simulation and theoretical analyses are presented to demonstrate their performance too.

In future, there are following possible topics to continue this research work.

1) So far, the amplitudes of desired users need be known beforehand. In the future, it may be interesting to investigate a simple amplitude estimation scheme which can be easily incorporated with the proposed semiblind approaches.

2) The proposed algorithms are not simple enough yet for practical applications, since they all involve the matrix inverse operation of the proposed semiblind spreading matrix. Therefore it is necessary to look for a good matrix computation scheme to simplify the multiuser detection structure.

3) In practical situations, it is possible that even desired users' spreading sequence is hard to be known beforehand. Therefore it is desired to investigation some spreading sequence estimation scheme with the proposed approaches.

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