Turbo Receiver for Spread Spectrum Systems Employing Parity Bit Selected Spreading Sequences

by

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To my parents,

and my love, Aphra.

Abstract

In spread spectrum systems employing parity bit selected spreading sequences, parity bits generated from a linear block encoder are used to select a spreading code from a set of mutually orthogonal spreading sequences. In this thesis, turbo receivers for SS-PB systems are proposed and investigated. In the transmitter, data bits are first convolutionally encoded before being fed into SS-PB modulator. In fact, the parity bit spreading code selection technique acts as an inner encoder in this system without allocating any transmit energy to the additional redundancy provided by this technique.

The receiver implements a turbo processing by iteratively exchanging the soft information on coded bits between a SISO detector and a SISO decoder. In this system, detection is performed by incorporating the extrinsic information provided by the decoder in the last iteration into the received signal to calculate the likelihood of each detected bit in terms of LLR which is used as the input for a SISO decoder.

In addition, SISO detectors are proposed for MC-CDMA and MIMO-CDMA systems that employ parity bit selected and permutation spreading. In the case of multiuser scenario, a turbo SISO multiuser detector is introduced for SS-PB systems for both synchronous and asynchronous channels. In such systems, MAI is estimated from the extrinsic information provided by the SISO channel decoder in the previous iteration. SISO multiuser detectors are also proposed for the case of multiple users in MC-CDMA and MIMO-CDMA systems when parity bit selected and permutation spreading are used.

Simulations performed for all the proposed turbo receivers show a significant reduction in BER in AWGN and fading channels over multiple iterations.

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List of Acronyms

3G Third Generation

3GPP Third Generation Partnership Project

3GPP2 Third Generation Partnership Project 2

AWGN Additive White Gaussian Noise

BER Bit Error Rate

BPF Band Pass Filter

BPSK Binary Phase Shift Keying

CDMA Code Division Multiple Access

CDMA-PB CDMA systems employing parity bit selected spreading codes

DS-CDMA Direct Sequence CDMA

E-MDFH Enhanced MDFH

EV-DO Evolution Data Optimized

FFH Fast Frequency Hopped

FH-CDMA Frequency Hopping CDMA

HSPA High Speed Packet Access

Hz Hertz

Kbps Kilo bit per second

kHz Kilohertz

LDP Low Probability of Detection

LLR Log Likelihood Ratio

LTE Long Term Evolution

MAI Multiple Access Interference

Mbps Mega bit per second

MC-DS-SS Multi Carrier DS-SS

MC-CDMA Multi Carrier CDMA

MC-SS-PB Multicarrier SS-PB

MDFH Message Driven Frequency Hopping

MHz Megahertz

MIMO Multi Input Multi Output

MLD Maximum Likelihood Detection

MUD Multi User Detection

OFDMA Orthogonal Frequency Division Multiple Access

PN Pseudonoise

P-SE-SS Product SE-SS

QAM Quadrature Amplitude Modulation

SE-CDMA Self Encoded CDMA

SE-SS Self Encoded Spread Spectrum

SFH Slow Frequency Hopped

SIC Soft Interference Cancellation

SISO Soft-Input Soft-Output

SNR Signal to Noise Ratio

SS Spread Spectrum

SS-PB Spread Spectrum system employing Parity Bit selected spreading codes

TCM Trellis Coded Modulation

TDMA Time Division Multiple Access

UMB Universal Mobile Broadband

WCDMA Wideband CDMA

WTC Woerner's Trellis Code

List of Symbols

A_k : signal amplitude of the kth user.

$\underline{A}$: matrix of all signal amplitudes of all users.

$\alpha(t)$: fading gain of the channel.

α_v : channel gain of the vth subcarrier.

α_{ij}^l : complex channel gain on the link between transmit antenna i and receive antenna j on lth signaling interval.

$\{b[i]\}$: stream of modulated bits.

$\underline{b}^l$: vector of modulated bits on the lth signaling interval.

$\mathcal{B}_+^k$: set of all modulated bit vectors whose kth bit is $+1$.

$\mathcal{B}_-^k$: set of all modulated bit vectors whose kth bit is -1

$\{c_q(t)\}$: Set of mutually orthogonal spreading sequences.

$c_j(t)$: jth spreading waveform.

$\{d[i]\}$: stream of data bits.

$\underline{d}^l$: vector of data bits on the lth signaling interval.

E_b : bit energy.

f_c : carrier frequency.

f_v : carrier frequency of the vth subcarrier.

f_d : Doppler spread.

$\underline{G}$: generator matrix of a linear code.

$\underline{H}^{(k)}$: channel gain matrix of the kth user.

$\underline{I}$: Identity matrix.

K : number of users in a CDMA system.

k_0 : length of the message word in a systematic block code.

$\Lambda_c(b_k)$: a posteriori LLR of b_k provided by the SISO channel decoder.

$\lambda_c(b_k)$: extrinsic information about b_k provided by the SISO channel decoder.

$\lambda_c^p(b_k)$: a posteriori LLR of b_k calculated by the SISO channel decoder in the previous iteration.

$\Lambda_c(d_k)$: a posteriori LLR of data bit d_k provided by the SISO channel decoder.

$\Lambda_d(b_k)$: a posteriori LLR of b_k provided by the SISO detector.

$\lambda_d(b_k)$: extrinsic information about b_k provided by the SISO detector.

$\lambda_d^p(b_k)$: a posteriori LLR of b_k calculated by the SISO detector in the previous iteration.

$\{m[i]\}$: stream of coded bits.

$\underline{m}^l$: vector of coded bits on the lth signaling interval.

M_i : ith coset.

$n(t)$ additive white Gaussian noise.

n_0 : length of the code word in a systematic block code.

N_0 : noise spectral density.

N_c : number of subcarriers.

N_t : number of transmit antennas.

N_r : number of receive antennas.

$\underline{p}$: parity vector.

$\underline{P}$: parity matrix.

P_b : probability of a bit error.

$p_{T_c}(.)$: takes value 1 on the interval $[0, T_c)$ and 0 otherwise.

Q : number of mutually orthogonal spreading sequences assigned to each user.

$r(t)$: received signal.

R : code rate.

$\underline{R}$: matrix of cross-correlations.

$\underline{\rho}^{(p,q)}$: matrix of cross-correlation between the spreading sequences of user p and q.

$\rho_{i,j}^{(p,q)}$: cross-correlation between the ith spreading code of user p and the jth spreading code of user q.

$s(t)$: transmitted signal.

σ_n : standard deviation of noise.

T_b : bit duration.

τ_k : random delay corresponding to the kth use.

$U_{b,c}^{(i)}$: decision variable of the cth spreading sequence in the bth subcarrier.

$U_i^{(j)}$: ith matched filter output on $j$the signaling interval.

$U_{b,c}^{(i)}$: decision variable of the cth spreading sequence in the bth subcarrier.

$\mathcal{W}$: set of the spreading sequences used by the transmit antennas.

$w_i(t)$: spreading waveform used by the ith transmit antenna.

$\underline{y}_{(\underline{b})}^l$: noise free matched filter outputs when $\underline{b}^l$ is transmitted.

$\underline{z}_q^l$: vector of the qth matched filter outputs on the l signaling interval.

$\underline{Z}$: matrix of all matched filter outputs.

Chapter 1

Introduction

1.1 Motivation

Presently, the world is experiencing a tremendous growth in the number of mobile communication devices as well as the amount of traffic carried on mobile networks. With the introduction of smartphones like iPhone, Blackberry and Android based devices, the emerging market of new tablet computers like iPad and other devices capable of connecting to mobile networks, it has never been this easy to access the Internet on the go and provide data communication on mobile devices. This revolution in mobile devices has led to a huge explosion in data traffic over mobile networks. In March 2010, Ericsson announced that mobile data surpassed voice on a global basis during December 2009 [1]. Cisco has predicted that mobile data traffic would increase by a factor of 39 times between 2009 and 2014 [2].

Code division multiple access (CDMA) has been the main technique used in the physical layer of the third generation of mobile communication systems (3G). As of August 2011, there are more than 1.3 billion mobile subscribers around the world that use CDMA technology in one way or the other. This statistic includes 720 million subscribers of the third generation partnership project (3GPP) family of standards (WCDMA/HSPA) [3] and 580 million subscribers of the third generation partnership project 2 (3GPP2) family of standards (CDMA2000/EV-DO) [4].This number will see a steady growth as more and more operators around the world opt to deploy one of these 3G networks which are all based on CDMA technology. This figure shows that making any kind of progress in CDMA technology will benefit many operators around the world as they can keep the existing equipment with minimum turn over cost while providing better quality of service and higher data rate to their costumers.

CDMA technology is based on direct-sequence spread spectrum (DS-SS) systems in which every user employs a unique spreading sequence in order to transmit its signal on a bandwidth considerably larger that the bandwidth of the original signal. Multiple users access the same channel and are separated at the receiver by the spreading sequences employed. Typically, the spreading sequences are not uncorrelated and this leads to multiple access interference (MAI) which causes a degradation in the bit error rate (BER) performance of all users. The maximum amount of users that can access a CDMA channel is attained once the BER performance of all users reaches a critical rate.

A spread spectrum system that employs parity bit selected spreading sequence (SS-PB) is first introduced in [5] in which the parity bits generated by a linear block encoder are used to select a spreading code from a set of orthogonal spreading sequences. Combining coding with spreading code selection provides a coding gain in the system's BER performance while keeping the same spectral efficiency. The application of the parity bit selected spreading technique is extended to multicarrier CDMA (MC-CDMA) as well as multi-input multi-output (MIMO) CDMA systems in [6] and [7] respectively. To obtain all of the benefits of SS-PB systems, an effective receiver is needed that is able to exploit the additional information that is conveyed by this hybrid of coding and spreading.

Turbo (iterative) processing techniques have been applied in a variety of ways to improve the performance of several communication systems. In this thesis, we study the application of turbo processing in the receiver of different SS-PB systems. In the transmitter, data bits are first convolutionally encoded before they are input to the SS-PB modulators. In such systems, the parity bit selected spreading technique acts as an inner encoder without requiring any additional transmit energy for the provided redundancy. In the receiver, the soft information provided by the soft-input soft-output (SISO) detector and decoder is iteratively exchanged. Designing of the SISO detector for different SS-PB systems is the main objective of this thesis.

1.2 Contribution of the Thesis

The main contributions of this thesis can be summarized as follows:

- The introduction of turbo processing in the structure of the receivers of SS-PB systems.
- Design of a SISO detector for a single user SS-PB system that provides the reliability of each detected bit in terms of log likelihood ratio (LLR).
- Proposal of turbo receivers for MC-CDMA and MIMO-CDMA systems employing parity bit selected spreading or, in the case of MIMO-CDMA, permutation spreading.
- Investigation of multiuser scenario of SS-PB systems and introduction of turbo multiuser receivers for such systems.
- Design of SISO multiuser detector for both synchronous and asynchronous CDMA systems that employ SS-PB techniques.

- Proposals of multiuser receivers for MC-CDMA and MIMIO-CDMA systems that employ parity bit selected and permutation spreading.

1.3 Organization of the Thesis

The remainder of this thesis is organized as follows:

- **Chapter 2** briefly presents an overview on basic CDMA systems. In this chapter, several CDMA systems in which the spreading codes are dependent on the transmitted data are explained. This chapter also includes an explanation of SS-PB systems.
- **Chapter 3** introduces the application of turbo processing in the receiver of SS-PB systems. This chapter proposes a SISO detector for SS-PB systems that calculates the LLRs of each detected bit.
- **Chapter 4** investigates the application of turbo processing in the receiver of MC-CDMA systems that employ parity bit selected spreading sequence.
- **Chapter 5** proposes a turbo receiver for MIMO-CDMA systems that employ parity bit selected and permutation spreading.
- **Chapter 6** studies the multiuser scenario of SS-PB systems. In this chapter, turbo multiuser receivers for both synchronous and asynchronous CDMA systems that employ parity bit selected spreading sequence are proposed. Turbo multiuser detection for MC-CDMA and MIMO-CDMA systems with parity bit selected and permutation spreading are also investigated
- **Chapter 7** concludes the thesis, provides the summary of the work and suggests potential further research to continue this work.

Chapter 2

Background and Literature Survey

2.1 Introduction

Spread spectrum (SS) is a technique in which a narrowband signal is intentionally spread over a much higher bandwidth. Spread spectrum systems were initially adopted for military applications as they are very robust against interception and jamming [8].

Due to its anti jam capabilities, spread spectrum can be used as a method of multiple access. Doing so can provide huge advantages in commercial applications. In code division multiple access (CDMA) systems, multiple access is provided by assigning each user a different spreading code [9]. There are two prominent CDMA systems: namely, direct-sequence CDMA (DS-CDMA) and frequency-hopping CDMA (FH-CDMA).

2.1.1 DS-CDMA Systems

DS-CDMA is the most popular technique in CDMA systems which has been employed in many cellular networks and telecommunication systems around the world. The basic concept behind DS-CDMA system is to multiply a high rate unique pseudonoise (PN) sequence, $c_k(t)$, by the modulated data transmitted by user k. PN sequence consists of a number of code bits called chips. Since the PN sequence has much higher chip rate than data bit rate, the resulting signal occupies much larger bandwidth than the original signal. The ratio of the PN sequence chip rate to the original data bit rate is called *processing gain*. Although all users occupy the same bandwidth, information of each user can be recovered without additional interference if mutually orthogonal PN sequences are employed by the users. The basic DS-CDMA system model is shown in Figure 2.1 [9] [10] .

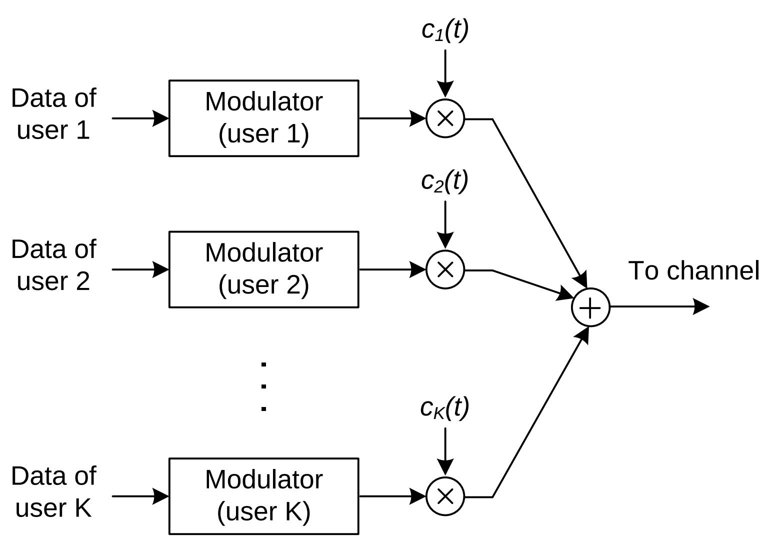

Figure 2.1: K-user DS-CDMA system.

Received signal in a K-user DS-CDMA system in an additive white Gaussian noise (AWGN) channel can be modeled as [9]:

$$r(t) = \sum_{k=1}^{K} A_k s_k(t - \tau_k) c_k(t - \tau_k) + n(t) \tag{2.1}$$

where A_k, $s_k(t)$, $c_k(t)$ are respectively the received amplitude, modulated signal and the spreading code of the kth user and $n(t)$ is an AWGN process with power spectral density of $N_o/2$. In equation (2.1), τ_k represents the random delay corresponding to the kth user and T_b is the duration of one bit.

The CDMA system is called synchronous if all the PN sequences of all users are aligned at the receiver (*i.e.*,$\tau_k = 0$ for all k), but if no attempts are made to align PN sequences of different users (*i.e.*,$\tau_k \neq 0$), the system is called asynchronous DS-CDMA. The asynchronous assumption is more realistic particularly for the received signal on the uplink of a CDMA system when users move and cause varying delays.

To despread the received signal, the receiver multiplies it by the locally generated PN sequence which is exactly the same as the spreading code used in the transmitter. This is accomplished by multiplying the received signal by $c_k(t)$ which in turn reproduces $s_k(t)$ since $c_k^2(t) = 1$.

The performance of DS-CDMA systems is limited by the interference caused by other users that are simultaneously transmitting signals in the same bandwidth when non-orthogonal spreading codes are used. This interference is known as MAI and it depends on the number of users in the system, cross-correlation between spreading codes used by different users, multipath effect of the channel, transmitted power level of different users and so on. Different techniques for multiuser detection (MUD) have been used to mitigate the effect of MAI and improve the performance of DS-CDMA systems [11].

2.1.2 FH-CDMA Systems

In FH-CDMA the available channel bandwidth is divided into a large number of contiguous frequency slots. In any signaling interval, each user transmits its signal in one or more available frequency slots. The distinct hopping pattern for each user is governed by a unique PN sequence in the transmitter. Figure 2.2 shows the block diagram of the

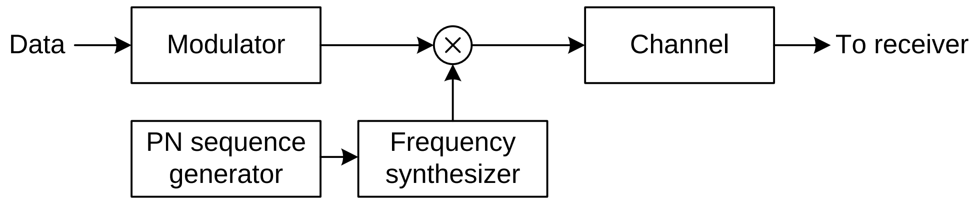

Figure 2.2: Transmitter of an FH-CDMA system.

transmitter of a FH spread spectrum system [10].

If the hopping rate of the system is less than the symbol rate of the original signal, the system is called slow frequency-hopped (SFH). Conversely, if the frequency hopping rate is faster than the symbol rate of the original signal, then it is a fast frequency-hopped (FFH) system [10]. The detection is achieved when the receiver is aware of the exact hopping pattern used by the transmitter. In FH-CDMA systems, MAI is created when collision happens on some frequency slots used by different users.

The main focus of this thesis is mainly on DS-CDMA systems.

2.2 Data Dependent Spreading Code for CDMA Systems

In this section we introduce different scenarios of spread spectrum systems in which the spreading codes are dependent on the original data. These systems include: message-driven frequency hopping system, Woerner's trellis-coded DS-CDMA system, self encoded spread spectrum systems and parity bit selected spreading sequence for DS-CDMA system. In the following subsections each of these systems are explained.

2.2.1 Message-Driven Frequency Hopping System

Message driven frequency hopping (MDFH) systems are introduced and investigated by Q. Ling et al. in [12], [13], [14] and [15]. Unlike the traditional FH systems where the hopping pattern is determined by a pre-selected PN sequence at the transmitter, in MDFH systems, part of the message itself acts as the PN sequence to determine the hopping pattern of the transmitting signal. As it can be seen in Figure 2.3, the nth block of information data is split into two parts: the first part are the bit vectors that are used to select the frequency carriers $(X_{n,1}, X_{n,2}, ..., X_{n,N_h})$ and the second part are the ordinary bit vectors that will be modulated by the selected frequency carries (Y_n).

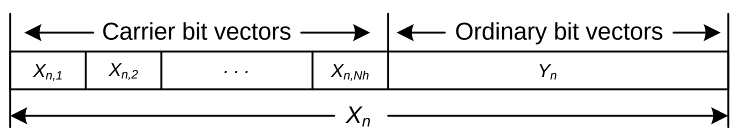

Figure 2.3: nth message vector in an MDFH system.

The transmitter block diagram of an MDFH system is shown in Figure 2.4. Each block of data is split into carrier bit vectors and ordinary bit vectors. Carrier bit vectors are then used to select carrier frequencies to modulate ordinary bit vectors.

In the receiver shown in Figure 2.5, a bank of bandpass filters (BPF) with detectors for all available carrier frequencies is used to determine which frequency is used in each hopping time. By finding the hopping pattern of the received signal, the carrier bit vectors can be easily determined and also the rest of ordinary bit vectors can be demodulated by knowing the exact frequency hopping pattern used at the transmitter.

Since the hopping pattern in MDFH systems also carries some information bits, as a result, they achieve higher spectral efficiency in comparison to the traditional FH systems

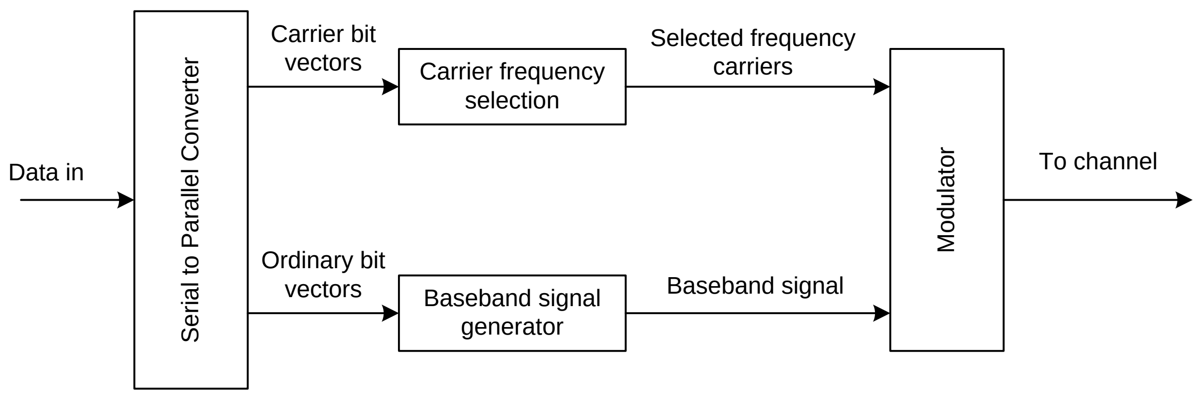

Figure 2.4: Transmitter block diagram of an MDFH system.

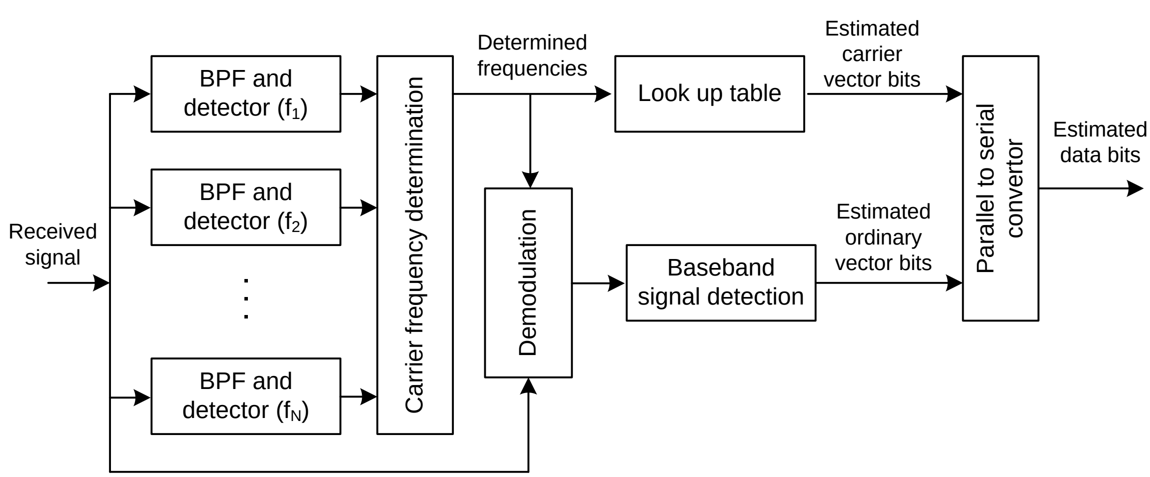

Figure 2.5: Receiver block diagram of an MDFH system.

[12]. To further improve the efficiency of these systems, the modified scheme of MDFH which is called enhanced MDFH (E-MDFH) is introduced in [15]. E-MDFH system is designed in such a way that simultaneous multiple transmissions are possible in each hopping time. If all available carrier frequencies are used in an E-MDFH system, it can be regarded as an orthogonal frequency division multiplexing (OFDM) system.

2.2.2 Woerner's Trellis-Coded DS-CDMA System

Woerner's trellis-coded (WTC) DS-CDMA system is introduced by Woerner and Stark [16] [17]. Contrary to the conventional trellis coded DS-CDMA systems where trellis coding and spread spectrum modulation are done separately over some standard two-dimensional signal constellations, in WTC-DS-CDMA systems, a trellis coding is used to select a combination of certain signature sequences that have large minimum distance.

Like trellis coded modulation (TCM), in the first stage, redundancy is added by means of convolutional code and then the output of the encoder is mapped to different PN sequence in such a way to increase the minimum distance between codes and hence increase the coding gain.

Let's assume $\{\boldsymbol{c}_i\}$ is a set of $Q/2$ orthogonal PN sequences:

$$\boldsymbol{c}_i \perp \boldsymbol{c}_j, \qquad i \neq j, i = 1, ..., Q/2. \tag{2.2}$$

This set can be further expanded by a factor of two if the antipodal of each PN sequence is also added to this set, in other words:

$$\boldsymbol{c}_{i+Q/2} = -\boldsymbol{c}_i, \qquad i = 1, ..., Q/2. \tag{2.3}$$

The new set of Q biorthogonal PN sequences is constructed over $Q/2$ dimensions and it can be used to transmit $\log_2(Q)$ bits per signaling interval.

In the first stage of the transmitter, there is a convolutional encoder of rate $\frac{\log_2(Q)-1}{\log_2(Q)}$ and in the second stage, the mapping is performed between the encoder output of $\log_2(Q)$ bits and Q PN sequences.

Using Ungerboeck's idea of set partitioning for TCM [18], Woerner and Stark have constructed trellis codes for biorthogonal PN sequence sets with 4, 8 and 16 PN sequences.

In a multiple access scenario, in order to implement a WTC-DS-CDMA system, each user must be given a separate set of biorthogonal signature sequences. In [17] some methods are introduced to construct a set of biorthogonal PN sequences for each user.

The simulation results shown in [17] confirm that the WTC-DS-CDMA system is able to accommodate a greater number of multiple access users for any given BER in comparison to the conventional DS-CDMA systems with convolutional codes. The performance improvement of such systems is a result of better distance properties of the codes and also cross-correlation properties of the signature sequences [17]. The performance of WT-DS-CDMA systems for fading channels are discussed in [19] and [20]. A multiuser receiver for WTC-DS-CDMA systems in asynchronous channels is proposed in [21]. It is shown that the proposed system is near-far resistant and has some coding gain over uncoded systems [21].

2.2.3 Self Encoded Spread Spectrum Systems

Self encoded spread spectrum (SE-SS) systems are first introduced by L. Nguyen in [22]. Unlike conventional spread spectrum systems with fixed pre-selected spreading sequences, in SE-SS systems, PN sequences are generated from the past information bits. At the transmitter there is an N-tap delay-register that provides a serial delay of past N information bits, where N is referred as the code length. In each signaling time, previous N information bits, form a time variant PN sequence that is used to spread the current bit. Figure 2.6 illustrates the block diagram of a direct sequence SE-SS system.

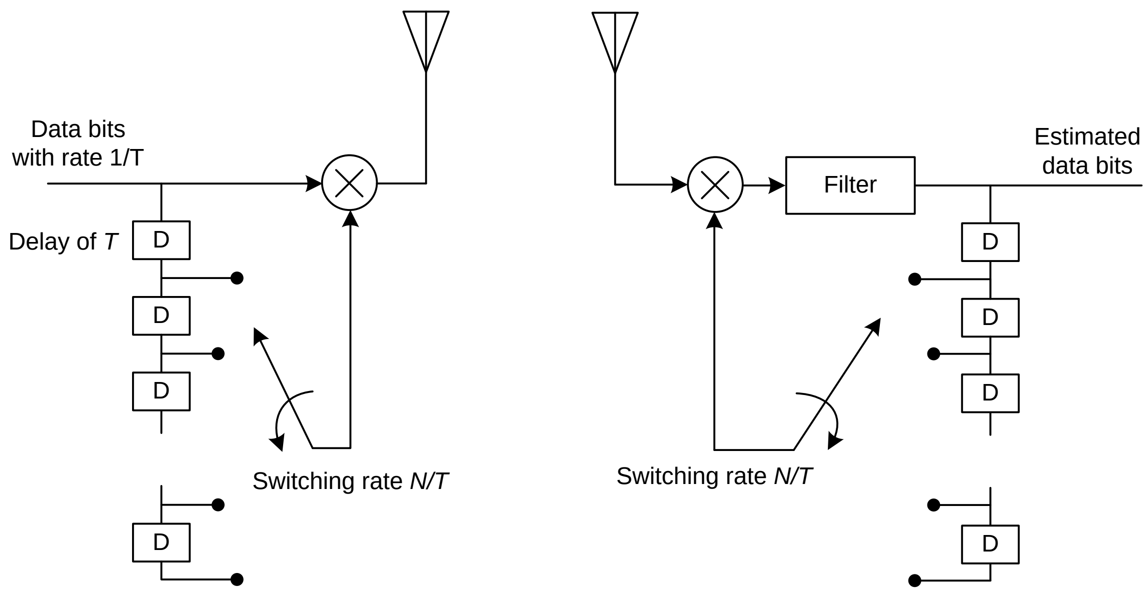

Figure 2.6: Direct sequence SE-SS system.

To assure the randomness of the spreading sequences, any redundancies of information bits can be removed by means of an appropriate data compression technique prior to be fed into the delay registers [22].

In the receiver, the reverse operation is performed. The recovered data bits are fed back into an N-tap delay-register that estimates the spreading sequence used in the transmitter. The estimated spreading sequence is then used to despread the received signal. To accomplish a successful recovery, the delay registers of both the transmitter and the receiver should be initiated with the same contents at the start of the transmission. Unless the receiver is fully aware of the delay-register structure and it is initially synchronized by the transmitter, data recovery will be extremely difficult. This property of SE-SS systems makes it more secure and more robust against unintended users. In other words, low probability of detection (LPD) is enhanced in SE-SS systems [22].

Random nature of spreading signals in SE-SS systems allow us to employ this technique in CDMA systems [23]. Based on the simulation results illustrated in [23], SE-CDMA systems have similar BER performance in AWGN channel in compare to the conventional CDMA systems using the correlation detector. The reason behind this lies in the fact that the randomness of spreading sequences in SE-CDMA systems is inherited from the random nature of the information bits.

Independent spreading codes for each user in a SE-CDMA system lead to more MAI that can be mitigated by means of appropriate channel coding [23]. Capacity analysis of m-user SE-CDMA systems in AWGN channels is presented in [24].

Synchronization in SE-SS systems is discussed in [25], [26] and [27] and multi-input multi-output (MIMO) SE-SS system is studied in [28] and [29].

Tomasin in [30] came up with similar idea as there is in SE-SS systems where the spreading signal for each bit is generated as a function of a number of past data bits. Figure 2.7 shows the block diagram of the transmitter in this system. The N-tap shift-register stores the past N data bits that are used to generate N_s chips of the spreading signal that spreads the current data bit.

Spreading sequence vector of length N_s that spreads the current bit, is generated as a function of past N data bits. Likewise the former SE-SS systems, initial synchronization and having a complete knowledge of the function used in the transmitter is crucial for the intended user to recover data bits. In contrast to the former format of SE-SS systems, more options are available to define the spreading coded as function of the past information bits. As an example of this function, product SE-SS (P-SE-SS) is considered in [30] where the chips of spreading code are generated as a product of a number of past information bits. The receiver structure discussed in [30] is in the form of successive

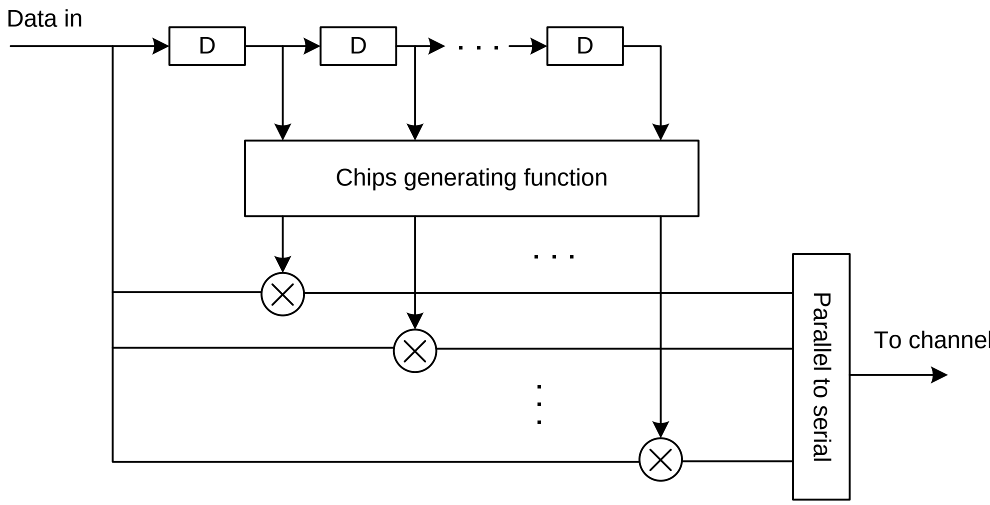

Figure 2.7: Block diagram of the SE-SS transmitter proposed in [30].

interference cancellation (SIC) where in each stage MAI is estimated and deducted from the desired signal. One of the main drawbacks of SE-SS system is error propagation that occurs due to the dependency between the data bits and the spreading codes. The improved detection scheme presented in [30] is shown to outperform the conventional CDMA system in a Rayleigh fading channel. Soft turbo despreading and decoding for SE-SS system is introduced in [31] that mitigates the problem of error propagation in such systems. In this system, despreading uses the extrinsic likelihood provided by decoding in an iterative way.

2.2.4 Spread Spectrum Systems Employing Parity Bit Selected Spreading Sequence

Spread spectrum systems that employ parity bit selected spreading sequence (SS-PB) is first introduced by D'Amours in [5]. This system utilizes systematic block codes to create a spread spectrum system with data dependent spreading code. Contrary to the

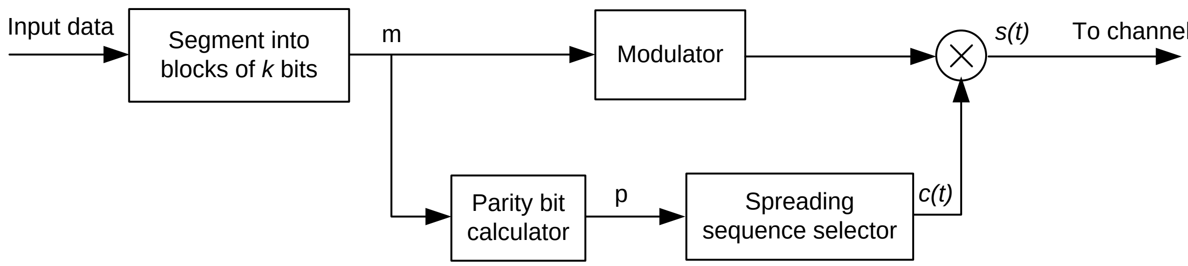

Figure 2.8: Transmitter of PB-SS system.

conventional systematic block codes where the parity bits are appended at the end of information sequence, these parity bits are used to select a spreading sequence from a set of mutually orthogonal spreading sequences.

The transmitter block diagram of a SS-PB system is depicted in Figure 2.8. The information stream is segmented into blocks of k_0 bits. Each information block is input to the parity bit calculator of a systematic (n_0, k_0) linear block encoder that generates $(n_0 - k_0)$ parity bits. To find the parity vector of $\underline{p}$, the information vector of $\underline{m}$ is multiplied by the parity matrix of $\underline{P}$. In other words:

$$\underline{p} = \underline{m}\ \underline{P} \tag{2.4}$$

Parity matrix of $\underline{P}$ is a part of the generator matrix of a systematic code $\underline{G} = [\underline{I}|\underline{P}]$ where $\underline{I}$ is a $k_0 \times k_0$ identity matrix.

The parity vector of $\underline{p}$ associated with the information vector of $\underline{m}$ is input to the spreading sequence selector which outputs one of the $2^{(n_0-k_0)}$ antipodal sequence $c_i(t)$ where i is the decimal representation of $\underline{p}$.

The unique set of spreading sequences allocated to the transmitter is made up of orthogonal sequences. In other words:

$$\int_{jT_b}^{(j+1)T_b} c_i(t)c_m(t)dt = 0 \qquad \text{for } \ i \neq m \tag{2.5}$$

The information is modulated using binary phase shift keying (BPSK) and then the modulated signal is multiplied by the selected spreading sequence. On the time interval $jT_b \leq t \leq (j+1)T_b$, the transmitted signal is:

$$s(t) = b_j A c_i(t) \cos(2\pi f_c t) \tag{2.6}$$

where $j = 0, 1, \ldots, k_0 - 1$, $b_j = 2m_j - 1$, A and f_c are the carrier amplitude and frequency respectively and $c_i(t)$ is the code selected by the parity bits. The spreading code $c_i(t)$ is used to spread all k bits of the information vector.

The receiver block diagram of a SS-PB system is shown in Figure 2.9. Detection of the received signal is done in two parts. In the first part, the spreading code used to spread the information bits in the transmitter is detected. Then the receiver has additional knowledge about the received data which can be used in detection.

To detect the spreading code used in the transmitter, the receiver consists of $2^{(n_0-k_0)}$ matched filters, each matched to different spreading signals. The ith matched filter output on jth signaling interval $(j = 0, 1, \ldots, k_0 - 1)$ is:

$$U_i^{(j)} = \int_{jT_b}^{(j+1)T_b} 2r(t)c_i(t)\cos(2\pi f_c t)dt \tag{2.7}$$

where $r(t)$ is the received signal.

Each message vector of k_0-bit is spread by a spreading sequence which has been selected by the codeword parity bit. Therefore, to find out the most likely employed spreading sequence, the outputs of $2^{(n_0-k_0)}$ matched filters over the k_0 signaling intervals

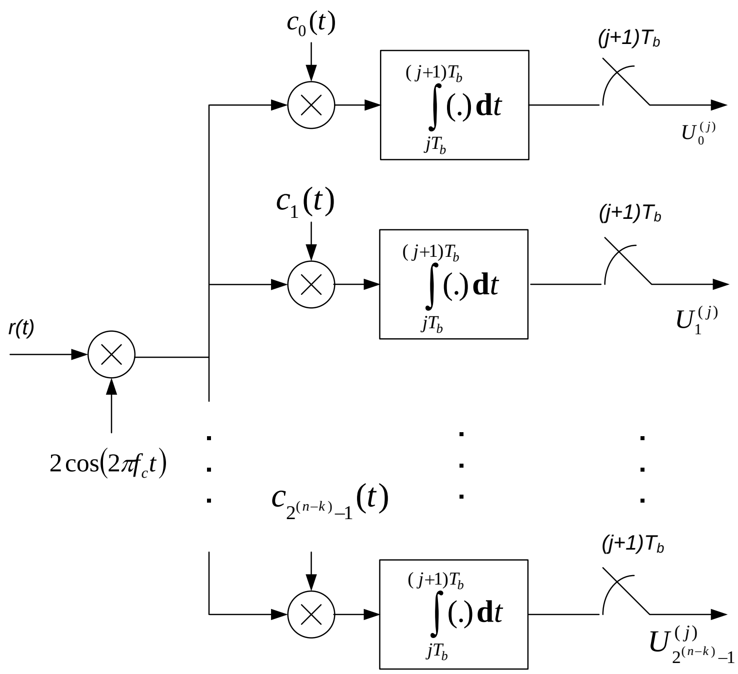

Figure 2.9: Receiver of PB-SS system.

of the message vector must be observed. The most likely employed spreading sequence is $c_i(t)$ if:

$$\sum_{j=0}^{(k_0-1)} \left|U_i^{(j)}\right|^2 > \sum_{j=0}^{(k_0-1)} \left|U_m^{(j)}\right|^2 \qquad \text{for all } \; m \neq i \tag{2.8}$$

When the receiver detects the most likely employed spreading sequence, it uses the matched filter outputs to find the most likely message vector that generates the parity vector that leads to select this spreading sequence. If the employed block code is short, maximum likelihood detection (MLD) would be a good option.

The SS-PB system explained in [5] is based on linear $(10, 6)$ and $(11, 7)$ codes whose parity matrices are:

$$\underline{P_{10,6}} = \begin{bmatrix} 1 & 0 & 1 & 1 \\ 0 & 1 & 1 & 1 \\ 1 & 1 & 0 & 1 \\ 0 & 1 & 0 & 1 \\ 0 & 0 & 1 & 1 \\ 1 & 0 & 0 & 1 \end{bmatrix} \qquad \underline{P_{11,7}} = \begin{bmatrix} 0 & 0 & 1 & 1 \\ 0 & 1 & 0 & 1 \\ 1 & 0 & 1 & 1 \\ 1 & 1 & 1 & 1 \\ 0 & 1 & 1 & 1 \\ 1 & 0 & 0 & 1 \\ 1 & 1 & 0 & 1 \end{bmatrix} \tag{2.9}$$

For example in the case of $(10, 6)$ code, in the transmitter side, information stream is segmented into blocks of $k_0 = 6$ bits that are used to generate parity vectors of length $n_0 - k_0 = 4$. These parity vectors are then used to select a spreading sequence from a set of 16 biorthogonal spreading sequences. For instance, 000000, 011011, 101110 and 110101 each produce parity vector $\underline{p}_0 = 0000$, which means they are all spread by $c_0(t)$.

The upper bound for the asymptotic bit BER performance of the SS-PB systems in an AWGN channel with the double sided noise spectral density of $N_0/2$ is [5]:

$$P_b < \frac{2^{n_0-k_0} - 1}{2^{2k_0}} e^{-k_0 E_b/N_0} \sum_{m=0}^{k_0-1} c_m \left(\frac{k_0 E_b/2}{2N_0} \right)^m \tag{2.10}$$

where P_b is the probability of a bit error, E_b is the bit energy and c_m is given by:

$$c_m = \frac{1}{m!} \sum_{l=0}^{k_0-1-m} \binom{2k_0 - 1}{l} \tag{2.11}$$

Figure 2.10 shows the BER performance of SS-PB systems based on $(10, 6)$ and $(11, 7)$ codes along with the theoretical performance of conventional spread spectrum in AWGN channels. It can be seen that at the BER of 10^{-3}, (10,6) and (11,7) based SS-PB systems, have 1.8dB and 2.3dB gain over uncoded DS-SS systems respectively.

It has been mentioned in [5] that in the SS-PB systems with two-stage detection algorithm, almost all the bit errors that occur in the simulation are due to incorrect detection of the spreading sequence.

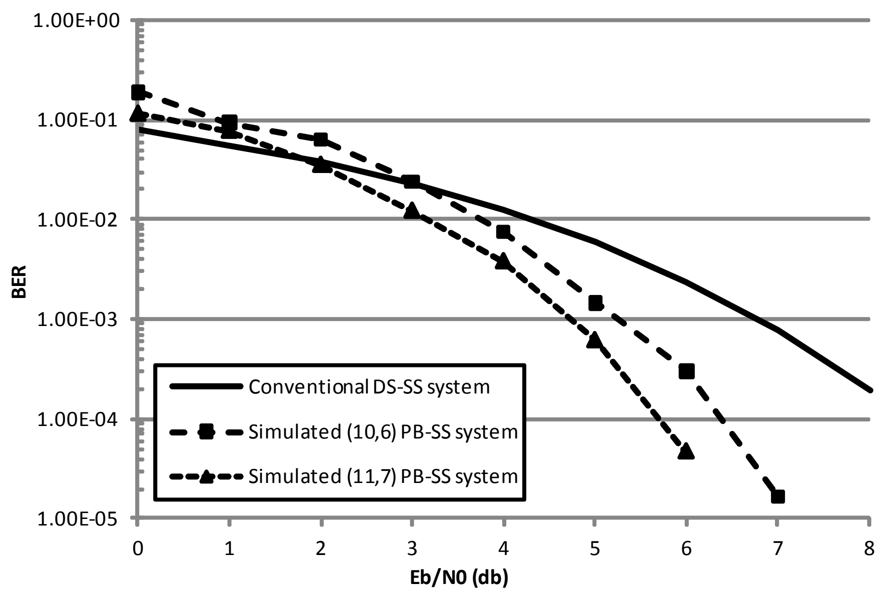

Figure 2.10: BER performance of SS-PB systems based on (10,6) and (11,7) codes in an AWGN channel.

In order to increase the probability of detecting the correct spreading sequence at the receiver, a new detection strategy is proposed by Lawrence and D'Amours in [32] where the decision variables of the two most likely spreading sequences are employed in the detection process. Based on the simulation results presented in [32], at a BER of 10^{-3}, for both (10,6) and (11,7) based SS-PB systems, this method provides 0.7dB improvement over the detection scheme of [5]. The performance of this detection strategy is also shown to be within 0.2 dB of the performance of MLD [32].

Multicarrier Direct Sequence Spread Spectrum Systems with Parity Bit Selected Spreading Sequence

The application of parity bit selected spreading sequence in multicarrier direct sequence spread spectrum (MC-DS-SS) systems is introduced in [6]. Figure 2.11 shows the transmitter block diagram of an MC-DS-SS system employing parity bit selected spreading sequence. The system depicted in Figure 2.11 consists of N_c subcarriers. At the transmitter, the information stream is first converted into N_c parallel streams. On the ith signaling interval, the output of the serial to parallel convertor is the message vector $\underline{m}^{(i)} = [m_0^{(i)}, m_1^{(i)}, \ldots, m_{N_c-1}^{(i)}]$.

Similar to the single carrier SS-PB systems explained in [5], this message vector is then input to the parity bit calculator that uses a systematic (n_0, k_0) linear block code to generate parity vector of $\underline{p}^{(i)} = [p_0^{(i)}, p_1^{(i)}, \ldots, p_{(n_0-k_0)-1}^{(i)}]$.

The parity bit vector of $\underline{p}^{(i)}$ is used to select one spreading sequence from a set of $Q = 2^{(n_0-k_0)}$ mutually orthogonal spreading sequences$\{c_q(t)\}, q = 1, 2, \cdots, Q$. The selected spreading sequence is employed in all N_c subcarriers to spread the modulated signals.

The receiver block diagram of an MC-DS-SS system with parity bit selected spreading waveform is shown in Figure 2.12. Similar to the detection scheme explained in [5] for the single carrier SS-PB systems, the detection of the received signal in the MC-DS-SS systems employing parity bit selected spreading sequence is also performed in two stages [6]. In the first stage, the spreading sequence used in all subcarriers is determined. To do so, decision variables for each spreading sequence in all subcarriers are calculated. The spreading sequence is determined as the sequence that generates the largest summation of decision variables in all N_c subcarriers. In other words, $c_q(t)$ is detected as the spreading

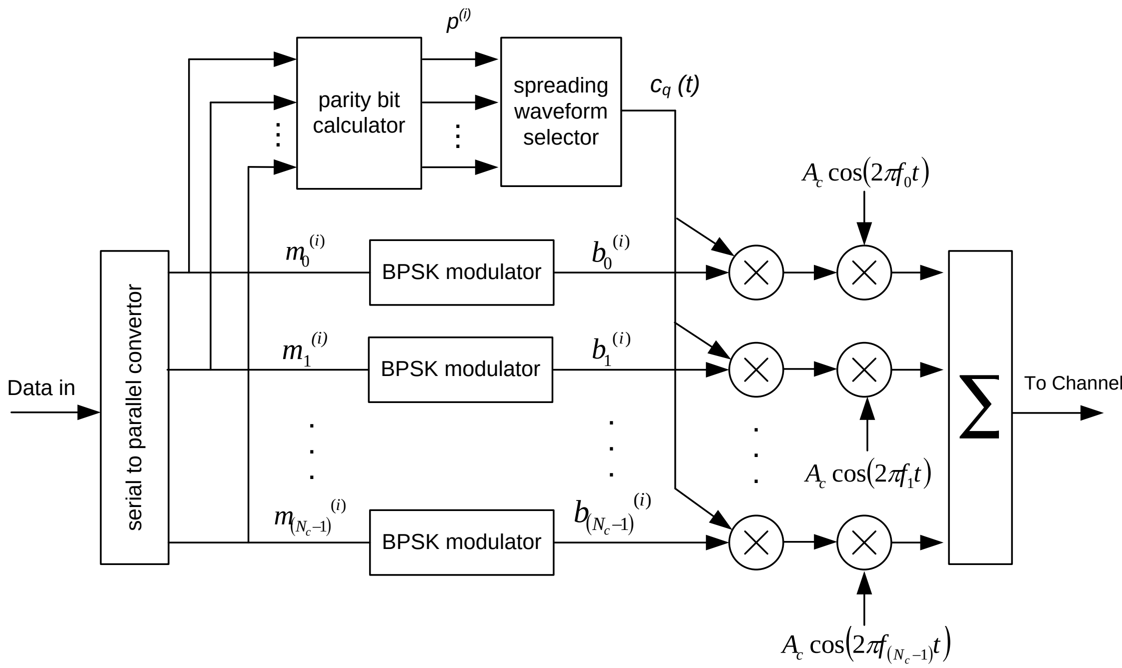

Figure 2.11: Transmitter of MC-DS-SS employing parity bit selected spreading sequence.

signal if:

$$\sum_{v=0}^{N_c-1} \left|U_{v,q}^{(i)}\right|^2 > \sum_{v=0}^{N_c-1} \left|U_{v,r}^{(i)}\right|^2 \quad \text{for all } r \neq q \tag{2.12}$$

where $U_{v,q}^{(i)}$ is the decision variable of the vth subcarrier about the qth spreading sequence on the ith signaling interval.

When the spreading sequence is detected, MLD is used to find the most likely message vector within the all possible message vectors that produce the parity vector of $\underline{p}$ that is associated to $c_q(t)$.

In [6] simulation is performed for a parity bit selected MC-DS-SS system with 4 subcarriers based on the linear (7,4) code. Figure 2.13 shows the BER performance of such system in addition to the conventional MC-DS-SS system in a slowly varying Rayleigh fading channel [6]. It can be seen that parity bit selected MC-DS-SS system

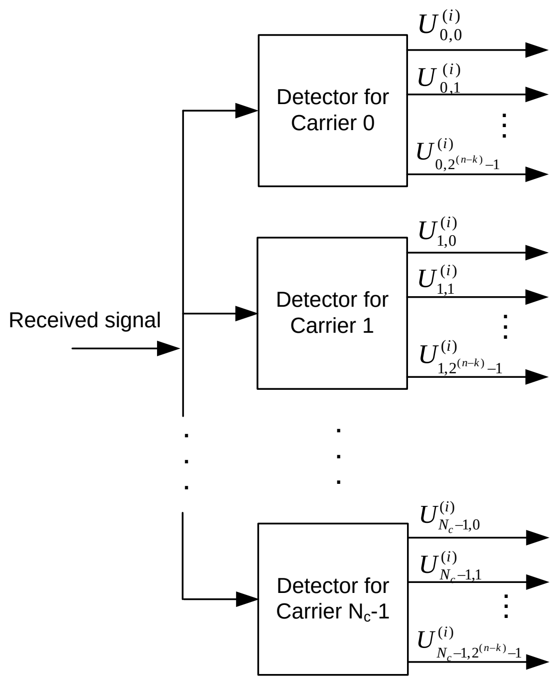

Figure 2.12: Receiver of MC-DS-SS employing parity bit selected spreading sequence.

has a steeper slope in compare to the conventional MC-DS-SS system.

The effect of increasing the number of subcarriers in a parity bit selected MC-DS-SS system is also studied in [6]. If the number of subcarriers in the system increases while the number of spreading waveforms does not change, then the number of message vectors associated with each spreading sequence also increases. Increasing the number of subcarriers leads to a better diversity exploitation by the receiver when determining which spreading code is employed. On the other hand, when we increase the number of message vectors associated with each spreading sequence, the minimum distance between the different message vectors associated with the same parity vector decreases and that leads to a degradation in the accuracy of the second stage of detection. Therefore, there is a trade-off between the number of subcarriers and the number of spreading sequence employed in the system [6].

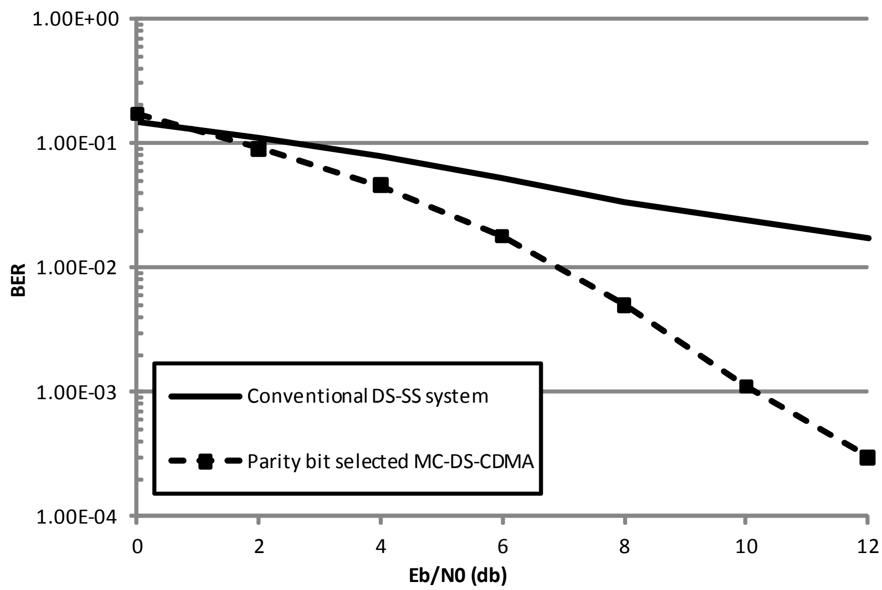

Figure 2.13: BER performance of an MC-DS-SS system with 4 subcarriers and based on (7, 4) code in Rayleigh fading channel.

Parity Bit Selected and Permutation Spreading for MIMO-CDMA Systems

The application of parity bit selected spreading sequence in MIMO-CDMA systems is introduced in [7] where two main spreading code selection techniques are explained.

Contrary to the conventional MIMO-CDMA systems where each transmit antenna uses a different spreading code, in a MIMO-CDMA system with parity bit selected spreading code, depending on the message vector, all the transmit antennas employ the same spreading sequence. Figure 2.14 shows the transmitter block diagram of this system. In such systems, during each signaling interval, the message vector is fed into a parity bit calculator. The calculated parity bits are then used to choose a spreading waveform from a set of mutually orthogonal spreading codes to be used in all transmit antennas during that signaling interval.

Similar to SS-PB systems, all possible message vectors are partitioned into a number

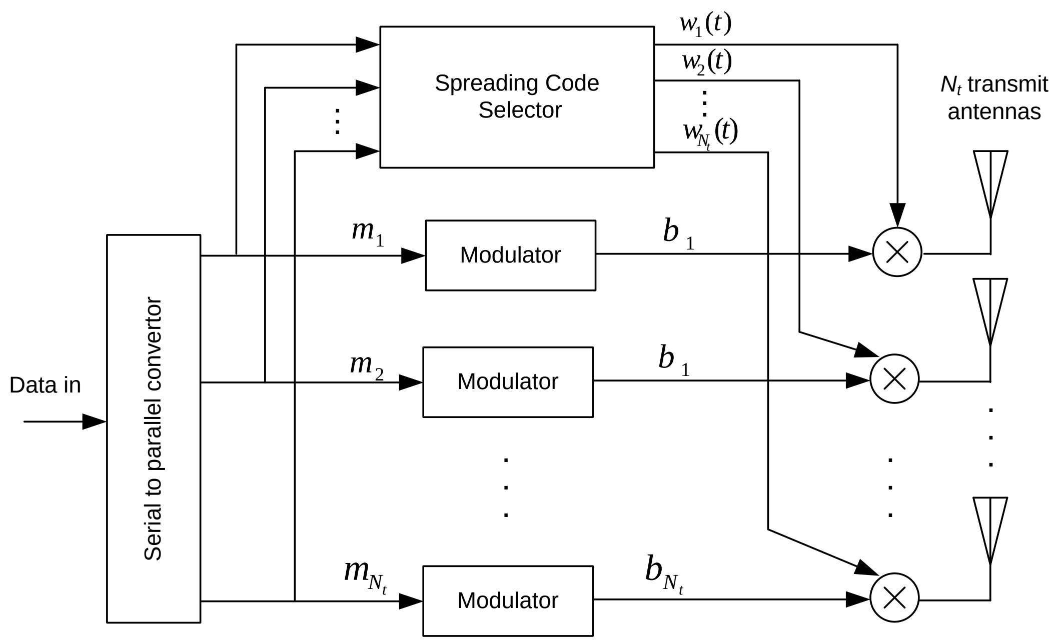

Figure 2.14: Transmitter of a MIMO-CDMA system with parity bit selected and permutation spreading.

of cosets in such a way to maximize the Euclidean distance between the messages in each coset. Depending on which coset the message comes from, a spreading waveform is assigned to all transmit antennas. The system explained in [7] has 4 transmit antennas and there are 8 mutually orthogonal spreading waveforms associated with 8 different cosets. In this case, cosets are shown in (2.13).

$$
\begin{aligned}
M_1 &= \{0000, 1111\}, M_2 = \{0001, 1110\} \\
M_3 &= \{0010, 1101\}, M_4 = \{0011, 1100\} \\
M_5 &= \{0100, 1011\}, M_6 = \{0101, 1010\} \\
M_7 &= \{0110, 1001\}, M_8 = \{0111, 1000\}
\end{aligned} \tag{2.13}
$$

For example if the message vector of [0010] is being transmitted, all 4 transmit an-

Coset	Message vectors	$w_1(t)$	$w_2(t)$	$w_3(t)$	$w_4(t)$
M_1	0000 1111	$c_1(t)$	$c_3(t)$	$c_5(t)$	$c_7(t)$
M_2	0001 1110	$c_8(t)$	$c_1(t)$	$c_4(t)$	$c_5(t)$
M_3	0010 1101	$c_2(t)$	$c_4(t)$	$c_3(t)$	$c_8(t)$
M_4	0011 1100	$c_5(t)$	$c_2(t)$	$c_6(t)$	$c_3(t)$
M_5	0100 1011	$c_6(t)$	$c_7(t)$	$c_1(t)$	$c_4(t)$
M_6	0101 1010	$c_3(t)$	$c_6(t)$	$c_8(t)$	$c_1(t)$
M_7	0110 1001	$c_7(t)$	$c_8(t)$	$c_2(t)$	$c_6(t)$
M_8	0111 1000	$c_4(t)$	$c_5(t)$	$c_7(t)$	$c_2(t)$

Table 2.1: Spreading permutations for MIMO-CDMA systems with 4 transmit antennas.

tennas use the spreading waveform associated with the coset M_3.

In contrast to the system with parity bit selected spreading, in MIMO-CDMA systems with permutation spreading, depending on the coset the message comes from, different permutation of spreading waveforms are assigned to transmit antennas. The permutations used in [7] are listed in Table 2.1. In this system, 8 orthogonal codes are available for a system with 4 transmit antennas. In Table 2.1, $w_i(t)$ represents the spreading waveform used in the ith antenna and $c_j(t)$ is the jth spreading waveform.

The advantage of permutation spreading over parity bit selected spreading is that it creates dependence between different parallel data streams and still maintains the orthogonality between the streams.

Figure 2.15 shows the receiver block diagram of a MIMO-CDMA system with parity bit selected and permutation spreading. As it can be seen in this figure, each receive

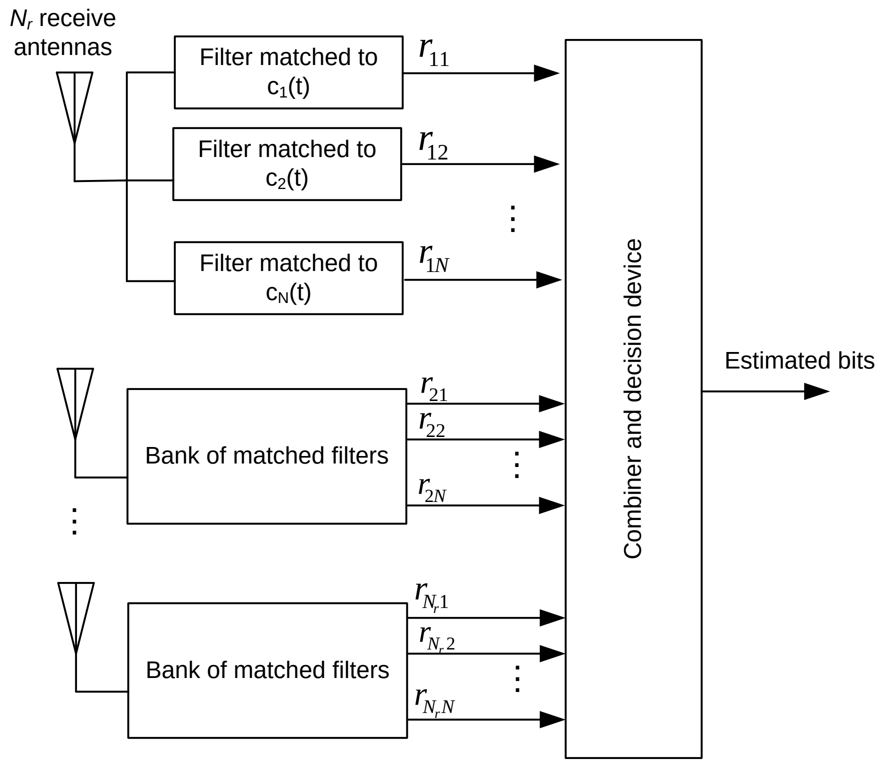

Figure 2.15: Receiver of a MIMO-CDMA system with parity bit selected and permutation spreading.

antenna is equipped with a bank of matched filter for each spreading waveform. The matched filter outputs are used as the decision variables to estimate the transmitted bits using MLD [7].

The BER performance of the 4×1 and 4×4 MIMO-CDMA systems with permutation spreading and parity bit selected spreading in frequency nonselective Rayleigh fading channel are shown in Figures 2.16 and 2.17 respectively [7].

In [33] a new method for designing the spreading permutations based on space time block code matrices is proposed. The proposed method shows a slight improvement in the BER performance compared to the MIMO-CDMA system employing permutation spreading explained in [7].

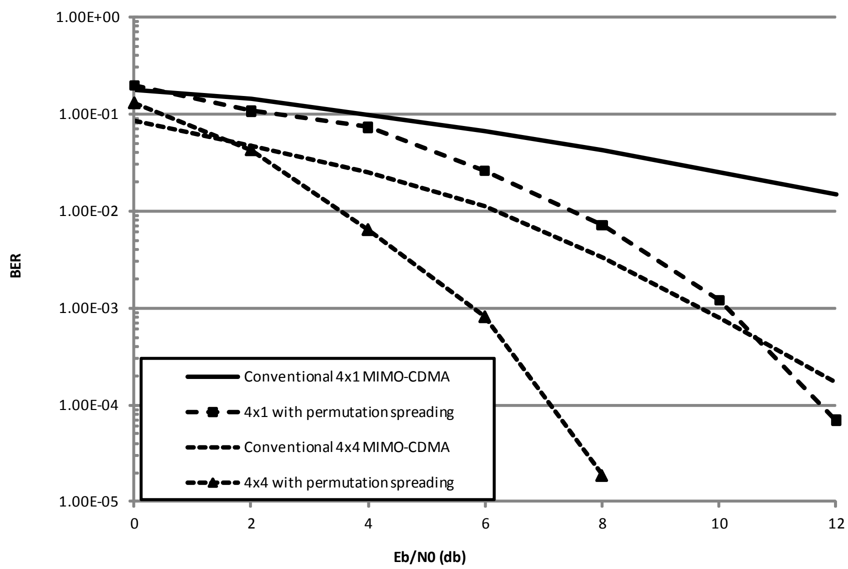

Figure 2.16: BER performance of MIMO-CDMA systems with permutation spreading in compare to the conventional systems.

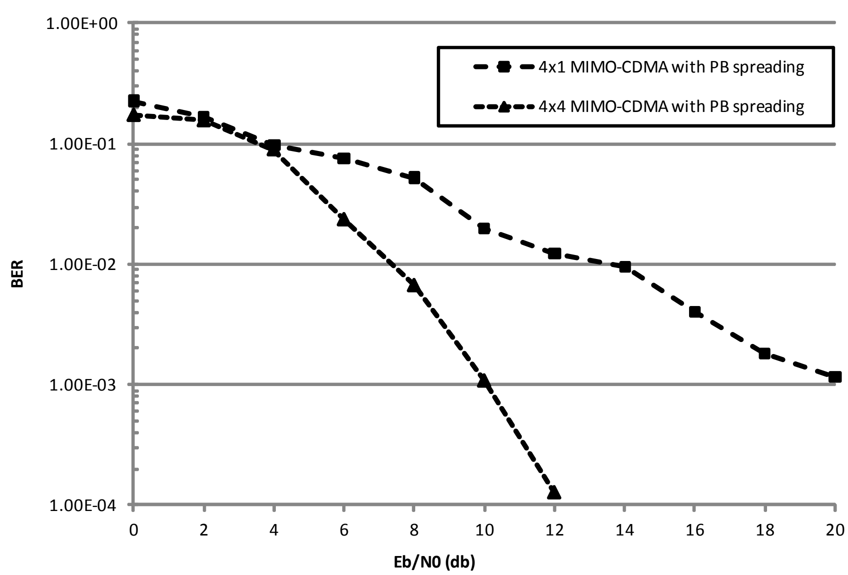

Figure 2.17: BER performance of MIMO-CDMA systems with parity bit selected spreading.

Effect of channel correlation and channel estimation errors on MIMO-CDMA systems with parity bit selected and permutation spreading are studied in [34] and it is shown that permutation spreading technique is more robust than parity bit selected spreading and conventional fixed spreading method in the presence of spatial correlation or channel estimation errors.

In [35], application of permutation spreading technique to asynchronous MIMO-CDMA systems is investigated. In such systems, Hadamard codes are used for antenna separation and different users are separated by Gold scrambling codes. Simulation results in [35] show an advantage of 7dB at the BER of 10^{-3} for MIMO-CDMA systems with permutation spreading compared to the conventional MIMO-CDMA systems in heavily loaded conditions.

Chapter 3

Turbo Receiver for Spread Spectrum Systems Employing Parity Bit Selected Spreading Sequences

3.1 Introduction

The detection strategy for SS-PB systems proposed in [5] consists of two stages. In the first stage, the receiver detects the spreading code used by the transmitter and then in the second stage, assuming that the spreading code determined in the first step is correct, the receiver then compares the output of the corresponding matched filter to the possible messages that can be carried by that specific spreading code. It then selects the most likely message block associated with the spreading code determined in step 1. In this detection strategy, most bit errors occur due to incorrect spreading code determination in the first stage [5]. In [36] a soft output detector for SS-PB systems is proposed. In the proposed system of [36], information bits are first convolutionally encoded prior to being used in a SS-PB system. Reliability of the detected bits in the receiver is calculated in terms of log likelihood ratios (LLRs) and then used as the input to a soft input Viterbi decoder. It is shown in [36] that the concatenation of convolutional coding and parity bit selected spreading provides a significant BER improvement even without the use of iterative turbo techniques.

The impressive performance of turbo (iterative) processing techniques that soon followed the discovery of turbo codes [37], has inspired many researchers to utilize the turbo principle in a variety of ways in the architecture of the receiver. In this chapter we introduce a new turbo receiver for SS-PB systems. In the transmitter data bits are first convolutionally encoded and interleaved before being used to calculate the parity bits and select the spreading codes in a SS-PB system. In this system, the parity bit selected technique acts as an inner encoder without allocating any transmit energy to the additional redundancy provided by this technique. Turbo processing is implemented in the receiver by iteratively exchanging soft information about the coded bits between a SISO detector and a SISO decoder. In this system, detection is performed by incorporating the *extrinsic* information provided by the decoder in the last iteration into the received signal to calculate the likelihood of each detected bit in terms of LLRs which are used as the input for a SISO decoder. Simulation results presented in this chapter show a significant improvement in the performance of SS-PB systems when a turbo receiver is implemented.

3.2 System Model

We consider a coded spread spectrum system where the information bits are convolutionally encoded, passed through an interleaver and then input to the SS-PB system. The block diagram of the transmitter is shown in Figure 3.1. The binary information stream $\{d[i]\}$ is convolutionally encoded with code rate R and after passing through

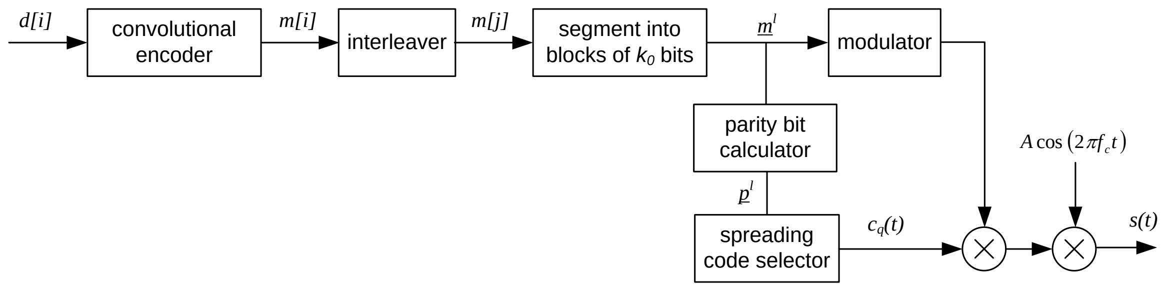

Figure 3.1: Transmitter of a coded SS-PB system.

an interleaver, the stream of interleaved coded bits $\{m[j]\}$ is segmented into blocks of k_0 bits: $\underline{m}^l = (m_1^l, m_2^l, \cdots, m_{k_0}^l)$. This message vector is then input to the parity bit calculator that uses a systematic (n_0, k_0) linear block code to generate parity vector of $\underline{p}^l = (p_1^l, p_2^l, \cdots, p_{(n_0-k_0)}^l)$.

The parity bit vector of $\underline{p}^l$ is used to select a spreading code from a set of $Q = 2^{(n_0-k_0)}$ mutually orthogonal spreading sequences: $\{c_q(t)\}$, $q = 1, 2, \cdots, Q$. The selected spreading code is then employed to spread the modulated signals of all k_0 bits of $\underline{m}^l$.

The stream of coded bits $\{m[j]\}$ is modulated using BPSK to obtain $\{b[j]\}$ and then each block of modulated bits $\underline{b}^l = (b_1^l, b_2^l, \cdots, b_{k_0}^l)$ spread by one of the Q spreading codes depending on which parity vector is generated by $\underline{m}^l$. On the time interval $jT_b \leq t \leq (j+1)T_b$, the transmitted signal is:

$$s(t) = b[j]Ac_q(t)\cos(2\pi f_c t) \tag{3.1}$$

where $q = 1, 2, \cdots, Q$, $b[j] = 2m[j]-1$, A and f_c are the carrier amplitude and frequency of the transmitted signal respectively. The spreading code $c_q(t)$ is chosen based on which block of coded bits $b[j]$ comes from.

After transmission through the channel, the received signal is $r(t) = \alpha(t)s(t) + n(t)$, where $\alpha(t)$ is the fading amplitude and $n(t)$ is a zero-mean white Gaussian noise process with power spectral density of $\frac{N_0}{2}$.

To be able to determine which spreading code is used in the transmitter, the receiver is equipped with Q matched filters, each matched to one of the spreading codes of set $\{c_q(t)\}$. The output of the qth matched filter on the jth signaling interval is:

$$z_q[j] = \int_{jT_b}^{(j+1)T_b} 2\alpha(t)r(t)c_q(t)\cos(2\pi f_c t)dt$$

$$= \begin{cases} \alpha[j]b[j]AT_b + n_q[j] & \text{if } c_q(t) \text{ is used} \\ n_q[j] & \text{otherwise} \end{cases} \tag{3.2}$$

where $n_q[j]$ is the response of the qth matched filter to the input noise on the jth signaling interval and has a Gaussian distribution with zero mean and variance $N_0 T_b$. As the filters are matched to orthogonal spreading codes, the output noise samples are uncorrelated. In other words, $E[n_q[j]n_r[j]] = 0$ for $q \neq r$. Assuming a frequency non-selective slowly varying fading channel where the fading amplitude remains constant over one bit interval, $\alpha[j]$ in (3.2) represents the complex fading gain of the channel over the jth signaling interval.

As the same spreading code is used to spread all the bits of each block of modulated bits, the output of the matched filters should be observed over k_0 signaling intervals: $\underline{z}_q^l = (z_{q1}^l, z_{q2}^l, \cdots, z_{qk_0}^l)$. In the following sections, for notational simplicity, we drop the block index l in all vectors representing a block of information bits whenever possible.

3.3 Turbo Receiver

The proposed turbo receiver is shown in Figure 3.2. It consists of two stages: a SISO detector specifically designed for SS-PB systems followed by a SISO channel decoder.

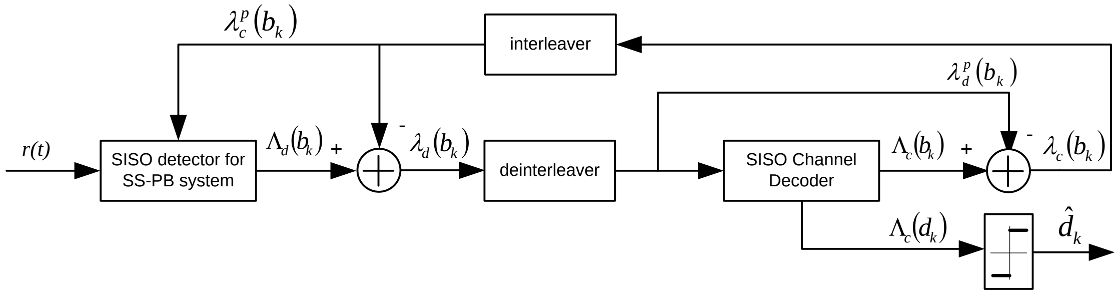

Figure 3.2: Proposed turbo receiver for SS-PB systems.

These two stages are connected via an interleaver and a deinterleaver. The proposed receiver structure in this chapter is inspired by the system introduced in [38] where a turbo multiuser detection is proposed for coded code division multiple access (CDMA) systems.

The SISO detector observes the matched filter outputs and by considering the *extrinsic* information provided by the SISO channel decoder in the previous iteration, delivers the *a posteriori* LLR of a transmitted "+1" and a transmitted "−1" for each coded bit,

$$\Lambda_d\left(b_k\right) \triangleq \ln \frac{P\left(b_k=+1|r(t)\right)}{P\left(b_k=-1|r(t)\right)}, \quad k=1,2,\cdots,k_0. \tag{3.3}$$

Using Bayes' rule, we can rewrite (3.3) as

$$\Lambda_d\left(b_k\right) = \underbrace{\ln \frac{P\left(r(t)|b_k=+1\right)}{P\left(r(t)|b_k=-1\right)}}_{\lambda_d(b_k)} + \underbrace{\ln \frac{P\left(b_k=+1\right)}{P\left(b_k=-1\right)}}_{\lambda_c^p(b_k)}. \tag{3.4}$$

The second term in (3.4), denoted by $\lambda_c^p\left(b_k\right)$, represents the *a priori* LLR of the coded bit b_k calculated by the SISO channel decoder in the previous iteration (as indicated by a superscript p to show the previous iteration). For the first round of detection (i.e. first iteration) when there is no prior information available about the likelihood of coded bits, we assume all the coded bits can be "+1" or "−1" equiprobably. In other words

$\lambda_c^p(b_k) = 0$ for $k = 1, 2, \cdots, k_0$.

The first term in (3.4), denoted by $\lambda_d(b_k)$, represents the *extrinsic* information provided by the SISO detector about the kth coded bit, b_k. This *extrinsic* information is calculated by the SISO detector based on the received signal $r(t)$ (or in other words the matched filter outputs), the structure of the SS-PB system and the prior information provided by the SISO channel decoder in the previous iteration about the coded bits other than the kth bit, $\{\lambda_c^p(b_j), j \neq k\}$. This *extrinsic* information is then passed through a deinterleaver and feed into the SISO channel decoder as *a priori* information in the next iteration.

By making use of the prior information provided by the SISO detector, $\lambda_d(b_k)$, and also considering the trellis structure of the convolutional code used in the transmitter, the SISO channel decoder calculate the *a posteriori* LLR of each coded bit,

$$\begin{aligned}\Lambda_c(b_k) &\triangleq \ln \frac{P\left(b_k = +1 | \{\lambda_d^p(b_k)\}_{k=1}^{k_0}; \text{decoding}\right)}{P\left(b_k = -1 | \{\lambda_d^p(b_k)\}_{k=1}^{k_0}; \text{decoding}\right)} \\ &= \lambda_c(b_k) + \lambda_d^p(b_k)\end{aligned} \tag{3.5}$$

where the *a posteriori* probabilities are conditioned on the *extrinsic* information from the SISO detector in the last iteration, $\{\lambda_d^p(b_k)\}_{k=1}^{k_0}$, and the coding structure shown by "decoding" in (3.5).

It can be seen in (3.5) that the *a posteriori* LLR $\Lambda_c(b_k)$, is the summation of the prior information $\lambda_d^p(b_k)$ and the *extrinsic* information $\lambda_c(b_k)$ that is the information gleaned about b_k from prior information of other coded bits based on the trellis structure of the channel encoder. The SISO channel decoder also computes the *a posteriori* LLR of the data bits $\Lambda_c(d_k)$, that is used in the last iteration to make decision about the decoded bits. The *extrinsic* information delivered in this stage $\lambda_c(b_k)$, is then passed to an interleaver and fed back to the SISO detector as the *a priori* information in the next iteration (i.e. $\lambda_c^p(b_k)$).

The SISO channel decoder discussed in this chapter that provides the *a posteriori* information for both coded and data bits is based on the algorithm introduced in [38] which is a slight modification of the well know BCJR algorithm [39].

In what follows we explain how the SISO detector for a SS-PB system uses the matched filter outputs to combine with the prior information from the channel decoder in the last iteration to calculate the *extrinsic* information for each detected bit.

Let us define

$$\mathcal{B}_+^k \triangleq \{\underline{b} \in (+1,-1)^{k_0}, b_k = +1\}, \tag{3.6}$$

that includes all the k_0 coded bit vectors whose kth bit is "+1". Similarly $\mathcal{B}_-^k$ is defined as

$$\mathcal{B}_-^k \triangleq \{\underline{b} \in (+1,-1)^{k_0}, b_k = -1\}. \tag{3.7}$$

The *extrinsic* information delivered by the SISO detector $\lambda_d(b_k)$ can be rewritten as

$$\begin{aligned} \lambda_d(b_k) &= \ln \frac{P(r(t)|b_k = +1)}{P(r(t)|b_k = -1)} \\ &= \ln \frac{\sum_{\underline{b}\in\mathcal{B}_+^k} P(\underline{Z}|\underline{b}) \prod_{j\neq k} P(b_j)}{\sum_{\underline{b}\in\mathcal{B}_-^k} P(\underline{Z}|\underline{b}) \prod_{j\neq k} P(b_j)} \end{aligned} \tag{3.8}$$

where $P(b_j)$ is the probability that the jth bit equals to b_j for $b_j \in \{+1,-1\}$, and $\underline{Z}$ is a $Q \times k_0$ matrix representing the Q matched filter outputs over k_0 signaling intervals. In other words the qth row of $\underline{Z}$ is $\underline{z}_q$ [cf. (3.2)]. Equation (3.8) holds true because the received signal is fully represented by the outputs of the filters matched to Q orthogonal spreading sequences. It is also necessary and sufficient to observe the matched filter outputs over k_0 signaling intervals to be able to completely extract the extrinsic information of each coded bit.

Let us define matrix $\underline{Y}$ as a noise free Q matched filter outputs over k_0 signaling interval when $\underline{b}$ is sent and let v be the index of the spreading code selected by $\underline{b}$. In other words,

$$y_{ij} = \begin{cases} b_j & \text{if} \quad i = v \\ 0 & \text{otherwise} \end{cases} \tag{3.9}$$

where y_{ij} is the ith noise free matched filter output in the jth signaling interval. From (3.2) since the noise has a normal distribution, (3.8) can be rewritten as

$$\lambda_d\left(b_k\right) = \ln \frac{\sum_{\underline{b}\in\mathcal{B}_+^k} \exp\left(-\frac{1}{N_0} d(\underline{b})^2\right) \prod_{j\neq k} P(b_j)}{\sum_{\underline{b}\in\mathcal{B}_-^k} \exp\left(-\frac{1}{N_0} d(\underline{b})^2\right) \prod_{j\neq k} P(b_j)} \tag{3.10}$$

where

$$d(\underline{b})^2 = \sum_{i=1}^{Q} \sum_{j=1}^{k_0} |z_{ij} - \alpha_j y_{ij}|^2. \tag{3.11}$$

in which α_j is the complex fading gain that affects the jth bit and $|.|$ denotes the magnitude.

Using (3.9), we can simplify (3.11) as

$$d(\underline{b})^2 = \sum_{i=1}^{Q} \sum_{j=1}^{K_0} |z_{ij}|^2 + \sum_{j=1}^{k_0} |\alpha_j y_{vj}|^2 - \sum_{j=1}^{K_0} (\alpha_j^* y_{vj}^* z_{vj} + \alpha_j y_{vj} z_{vj}^*) \tag{3.12}$$

where the superscript $*$ denotes the complex conjugate.The second term in (3.12) can be written as

$$\sum_{j=1}^{k_0} |\alpha_j y_{vj}|^2 = \sum_{j=1}^{k_0} |\alpha_j b_j|^2 = \sum_{j=1}^{k_0} |\alpha_j|^2 \tag{3.13}$$

which is the summation of squared fading amplitudes that affect all k_0 bits and is independent of $\underline{b}$. The third term in (3.12) denoted by $e(\underline{b})$ can also be simplified as

$$e(\underline{b}) = -\sum_{j=1}^{k_0}(\alpha_j^* y_{vj}^* z_{vj} + \alpha_j y_{vj} z_{vj}^*) = -\sum_{j=1}^{k_0}(\alpha_j^* b_j z_{vj} + \alpha_j b_j z_{vj}^*). \tag{3.14}$$

When the signal is passed through an AWGN channel, $\alpha_j = 1$ for all $j = 1, \cdots, k_0$, the derivation of $e(\underline{b})$ in (3.14) can be simplified as

$$e(\underline{b}) = -2\sum_{j=1}^{k_0} z_{vj} b_j = -2\underline{z}_v \cdot \underline{b} \tag{3.15}$$

where the last term denotes the dot product between the vector of the vth matched filter output $\underline{z}_v$, and the coded bit vector $\underline{b}$. Note that the first and the second terms in (3.12) are independent of $\underline{b}$ and therefore will be canceled in (3.10).

The *a priori* probabilities in (3.8) can be calculated based on their LLR's $\lambda_c^p(b_k)$ as follows [38]

$$\begin{aligned} P(b_j) &= \frac{\exp\left(b_j \lambda_c^p(b_j)\right)}{1 + \exp\left(b_j \lambda_c^p(b_j)\right)} \\ &= \frac{\exp\left(\frac{1}{2} b_j \lambda_c^p(b_j)\right)}{\exp\left(-\frac{1}{2} b_j \lambda_c^p(b_j)\right) + \exp\left(\frac{1}{2} b_j \lambda_c^p(b_j)\right)} \\ &= \frac{\cosh\left(\frac{1}{2} b_j \lambda_c^p(b_j)\right)\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p(b_j)\right)\right]}{2\cosh\left(\frac{1}{2} b_j \lambda_c^p(b_j)\right)} \\ &= \frac{1}{2}\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p(b_j)\right)\right]. \end{aligned} \tag{3.16}$$

Substituting (3.12), (3.14) (or (3.15) in the case of AWGN channel) and (3.16) into

(3.10), we obtain

$$\lambda_d\left(b_k\right) = \ln \frac{\sum_{\underline{b}\in\mathcal{B}_+^k} \exp\left(\frac{-e(\underline{b})}{N_0}\right) \prod_{j\neq k}\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p\left(b_j\right)\right)\right]}{\sum_{\underline{b}\in\mathcal{B}_-^k} \exp\left(\frac{-e(\underline{b})}{N_0}\right) \prod_{j\neq k}\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p\left(b_j\right)\right)\right]}. \tag{3.17}$$

By having (3.17) we can summarize the algorithm of the SISO detector in finding the *extrinsic* information of each coded bit as follows:

- First $\mathcal{B}_+^k$ and $\mathcal{B}_-^k$ are formed by dividing all the vectors of k_0 coded bits, $\underline{b}$, into the vectors in which the kth bit is "+1" and "−1" respectively.

- Based on the code structure used in the transmitter for parity bit calculation, the spreading code index v is determined for each vector $\underline{b}$.

- $e(\underline{b})$ is calculated for each vector $\underline{b}$ using (3.14) (or (3.15) in the case of AWGN channel) that computes the dot product between $\underline{b}$ and the vector of the vth matched filter output over k_0 signaling intervals.

- *a priori* probabilities $P(b_k)$ are calculated using (3.16) based on the *extrinsic* LLR provided by the SISO channel decoder in the previous iteration.

- By knowing $e(\underline{b})$ and $\lambda_c^p\left(b_k\right)$ for each coded bit, the *extrinsic* LLR in the output of the SISO detector is computed by (3.17).

The computed *extrinsic* LLRs $\lambda_d\left(b_k\right)$ are then passed through a deinterleaver and fed into the SISO channel decoder as the *a priori* information. The SISO channel decoder provides the *a posteriori* LLRs for both coded and decoded (data) bits. The *extrinsic* LLRs $\lambda_c\left(b_k\right)$ at the output of the channel decoder are formed by subtracting the the *a priori* LLRs $\lambda_d^p\left(b_k\right)$ from the *a posteriori* LLRs of the coded bits $\Lambda_c\left(b_k\right)$. In the last iteration, *a posteriori* LLRs of the data bits are used to make decision on data bits, in other words the estimated data bit, $\hat{d}$, is the sign of $\lambda_c(d_k)$ in the last iteration.

3.4 Simulation Results

In the transmitter, at first, data bits are encoded by a 1/2 rate convolutional encoder with constraint length of 5 and generators 23, 35 in octal notation. The coded bits are passed through a random interleaver before being fed into the SS-PB modulator. The simulations are performed for both AWGN and Rayleigh fading channels. The Rayleigh fading is assumed to be frequency non-selective slowly varying with autocorrelation function of $J_0(2\pi f_d T_b)$ where J_0 is the zero order modified Bessel function of the first type and f_d is the Doppler spread [40]. The simulation is performed for the channel with fading rate of $f_d T_b = 0.05$ and a coherent detection is assumed. We consider the SS-PB systems based on the $(10, 6)$ and $(11, 7)$ block encoder whose parity matrices are shown in (2.9).

The BER performance of the proposed turbo receiver in AWGN and Rayleigh fading channels for the coded $(10, 6)$ based SS-PB system along with the conventional coded DS-SS system are shown in Figures 3.3 and 3.4 respectively. These simulations are repeated for $(11, 7)$ based PB-SS systems and the results are shown in Figures 3.5 and 3.6. The simulation results in both AWGN and fading channels show a significant performance improvement of the coded SS-PB system over multiple iterations.

3.5 Discussion

In this chapter, we have introduced a novel turbo receiver for spread spectrum systems with parity bit selected spreading sequences. An algorithm for the SISO detector in

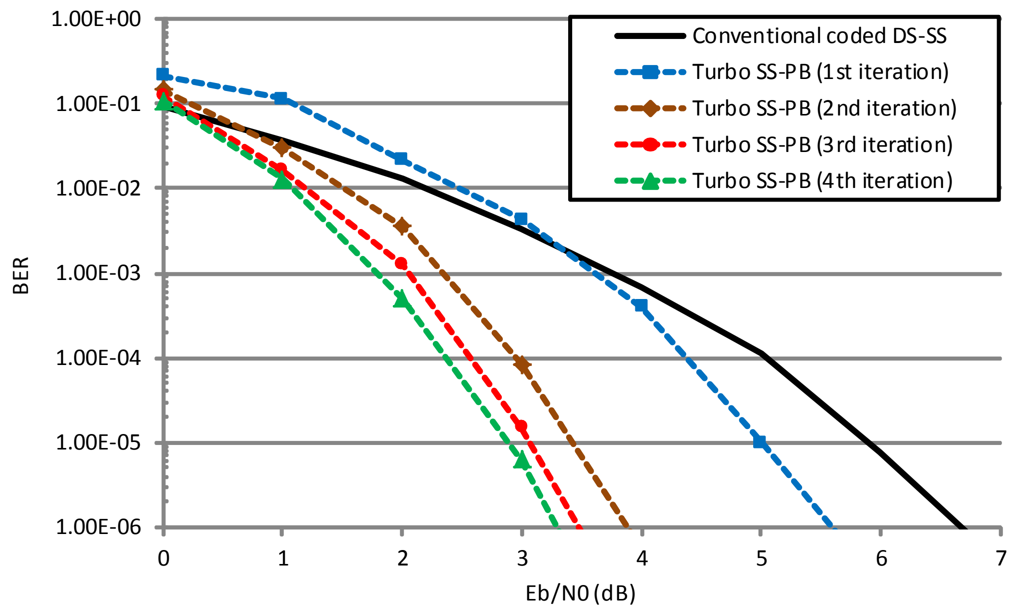

Figure 3.3: BER Performance of the proposed turbo receiver for the coded $(10, 6)$ based SS-PB system in the AWGN channel.

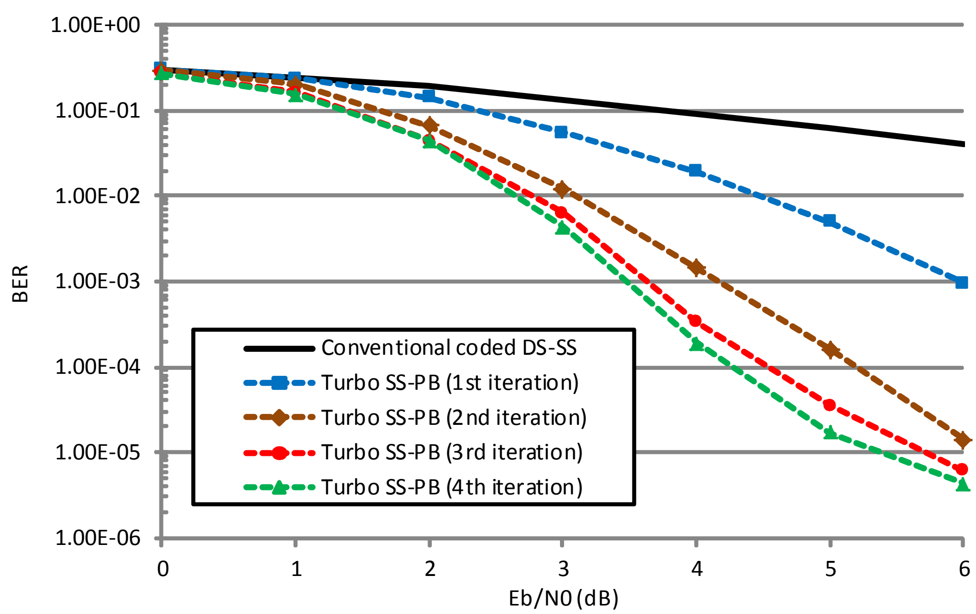

Figure 3.4: BER Performance of the proposed turbo receiver for the coded $(10, 6)$ based SS-PB system in the Rayleigh fading channel.

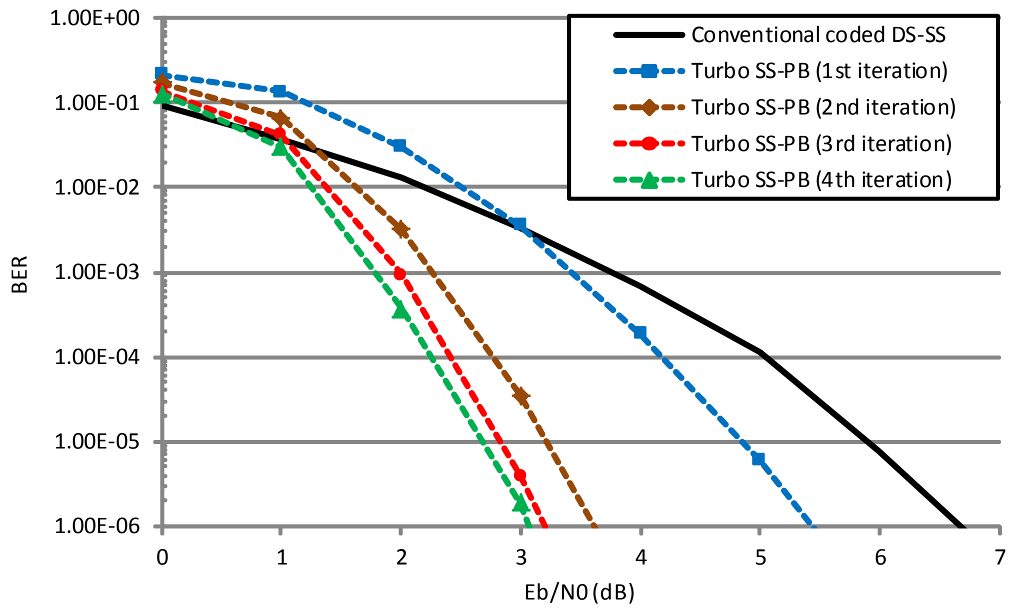

Figure 3.5: BER Performance of the proposed turbo receiver for the coded $(11, 7)$ based SS-PB system in the AWGN channel.

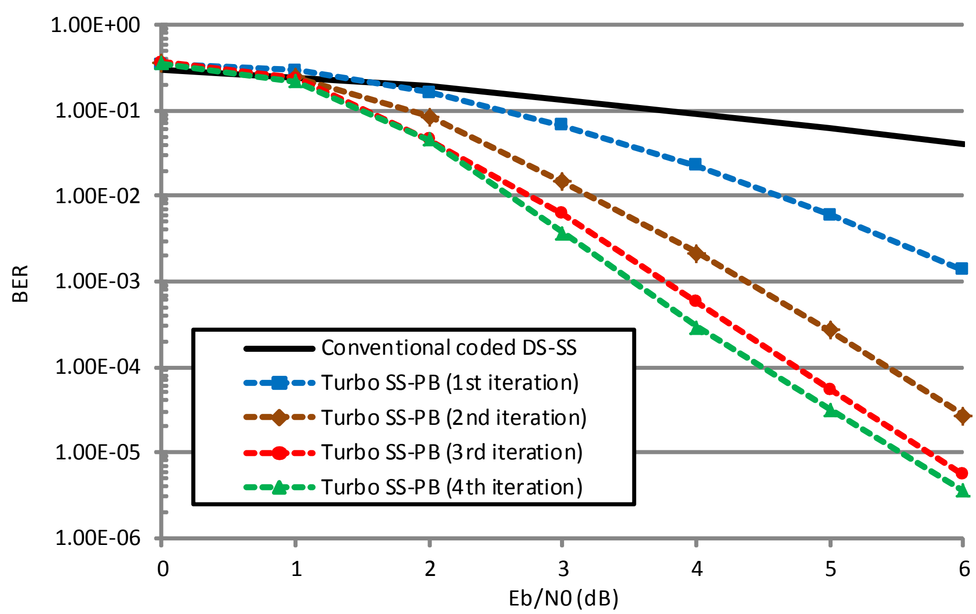

Figure 3.6: BER Performance of the proposed turbo receiver for the coded $(11, 7)$ based SS-PB system in the Rayleigh fading channel.

SS-PB systems is developed in which the LLRs of each bit is calculated based on the received signal, the *a priori* information from the previous iteration and the code structure of the SS-PB system. The proposed turbo receiver consists of a SISO detector for SS-PB systems followed by a SISO channel decoder. Iteration of LLRs between these two components provides a significant performance improvement when compared to the conventional coded SS systems. If the SS-PB system is used in a multiuser scenario, the improved BER performance of the proposed system can be traded off against additional users in a CDMA system.

Chapter 4

Turbo Receiver for Multicarrier Spread Spectrum Systems Employing Parity Bit Selected Spreading Sequences

4.1 Introduction

In multicarrier direct-sequence spread spectrum systems (MC-DS-SS), in order to combat the frequency selectivity of the channel, the available spectrum is divided into a set of distinct sub-bands. Using direct sequence spreading, the data to be transmitted is spread to the bandwidth of each sub-band and then used to modulate one or more orthogonal carriers. Frequency diversity is obtained by transmitting the same data on multiple carriers. This comes at the expense of spectral efficiency. For optimal spectral efficiency, distinct information is carried by each carrier. In order to maximize the frequency diversity, in the discussed system of [41] and [42], the same data is transmitted on every carrier. In these works, the fading on each sub-band is assumed to be Rayleigh and independent of the fading on other sub-bands. In [42], each user is assigned a unique set of mutually orthogonal spreading waveforms and then each sub-band is spread with a different spreading waveform. In [43], the data to be transmitted is serial to parallel converted and then used to modulate a subset of the orthogonal carriers. The MC-DS-SS system considered in this chapter is of the later one where every subcarrier transmits different sets of modulated bits.

The extension of parity bit selected spreading technique into MC-DS-SS systems is proposed in [6] and is explained in Chapter 2. A soft output detector for MC-SS-PB systems is proposed in [44]. In the proposed system of [44], information bits are first convolutionally encoded prior to being used in a MC-SS-PB system. Reliability of the detected bits in the receiver is calculated in terms of LLRs and then used as the input to a soft input Viterbi decoder. It is shown in [44] that a significant BER improvement can be achieved when a convolutional encoder is concatenated with a MC-SS-PB system even without the use of any iterative turbo techniques.

In this chapter we proposed a turbo receiver for MC-SS-PB systems. In the transmitter, the parity bit selected scheme of MC-SS-PB is used as an inner encoder that is concatenated by a convolutional encoder. Similar to the turbo receiver of SS-PB systems discussed in Chapter 3, the receiver implements a turbo processing by iteratively exchanging soft information about the coded bits between a SISO detector and a SISO decoder.

4.2 System Model

The block diagram of the transmitter is shown in Figure 4.1. The binary information stream $\{d[i]\}$ is convolutionally encoded with code rate R and after passing through an interleaver, the stream of interleaved coded bits $\{m[j]\}$ is converted into N_c (number of subcarriers) parallel streams. On the lth signaling interval, the vector of bits to be transmitted on N_c subcarriers is $\underline{m}^l = (m_1^l, m_2^l, \cdots, m_{N_c}^l)$. This vector is then input to the parity bit calculator of a systematic (n_0, k_0) linear block encoder where $k_0 = N_c$ to generate the parity vector of $\underline{p}^l = (p_1^l, p_2^l, \cdots, p_{(n_0-k_0)}^l)$.

The parity bit vector of $\underline{p}^l$ is used to select a spreading code from a set of $Q = 2^{(n_0-k_0)}$ mutually orthogonal spreading sequences: $\{c_q(t)\}$, $q = 1, 2, \cdots, Q$. The selected spreading code is then employed to spread the modulated signals of all N_c subcarriers on the lth signaling interval.

The streams of parallel bits in all subcarriers are modulated using BPSK and then the

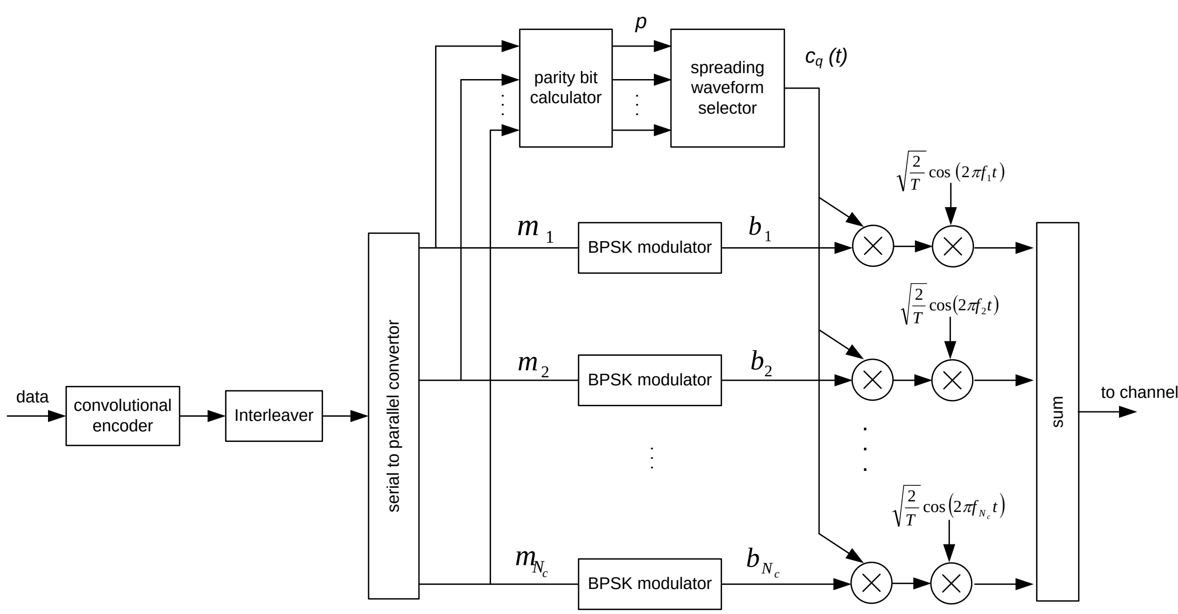

Figure 4.1: Transmitter of a coded MC-SS-PB system.

modulated bits in all subcarriers $\underline{b}^l = (b_1^l, b_2^l, \cdots, b_{N_c}^l)$ spread by one of the Q spreading codes depending on which parity vector is generated by $\underline{m}^l$. Each parallel spread signal is then upconverted by a unique orthogonal carrier and multiplexed with other subcarrier signals. We assume that the carriers are sufficiently spaced so that the fading encountered by each can be assumed independent. We also assume that the data rates of each parallel data stream and spreading factor are low enough that the fading encountered by each carrier is frequency nonselective.

On the time interval $lT \leq t \leq (l+1)T$, the transmitted signal is:

$$s(t) = \sum_{v=1}^{N_c} \text{Re}\left[\sqrt{\frac{2}{T}} b_v^l c_q(t) e^{j2\pi f_v t}\right] \tag{4.1}$$

where f_v is the carrier frequency of the vth subcarrier and the spreading code $c_q(t)$, $q = 1, 2, \cdots, Q$ is chosen by parity vector $\underline{p}^l$.

The signal is transmitted through the channel that is assumed to be slowly-varying, frequency selective, Rayleigh channel, where the DS-SS signals in each sub-channel is fading nonselectively and independently. After transmission through the channel, the received signal is

$$r(t) = \sum_{v=1}^{N_c} \text{Re}\left[\alpha_v \sqrt{\frac{2}{T}} b_v^l c_q(t) e^{j2\pi f_v t}\right] + n(t) \tag{4.2}$$

where α_v is the complex fading gain of the vth sub-channel and $n(t)$ is a zero-mean Gaussian noise with double sided spectral density of $N_0/2$. The channel is assumed to be slowly-varying Rayleigh fading; therefore, α_v is constant over one signaling interval and $|\alpha_v|$ and $\angle\alpha_v$ are, respectively, an independently, identical distribution (i.i.d) Rayleigh random variable with unit variance and i.i.d uniform random variable over $[0, 2\pi)$.

To be able to determine which spreading code is used in the transmitter, the receiver is equipped with Q matched filters in each subcarrier, each matched to one of the spreading codes of set $\{c_q(t)\}$. The output of the qth matched filter in the vth subcarrier on the lth signaling interval is:

$$z_{v,q}^l = \begin{cases} \alpha_v^l b_v^l + n_{v,q}^l & \text{if } c_q(t) \text{ is used} \\ n_{v,q}^l & \text{otherwise} \end{cases} \tag{4.3}$$

where $n_{v,q}^l$ is the response of the qth matched filter of the vth sub-channel to the input noise on the lth signaling interval and has a Gaussian distribution with zero mean and variance N_0T. As the filters are matched to orthogonal spreading codes, the output noise samples are uncorrelated. In other words, $E\left[n_{v,q}^l n_{w,r}^l\right] = 0$ for $v \neq w$ and $q \neq r$.

As the same spreading code is used in all sub-channels, in order to optimally detect the spreading code, the output of the matched filters in all N_c sub-channels are observed: $\underline{z}^l = (z_{1,1}^l, \cdots, z_{v,q}^l, \cdots, z_{N_c,Q}^l)$. In the following sections, for notational simplicity, we drop the signaling interval index l whenever possible.

4.3 Turbo Receiver

The detection in the receiver is carried out by iteratively exchanging soft information about the coded bits between two sections: a SISO detector specifically designed for MC-SS-PB systems and a SISO channel decoder. The block diagram of the receiver is shown in Figure 4.2. Using similar receiver structure of single carrier SS-PB systems explained in Chapter 3, the SISO detector observes the matched filter outputs of all N_c subcarriers and by considering the *extrinsic* information provided by the SISO channel decoder in the previous iteration, delivers the *a posteriori* LLR of a transmitted “+1”

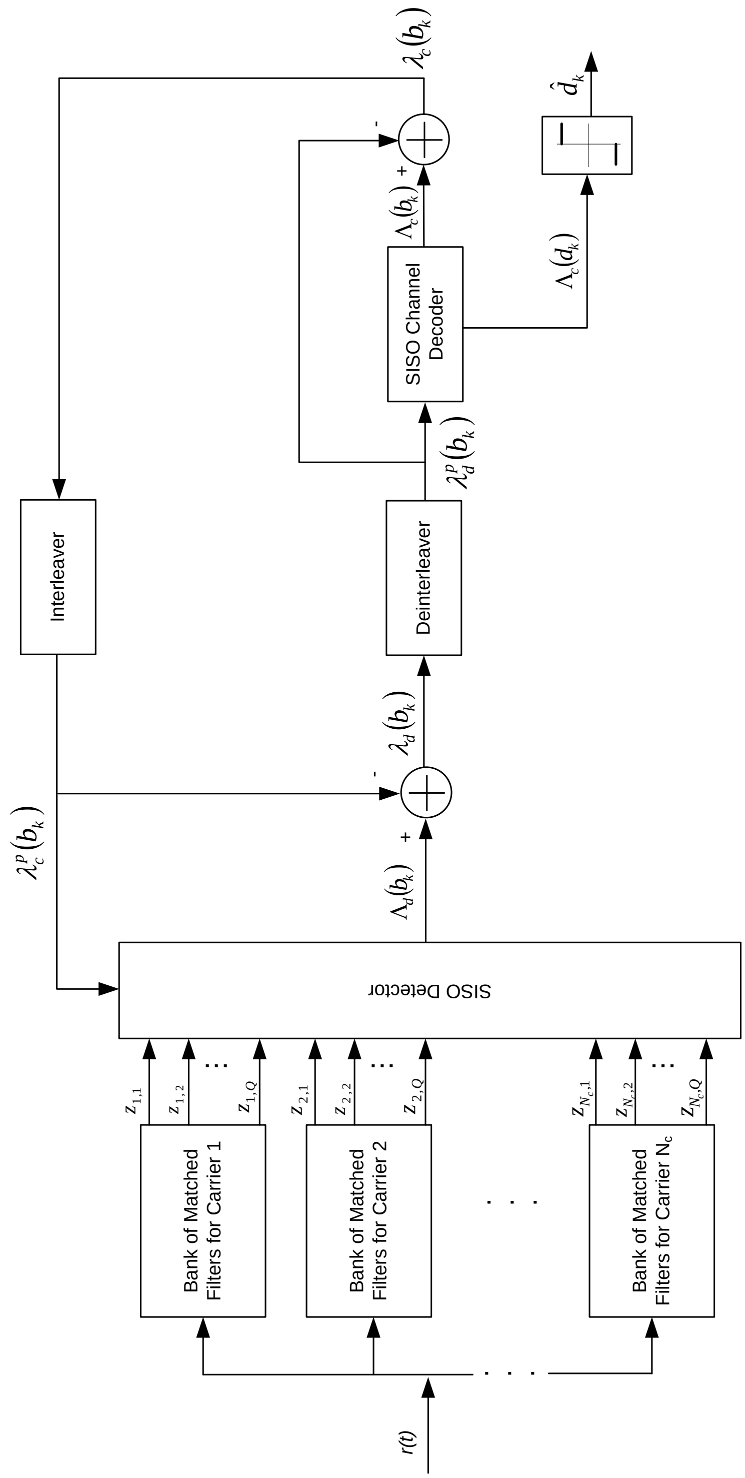

Figure 4.2: Turbo receiver for MC-SS-PB systems.

and a transmitted "−1" for the coded bit in each subcarrier,

$$\Lambda_d(b_k) \triangleq \ln \frac{P(b_k = +1|r(t))}{P(b_k = -1|r(t))}, \quad k = 1, 2, \cdots, N_c. \tag{4.4}$$

Equation (4.4) can be rewritten using Bayes' rule as

$$\Lambda_d(b_k) = \underbrace{\ln \frac{P(r(t)|b_k = +1)}{P(r(t)|b_k = -1)}}_{\lambda_d(b_k)} + \underbrace{\ln \frac{P(b_k = +1)}{P(b_k = -1)}}_{\lambda_c^p(b_k)}. \tag{4.5}$$

In this equation, $\lambda_c^p(b_k)$, represents the *a priori* LLR of the coded bit b_k calculated by the SISO channel decoder in the previous iteration. For the first round of detection (i.e. first iteration) when there is no prior information available about the likelihood of coded bits, we assume all the coded bits can be "+1" or "−1" equiprobably. In other words $\lambda_c^p(b_k) = 0$ for $k = 1, 2, \cdots, N_c$.

The *extrinsic* information provided by the SISO detector about the coded bit of the kth sub-channel, b_k, is denoted by $\lambda_d(b_k)$ in (4.5). This *extrinsic* information is calculated by the SISO detector based on the received signal $r(t)$ (or in other words the matched filter outputs of all sub-channels), the structure of the MC-SS-PB system and the prior information provided by the SISO channel decoder in the previous iteration about the coded bits other than the kth sub-channel bit, $\{\lambda_c^p(b_j), j \neq k\}$. This *extrinsic* information is then passed through a deinterleaver and feed into the SISO channel decoder as *a priori* information in the next iteration.

Similar to the discussion of Chapter 3, the SISO decoder calculates the *a posteriori* LLR of each coded bit by using the LLRs provided by the SISO detector and also considering the trellis structure of the convolutional encoder.

Now we focus our attention on the algorithm the SISO detector uses to calculate the *extrinsic* information for each detected bit.

Let us define

$$\mathcal{B}_{+}^{k} \triangleq \{\underline{b} \in (+1,-1)^{N_c}, b_k = +1\}, \tag{4.6}$$

that includes all the N_c coded bit vectors whose kth bit is "+1". Similarly $\mathcal{B}_{-}^{k}$ is defined as

$$\mathcal{B}_{-}^{k} \triangleq \{\underline{b} \in (+1,-1)^{N_c}, b_k = -1\}. \tag{4.7}$$

The *extrinsic* information delivered by the SISO detector $\lambda_d(b_k)$ can be rewritten as

$$\begin{aligned} \lambda_d(b_k) &= \ln \frac{P(r(t)|b_k = +1)}{P(r(t)|b_k = -1)} \\ &= \ln \frac{\sum_{\underline{b} \in \mathcal{B}_{+}^{k}} P(\underline{z}|\underline{b}) \prod_{j \neq k} P(b_j)}{\sum_{\underline{b} \in \mathcal{B}_{-}^{k}} P(\underline{z}|\underline{b}) \prod_{j \neq k} P(b_j)} \end{aligned} \tag{4.8}$$

where $P(b_j)$ is the probability that the bit in the jth subcarrier equals to b_j for $b_j \in \{+1,-1\}$. Equation (4.8) holds true because the received signal is fully represented by the outputs of the filters matched to Q orthogonal spreading sequences. It is also necessary and sufficient to observe the matched filter outputs of all N_c sub-channels to be able to completely extract the extrinsic information of each coded bit.

Let us define $\underline{y}$ as a noise and fading free Q matched filter outputs of all N_c sub-channels when $\underline{b}$ is sent and let w be the index of the spreading code selected by $\underline{b}$. In other words,

$$\underline{y} = (y_{1,1}, \cdots, y_{v,q}, \cdots, y_{N_c,Q}) \quad \text{in which} \quad y_{v,q} = \begin{cases} b_v & \text{if} \quad q = w \\ 0 & \text{otherwise} \end{cases} \tag{4.9}$$

where $y_{v,q}$ is the qth noise and fading free matched filter output of the vth sub-channel. From (4.3) since the noise has a normal distribution and assuming a coherent detection with perfect channel estimation, (4.8) can be rewritten as

$$\lambda_d\left(b_k\right) = \ln \frac{\sum_{\underline{b}\in\mathcal{B}_+^k} \exp\left(-\frac{1}{N_0} d(\underline{b})^2\right) \prod_{j\neq k} P(b_j)}{\sum_{\underline{b}\in\mathcal{B}_-^k} \exp\left(-\frac{1}{N_0} d(\underline{b})^2\right) \prod_{j\neq k} P(b_j)} \tag{4.10}$$

where

$$d(\underline{b})^2 = \sum_{i=1}^{N_c} \sum_{j=1}^{Q} \left|z_{i,j} - \alpha_i y_{i,j}\right|^2 \tag{4.11}$$

in which α_i is the complex fading gain that affects the ith sub-channel and $|.|$ denotes the magnitude.

Using (4.9), we can simplify (4.11) as

$$d(\underline{b})^2 = \sum_{i=1}^{N_c} \sum_{j=1}^{Q} \left|z_{i,j}\right|^2 + \sum_{i=1}^{N_c} \left|\alpha_i y_{i,w}\right|^2 - \sum_{i=1}^{N_c} (\alpha_i^* y_{i,w}^* z_{i,w} + \alpha_i y_{i,w} z_{i,w}^*) \tag{4.12}$$

where the superscript $*$ denotes the complex conjugate.The second term in (4.12) can be written as

$$\sum_{i=1}^{N_c} \left|\alpha_i y_{i,w}\right|^2 = \sum_{i=1}^{N_c} \left|\alpha_i b_i\right|^2 = \sum_{i=1}^{N_c} \left|\alpha_i\right|^2 \tag{4.13}$$

which is the summation of squared fading amplitudes of all N_c sub-channels and is independent of $\underline{b}$. The third term in (4.12) denoted by $e(\underline{b})$ can also be simplified as

$$e(\underline{b}) = -\sum_{i=1}^{N_c} (\alpha_i^* y_{i,w}^* z_{i,w} + \alpha_i y_{i,w} z_{i,w}^*) = -\sum_{i=1}^{N_c} (\alpha_i^* b_i z_{i,w} + \alpha_i b_i z_{i,w}^*). \tag{4.14}$$

Note that the first and the second terms in (4.12) are independent of $\underline{b}$ and therefore will be canceled in (4.10). Using the derivation of (3.16), the *a priori* probabilities in (4.8) can be calculated based on their LLR's $\lambda_c^p\left(b_k\right)$ as follows

$$P(b_j) = \frac{1}{2}\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p\left(b_j\right)\right)\right]. \tag{4.15}$$

Using (4.12), (4.14) and (4.15) in (4.10), we obtain

$$\lambda_d\left(b_k\right) = \ln \frac{\sum_{\underline{b}\in\mathcal{B}_+^k} \exp\left(-\frac{1}{N_0}e(\underline{b})\right)\prod_{j\neq k}\left[1+b_j\tanh\left(\frac{1}{2}\lambda_c^p\left(b_j\right)\right)\right]}{\sum_{\underline{b}\in\mathcal{B}_-^k} \exp\left(-\frac{1}{N_0}e(\underline{b})\right)\prod_{j\neq k}\left[1+b_j\tanh\left(\frac{1}{2}\lambda_c^p\left(b_j\right)\right)\right]}. \tag{4.16}$$

Now that we have (4.16), we can summarize the algorithm of the SISO detector in finding the *extrinsic* information of each coded bit as follows:

- First $\mathcal{B}_+^k$ and $\mathcal{B}_-^k$ are formed by dividing all the vectors of N_c coded bits, $\underline{b}$, transmitted in all sub-channels into the vectors in which the kth sub-channel transmits "+1" and "−1" respectively.

- Based on the code structure used in the transmitter for parity bit calculation, the spreading code index w is determined for each vector $\underline{b}$.

- $e(\underline{b})$ is calculated for each vector $\underline{b}$ using (4.14) that uses the wth matched filter output of all N_c sub-channels in addition to the fading gain amplitudes in all sub-channels. (If the fading amplitudes are known.)

- By knowing $e(\underline{b})$ and $\lambda_c^p\left(b_k\right)$ for each coded bits, the *extrinsic* LLR in the output of the SISO detector is computed by (4.16).

It can be seen that this algorithm is similar to the algorithm explained in Chapter 3 for the SISO detector of the SS-PB systems with some minor differences. When the *extrinsic* LLRs, $\lambda_d\left(b_k\right)$, are calculated by the SISO detector, they are passed through a deinterleaver and fed into the SISO channel decoder as the *a priori* information. In the last iteration, *a posteriori* LLRs provided by the SISO channel decoder about the data bits are used to make decision on data bits, in other words the estimated data bit, $\hat{d}$, is the sign of $\lambda_c\left(d_k\right)$ in the last iteration.

4.4 Simulation Results

The simulations are performed for coded MC-SS-PB systems that are based on the linear $(7, 4)$, $(10, 6)$ and $(11, 7)$ block codes. In the transmitter, at first, data bits are encoded by a 1/2 rate convolutional encoder with the constraint length of 5 and the generators of 23 and 35 in octal notation. The coded bits are passed through a random interleaver before being fed into the MC-SS-PB modulator. The simulations are performed considering two different scenarios. First when the receiver does not have information about the fading gain in each sub-channel and second when the receiver is able to perfectly estimate the fading gains of all subcarriers. In both scenarios coherent detection is assumed. The cross-correlation between all spreading codes is assumed to be zero (orthogonal codes) and the channel is a slowly-varying, frequency selective, Rayleigh channel, where the DS-SS signals in each frequency band is fading nonselectively and independently.

In the first scenario where the receiver is unable to estimate the fading gains, the channel gains are assumed to be equal to one in the calculation of *extrinsic* information of the SISO detector. In other words, $\alpha_i = 1$ for all i. The BER performance of the proposed turbo receiver when the channel gains are not available for the coded MC-SS-PB system based on $(7, 4)$, $(10, 6)$ and $(11, 7)$ block encoder along with the conventional coded MC-DS-SS system are shown in Figures 4.3, 4.4 and 4.5 respectively.

The simulation results show that even when the channel gains are not known, the coded MC-SS-PB system in the first iteration outperforms the conventional DS-SS system higher E_b/N_0s (where E_b denotes the bit energy). Significant performance improvement is achieved in the second, third and fourth iteration of detection and decoding.

The BER performance of the proposed turbo receiver of the coded MC-SS-PB system based on $(7, 4)$, $(10, 6)$ and $(11, 7)$ codes when the channel gains of all sub-channels are perfectly known by the receiver are shown in Figures 4.6, 4.7 and 4.8 respectively. These

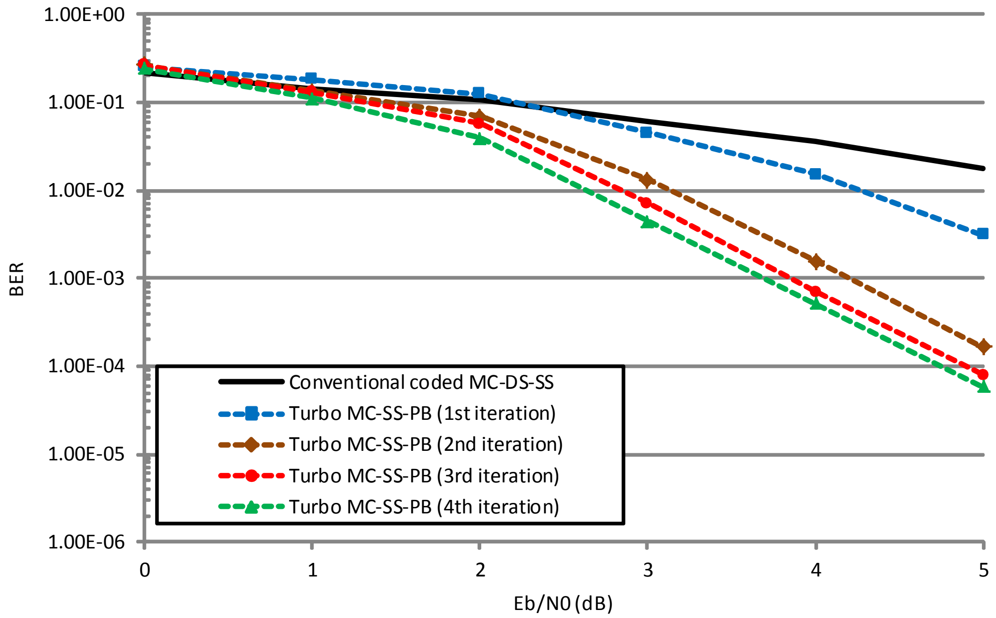

Figure 4.3: BER Performance of the proposed turbo receiver for the coded $(7, 4)$ based MC-SS-PB system with unknown fading gains.

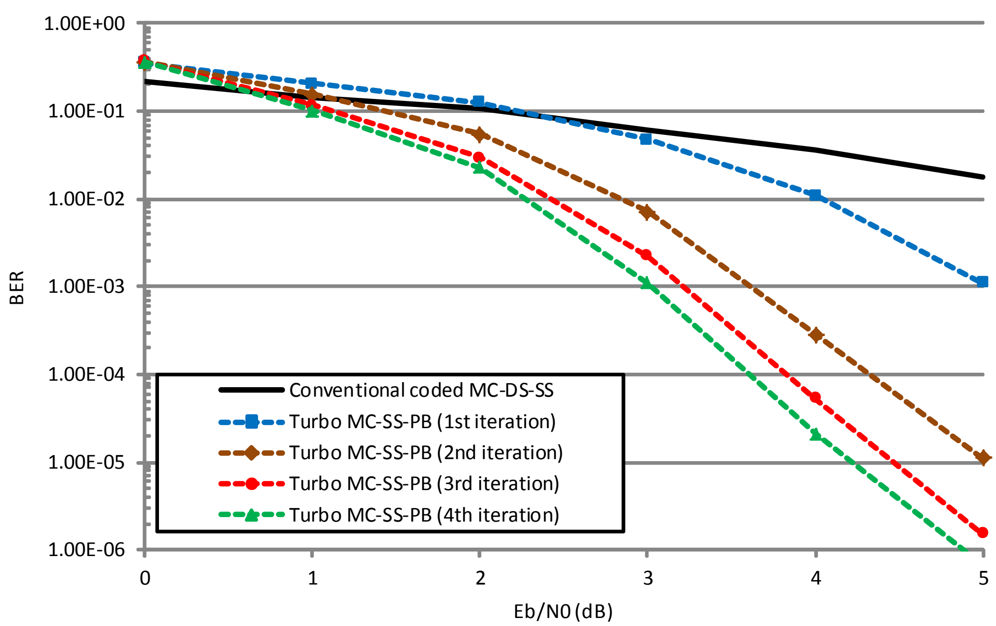

Figure 4.4: BER Performance of the proposed turbo receiver for the coded $(10, 6)$ based MC-SS-PB system with unknown fading gains.

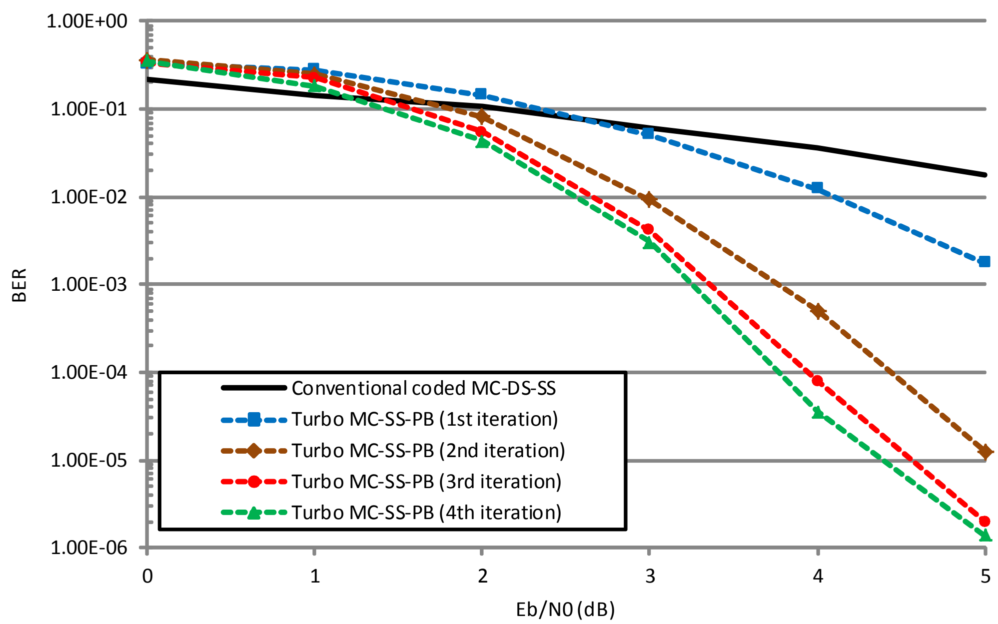

Figure 4.5: BER Performance of the proposed turbo receiver for the coded $(11, 7)$ based MC-SS-PB system with unknown fading gains.

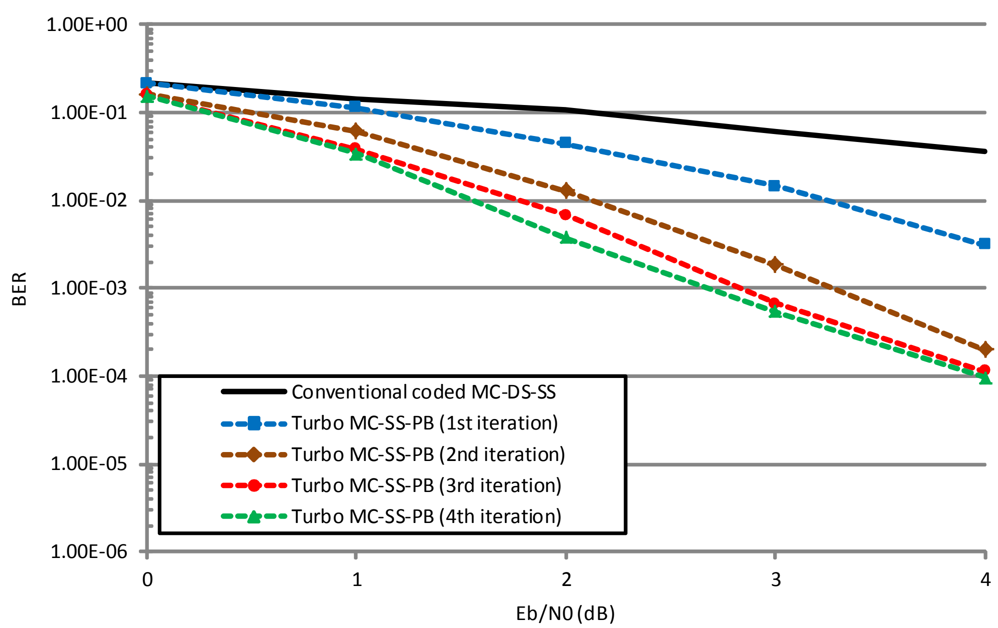

Figure 4.6: BER Performance of the proposed turbo receiver for the coded $(7, 4)$ based MC-SS-PB system with known fading gains.

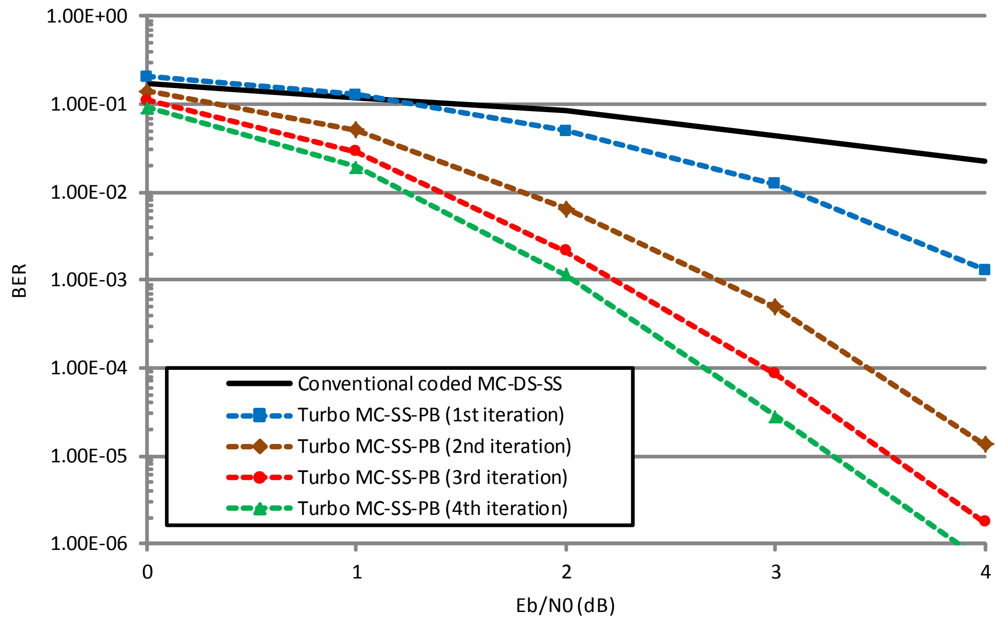

Figure 4.7: BER Performance of the proposed turbo receiver for the coded (10, 6) based MC-SS-PB system with known fading gains.

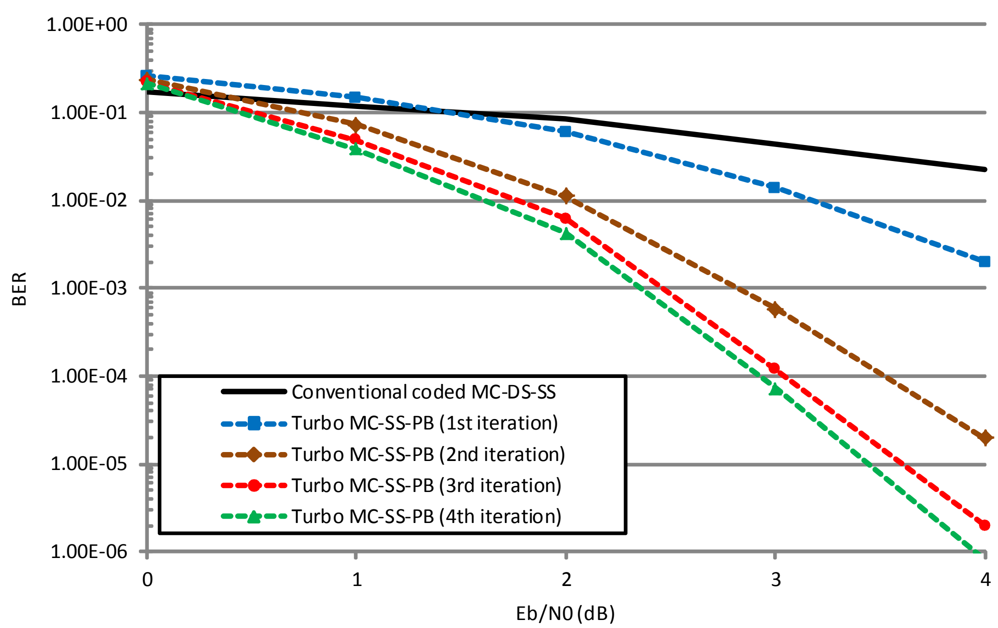

Figure 4.8: BER Performance of the proposed turbo receiver for the coded (11, 7) based MC-SS-PB system with known fading gains.

figures show a significant performance improvement when turbo detection and decoding is performed in the receiver. It can be seen in Figure 4.7 that the performance of the system in the fourth iteration at 10^{-2} achieves more than 3 dB gain over the conventional system. Comparing Figures 4.7 and 4.4 reveals that knowing the fading gains in all sub-channels provide around 1 dB gain at a BER of 10^{-2} in the fourth iteration. It can also be seen that more coding gain is achievable in lower BERs.

The improved BER in MC-SS-PB systems is achieved by creating dependency between the bits carried on different carriers. In other words the information bits transmitted on the sub-bands encountering deep fades can still be recovered with high probability because of the relationship between the spreading codes used and the information it carries.

4.5 Discussion

In this chapter, we have introduced a novel turbo receiver for MC-SS-PB systems. An algorithm for the SISO detector in MC-SS-PB systems is developed in which the LLRs of each bit is calculated based on the received signal, the *a priori* information from the previous iteration and the code structure of the MC-SS-PB system. The proposed turbo receiver consists of a SISO detector for MC-SS-PB systems followed by a SISO channel decoder. Exchange of LLRs between these two components provides a significant performance improvement compared to the conventional coded MC-SS systems. If the MC-SS-PB system is used in a multiuser scenario, the improved BER performance of the proposed system can be traded off against additional users in a CDMA system.

Chapter 5

Turbo Receiver for MIMO-CDMA Systems Employing Parity Bit Selected and Permutation Spreading

5.1 Introduction

In multiple input multiple output (MIMO) systems where multiple antennas are used at both ends of the wireless channels, spectacular benefits are achieved by taking advantage of spatial diversity. By using space-time codes in MIMO systems, a trade-off between spatial multiplexing and diversity order can be achieved [45]. When CDMA is combined with MIMO systems (MIMO-CDMA), orthogonal spreading sequences can be used in each transmit antenna in order to allow the receiver to accurately estimate the channel gains which are very crucial in the performance of MIMO systems.

The extension of parity bit selected spreading technique to MIMO-CDMA systems is introduced in [7] and is explained in Chapter 2 where two main techniques to select spreading waveforms are explained. In a MIMO-CDMA system employing parity bit selected spreading code, the calculated parity bits are used to choose a spreading waveform from a set of mutually orthogonal spreading codes to be used in all transmit antenna during that signaling interval. In contrast, in a system employing permutation spreading, each block of information bits, assigns a different permutation of spreading waveforms to transmit antennas. In [46] and [47] the soft output detection for MIMO-CDMA systems that employ parity bit selected and permutation spreading is proposed. In the proposed systems of [46] and [47], information bits are first convolutionally encoded prior to being used in a MIMO-CDMA system. Reliability of the detected bits in the receiver is calculated in terms of LLRs and then used as the input to a soft input Viterbi decoder. It is shown in [46] and [47] that when a convolutional encoder is concatenated with a MIMO-CDMA system that employs either parity bit selected or permutation spreading, a significant BER improvement is achieved even without iteratively exchanging the soft information.

The application of turbo processing in MIMO systems has been discussed in different literatures and has been shown that they can attain a channel capacity close to Shannon limit [48], [49]. In this chapter we introduce a new turbo receiver for MIMO-CDMA systems that employ parity bit selected and permutation spreading. Similar to the systems introduced in chapters 3 and 4, a serially concatenated convolutional code is introduced into the transmitter and the parity bit selected or permutation spreading techniques act as an inner code. In the receiver, the soft information about the coded bits are iteratively exchanged between a SISO detector and a SISO decoder. Simulation results presented in this chapter show a significant improvement in the performance of MIMO-CDMA systems of both parity bit selected and permutation spreading when a turbo receiver is implemented.

5.2 System Model

Consider a coded MIMO system with N_t transmit antennas and N_r receive antennas. The block diagram of the transmitter is shown in Figure 5.1. The binary information stream $\{d[i]\}$ is convolutionally encoded with code rate R and after passing through an interleaver, the stream of interleaved coded bits $\{m[j]\}$ is converted into N_t parallel streams. On the lth signaling interval, the vector of bits to be transmitted by N_t transmit antennas is $\underline{m}^l = (m_1^l, m_2^l, \cdots, m_{N_t}^l)$. Each message vector of $\underline{m}^l$ is used to select a set of spreading waveforms for all N_t transmit antennas, $\mathcal{W}^l = \{w_1^l(t), w_2^l(t), \cdots, w_{N_t}^l(t)\}$, from a set of Q mutually orthogonal spreading sequences: $\{c_q(t)\}$, $q = 1, 2, \cdots, Q$, where $w_i^l(t)$ is the spreading waveform used to spread the data transmitted by the ith antenna on time interval l. Data bits of all N_t transmit antennas are modulated using BPSK to obtain $\underline{b}^l = (b_1^l, b_2^l, \cdots, b_{N_t}^l)$. Using the selected set of spreading waveforms, $\mathcal{W}^l$, each modulated bit is spread before being transmitted by one of N_t transmit antennas.

5.2.1 MIMO-CDMA Systems Employing Parity Bit Selected Spreading

As discussed in Chapter 2, in SS-PB systems, unlike the conventional systems, a parity bit calculator of a systematic block code (n_0, k_0) are used to select one of a set of $2^{(n_0-k_0)}$ spreading sequences, where n_0 is the length of message vector and (n_0-k_0) is the number of parity bits. In this way, depending on the coding scheme, all possible message vectors are divided into groups that are spread by the same spreading waveform.

Let $\mathcal{M}$ be a set of all $M = 2^{N_t}$ possible message vectors of length N_t. We can partition $\mathcal{M}$ into Q cosets of $\mathcal{M}_1, \mathcal{M}_2, \cdots, \mathcal{M}_Q$ where each one has M/Q message vectors. Depending on which coset the message vector $\underline{m}^l$ comes from, a spreading waveform is

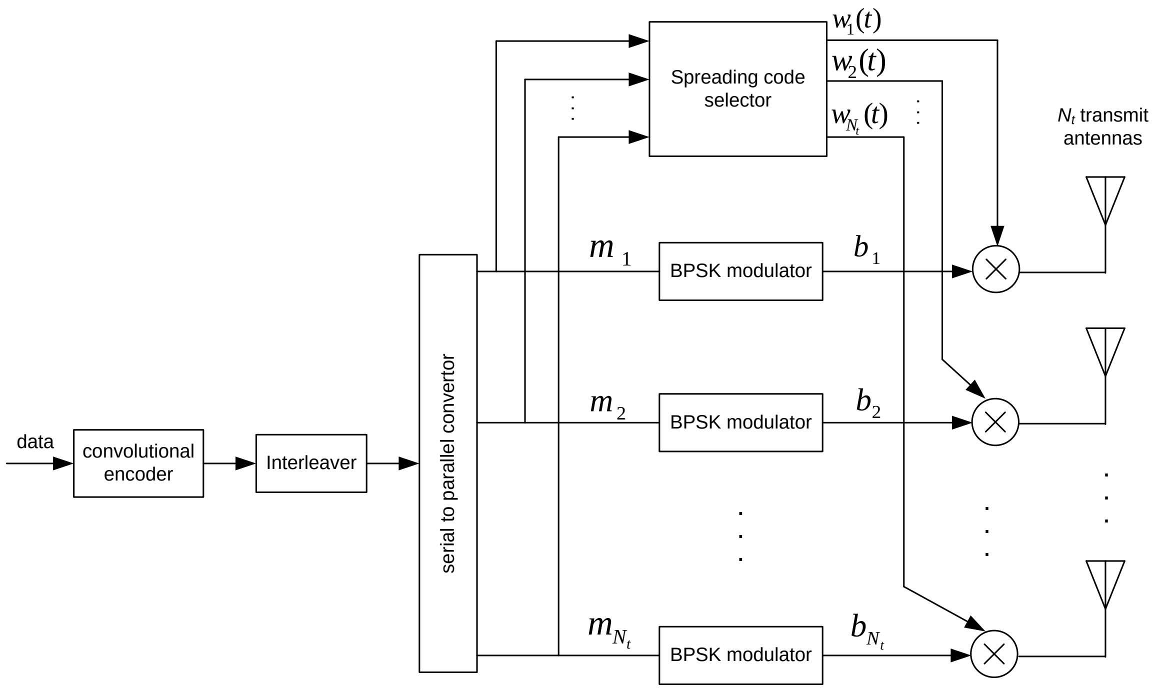

Figure 5.1: Transmitter of a coded MIMO-CDMA system that employs parity bit selected or permutation spreading.

assigned to all transmit antennas. In other words; if $\underline{m}^l \in \mathcal{M}_q$ then

$$w_1^l(t) = w_2^l(t) = \cdots = w_{N_t}^l(t) = c_q(t). \tag{5.1}$$

In this chapter we consider the same system explained in Chapter 2 where $Q = 8$ mutually orthogonal spreading waveforms are used in a system with $N_t = 4$ transmit antennas and the cosets are as in (2.13). Considering this system, if $\underline{m}^l = (0, 0, 1, 0)$, all 4 transmit antennas use $c_3(t)$ as the spreading waveform.

To be able to determine which spreading code is used in the transmitter, each receive antenna is equipped with Q matched filters, each matched to one of the spreading codes of set $\{c_q(t)\}$. The output of the qth matched filter of the jth receive antenna on the lth signaling interval is:

$$z_{j,q}^l = \begin{cases} \sqrt{\frac{E_b}{N_r}} \sum_{i=1}^{N_t} b_i^l \alpha_{ij}^l + n_{j,q}^l & \text{if } c_q(t) \text{ is used} \\ n_{j,q}^l & \text{otherwise} \end{cases} \quad (5.2)$$

where α_{ij}^l is the complex channel gain on the link between transmit antenna i and receive antenna j on lth signaling interval; E_b is the averaged received energy per bit; and $n_{j,q}^l$ is the sampled noise from the qth matched filter of the jth receive antenna which is a zero mean Gaussian random variable with variance $N_0/2$.

5.2.2 MIMO-CDMA Systems Employing Permutation Spreading

Contrary to the MIMO-CDMA system that employs parity bit selected spreading, in MIMO-CDMA system employing permutation spreading, depending on the coset the message comes from, different permutation of spreading waveforms are assigned to transmit antennas. In this section we consider the system discussed in Chapter 2 with spreading permutations listed in Table 2.1. In this system, $Q = 8$ orthogonal codes are available for a system with $N_t = 4$ transmit antennas. The qth matched filter output of the j receive antenna is

$$z_{j,q}^l = \begin{cases} b_i^l \sqrt{\frac{E_b}{N_r}} \alpha_{ij}^l + n_{j,q}^l & \text{if } w_i^l(t) = c_q(t) \\ n_{j,q}^l & \text{otherwise} \end{cases} \quad (5.3)$$

In order to optimally determine the spreading waveforms used in the transmitter, the matched filter outputs of all N_r receive antennas are observed: $\underline{z}^l = (z_{1,1}^l, \cdots, z_{j,q}^l, \cdots, z_{N_r,Q}^l)$. Let us define $\underline{y}_{(\underline{b})}^l$ as a noise free matched filter outputs of all N_r receive antennas when

$\underline{b}^l$ is sent. Then we can write:

$$\underline{z}^l = \underline{y}^l_{(\underline{b})} + \underline{n}^l \tag{5.4}$$

where $\underline{n}^l = (n^l_{1,1}, \cdots, n^l_{j,q}, \cdots, n^l_{N_r,Q})$.

For example for a MIMO-CDMA system with $N_t = N_r = 4$ that employs parity bit selected spreading using the cosets of (2.13), if we consider $\underline{m}^l = (0, 0, 1, 0)$ from coset $\mathcal{M}_3$, then $\underline{b}^l = (-1, -1, +1, -1)$ and

$$\underline{y}^l_{(\underline{b})} = \sqrt{\frac{E_b}{N_r}}(0, 0, -\alpha^l_{11} - \alpha^l_{21} + \alpha^l_{31} - \alpha^l_{41}, 0, 0, 0, 0, 0, 0, 0, -\alpha^l_{12} - \alpha^l_{22} + \alpha^l_{32} - \alpha^l_{42}, 0, 0, 0, 0, 0, \\ 0, 0, -\alpha^l_{13} - \alpha^l_{23} + \alpha^l_{33} - \alpha^l_{43}, 0, 0, 0, 0, 0, 0, 0, -\alpha^l_{14} - \alpha_{24} + \alpha^l_{34} - \alpha^l_{44}, 0, 0, 0, 0, 0).$$

Likewise, for the system employing permutation spreading shown in Table 2.1, if the message vector is $\underline{m}^l = (0, 0, 0, 0)$ from coset M_1, then $\underline{b}^l = (-1, -1, -1, -1)$ and

$$\underline{y}^l_{(\underline{b})} = \sqrt{\frac{E_b}{N_r}}(- \alpha^l_{11}, 0, -\alpha^l_{21}, 0, -\alpha^l_{31}, 0, -\alpha^l_{41}, 0, -\alpha^l_{12}, 0, -\alpha^l_{22}, 0, -\alpha^l_{32}, 0, -\alpha^l_{42}, 0, \\ - \alpha^l_{13}, 0, -\alpha^l_{23}, 0, -\alpha^l_{33}, 0, -\alpha^l_{43}, 0, -\alpha^l_{14}, 0, -\alpha^l_{24}, 0, -\alpha^l_{34}, 0, -\alpha^l_{44}, 0).$$

In the following sections, for notational simplicity, we drop the signaling interval index l whenever possible.

5.3 Turbo Receiver

The proposed turbo receiver is shown in Figure 5.2. Similar to the structure of other turbo-based receivers, it consists of two main parts: a SISO detector specifically designed for MIMO-CDMA systems that employ parity bit selected and permutation spreading and a SISO channel decoder. These two parts are connected via an interleaver and a deinterleaver. Similar to the turbo receivers of SS-PB and MC-SS-PB systems discussed

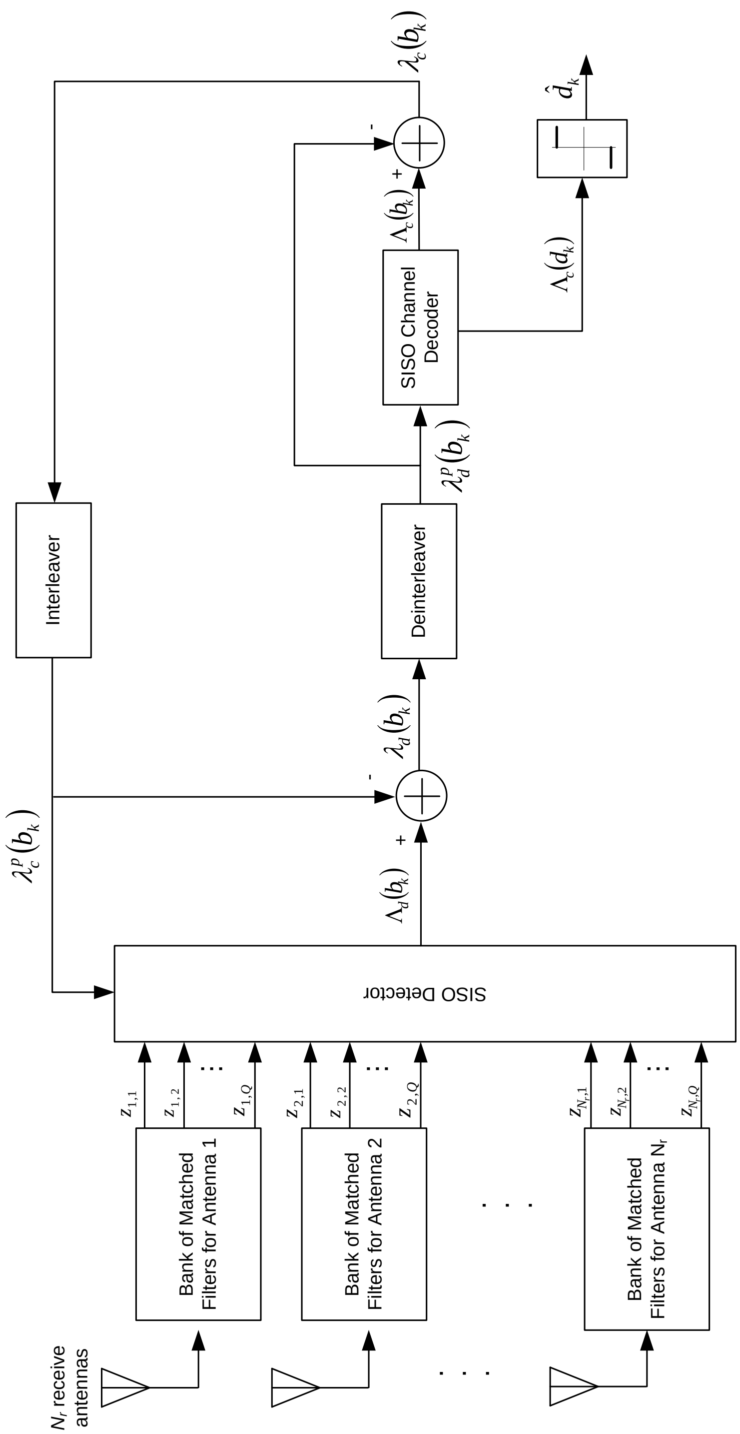

Figure 5.2: Turbo receiver for MIMO-CDMA systems that employ parity bit selected or permutation spreading.

in Chapters 3 and 4, the SISO detector observes the matched filter outputs of all N_r antennas and by considering the *extrinsic* information provided by the SISO channel decoder in the previous iteration, delivers the *a posteriori* LLR of a transmitted "+1" and a transmitted "−1" for the coded bit in each transmit antenna,

$$\Lambda_d\left(b_k\right) \triangleq \ln \frac{P\left(b_k=+1|\underline{z}\right)}{P\left(b_k=-1|\underline{z}\right)}, \qquad k=1,2,\cdots,N_t. \tag{5.5}$$

Using Bayes' rule, we can rewrite (5.5) as

$$\Lambda_d\left(b_k\right) = \underbrace{\ln \frac{P\left(\underline{z}|b_k=+1\right)}{P\left(\underline{z}|b_k=-1\right)}}_{\lambda_d\left(b_k\right)} + \underbrace{\ln \frac{P\left(b_k=+1\right)}{P\left(b_k=-1\right)}}_{\lambda_c^p\left(b_k\right)}. \tag{5.6}$$

The *a priori* LLR of each coded bit (shown by $\lambda_c^p\left(b_k\right)$ in (5.6)) is calculated by the SISO channel decoder. These LLRs are assumed to be zero in the first iteration when there is no prior information available about the likelihood of the coded bits. In other words $\lambda_c^p\left(b_k\right)=0$ for $k=1,2,\cdots,N_r$.

The SISO detector calculates the *extrinsic* information about each coded bit (shown by $\lambda_d\left(b_k\right)$ in (5.6)) by considering the matched filter outputs, $\underline{z}$, the spreading code selection strategy of the MIMO-CDMA system and the prior information provided by the SISO channel decoder in the previous iteration. This *extrinsic* information is then passed through a deinterleaver and feed into the SISO channel decoder as *a priori* information in the next iteration.

Similar to the discussion of Chapter 3, the SISO decoder gets the *a priori* LLRs from the SISO detector and by considering the trellis structure of the convolutional code, calculates the *a posteriori* LLRs of each coded and also data bit, namely $\Lambda_c\left(b_k\right)$ and $\Lambda_c\left(d_k\right)$ respectively.

Now we concentrate on the algorithm used by the SISO detector that is specifically designed for the MIMO-CDMA systems that employ parity bit selected or permutation spreading. The SISO detector observes the matched filter outputs along with the prior information from the channel decoder in the previous iteration to calculate the *extrinsic* information for each detected bit.

Let us define

$$\mathcal{B}_{+}^{k} \triangleq \{\underline{b} \in (+1,-1)^{N_t}, b_k = +1\}, \tag{5.7}$$

that includes all the N_t coded bit vectors whose kth bit is "+1". Similarly $\mathcal{B}_{-}^{k}$ is defined as

$$\mathcal{B}_{-}^{k} \triangleq \{\underline{b} \in (+1,-1)^{N_t}, b_k = -1\}. \tag{5.8}$$

The *extrinsic* information delivered by the SISO detector $\lambda_d(b_k)$ can be rewritten as

$$\begin{aligned}\lambda_d(b_k) &= \ln \frac{P(\underline{z}|b_k = +1)}{P(\underline{z}|b_k = -1)} \\ &= \ln \frac{\sum_{\underline{b}\in\mathcal{B}_{+}^{k}} P(\underline{z}|\underline{b}) \prod_{j\neq k} P(b_j)}{\sum_{\underline{b}\in\mathcal{B}_{-}^{k}} P(\underline{z}|\underline{b}) \prod_{j\neq k} P(b_j)}\end{aligned} \tag{5.9}$$

where $P(b_j)$ is the probability that the bit transmitted by the jth antenna equals to b_j for $b_j \in \{+1,-1\}$. Equation (5.9) holds true because the received signal is fully represented by the outputs of the filters matched to Q orthogonal spreading sequences. It is also necessary and sufficient to observe the matched filter outputs of all N_t transmit antennas to be able to completely extract the extrinsic information of each coded bit.

From (5.2) and (5.3), since the noise has a normal distribution and assuming that the receiver has a perfect knowledge about all channel gains, (5.9) can be rewritten as

$$\lambda_d(b_k) = \ln \frac{\sum_{\underline{b}\in\mathcal{B}_{+}^{k}} \exp\left(-\frac{1}{N_0} d(\underline{b})^2\right) \prod_{j\neq k} P(b_j)}{\sum_{\underline{b}\in\mathcal{B}_{-}^{k}} \exp\left(-\frac{1}{N_0} d(\underline{b})^2\right) \prod_{j\neq k} P(b_j)} \tag{5.10}$$

where

$$d(\underline{b})^2 = ||\underline{z} - \underline{y}_{(\underline{b})}||^2 \tag{5.11}$$

is the squared Euclidean distance between the vector of received and noise free matched filter outputs when $\underline{b}$ is sent.

From (3.16), the *a priori* probabilities in (5.9) can be calculated based on their LLR's $\lambda_c^p(b_k)$ as

$$P(b_j) = \frac{1}{2}\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p(b_j)\right)\right]. \tag{5.12}$$

Substituting (5.12) in (5.10), we obtain

$$\lambda_d(b_k) = \ln \frac{\sum_{\underline{b}\in\mathcal{B}_+^k} \exp\left(-\frac{1}{N_0}d(\underline{b})^2\right)\prod_{j\neq k}\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p(b_j)\right)\right]}{\sum_{\underline{b}\in\mathcal{B}_-^k} \exp\left(-\frac{1}{N_0}d(\underline{b})^2\right)\prod_{j\neq k}\left[1 + b_j \tanh\left(\frac{1}{2}\lambda_c^p(b_j)\right)\right]}. \tag{5.13}$$

By having (5.13), the algorithm of the SISO detector in computing the *extrinsic* information of each coded bit can be summarized as follows:

- First $\mathcal{B}_+^k$ and $\mathcal{B}_-^k$ are formed by dividing all the vectors of N_t coded bits, $\underline{b}$, transmitted by all antennas into the vectors in which the kth antenna transmits "+1" and "−1" respectively.
- Based on weather parity bit selected or permutation spreading technique is used in the transmitter, the expected noise free matched filter outputs of $\underline{y}_{(\underline{b})}$ is determined for each vector $\underline{b}$.
- The squared Euclidean distance, $d(\underline{b})^2$, is calculated for each vector $\underline{b}$ using (5.11).
- By knowing $d(\underline{b})^2$ and $\lambda_c^p(b_k)$ for each coded bits, the *extrinsic* LLR in the output of the SISO detector is computed by (5.13).

Similar to the discussion of Chapter 3 and 4, when the SISO detector calculates the *extrinsic* LLRs for each coded bit, $\lambda_d(b_k)$, they are passed through a deinterleaver and fed into the SISO channel decoder as the *a priori* information. The SISO channel decoder computes the *a posteriori* LLRs for both coded and decoded (data) bits. The *extrinsic* LLRs $\lambda_c(b_k)$ at the output of the channel decoder are formed by subtracting the *a priori* LLRs $\lambda_d^p(b_k)$ from the *a posteriori* LLRs of the coded bits $\Lambda_c(b_k)$. In the last iteration, *a posteriori* LLRs of the data bits are used to make decision on data bits, in other words the estimated data bit, $\hat{d}$, is the sign of $\lambda_c(d_k)$ in the last iteration.

5.4 Simulation Results

The simulations are performed for coded MIMO-CDMA systems that employ parity bit selected and permutation spreading.

In the transmitter, at first, data bits are encoded by a 1/2 rate convolutional encoder with the constraint length of 5 and the generators of 23 and 35 in octal notation. The coded bits are randomly interleaved and serial-to-parallel converted such that 4 parallel coded bits are transmitted simultaneously on 4 transmit antennas. Depending on the spreading code selection technique, a message vector of length 4 is used to choose spreading waveforms for transmit antennas. If parity bit selected technique is used, using (2.13), depending on the coset the message vector comes from, a spreading waveform is selected to be used by all transmit antennas. In the case of MIMO-CDMA system with permutation spreading, the message vector follows Table 2.1 to select permutation of spreading waveforms for each transmit antenna. The channel is modeled as a slowly-varying, frequency nonselective, Rayleigh fading channel, where the channel gains of different transmit-receive links are uncorrelated. We assume that perfect channel estimation is available at the receiver without a power penalty. We consider MIMO-CDMA

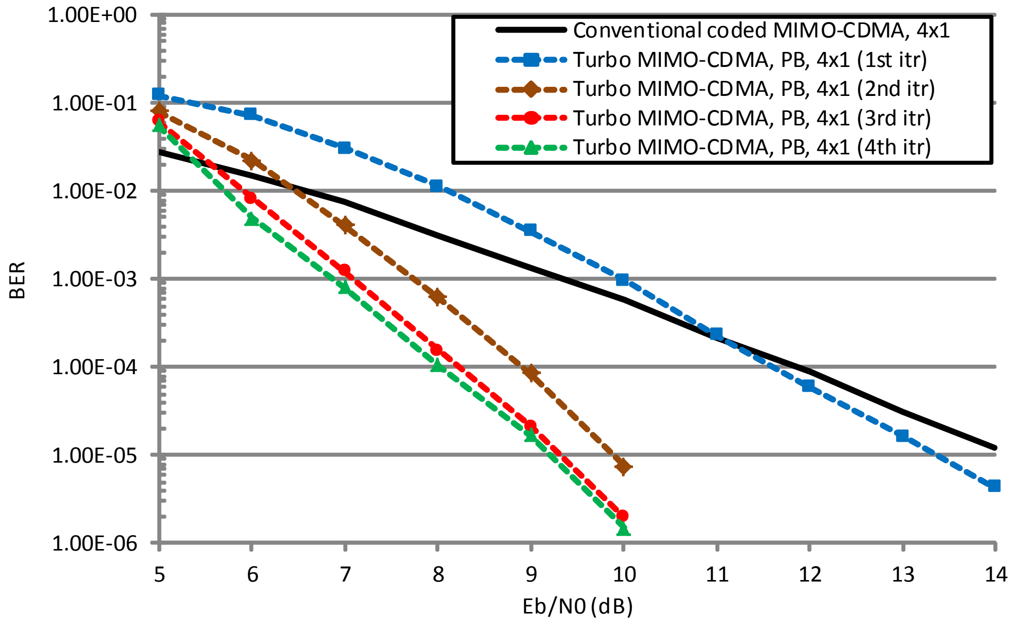

Figure 5.3: BER for the MIMO-CDMA system employing parity bit selected spreading with $Nt = 4$ and $Nr = 1$.

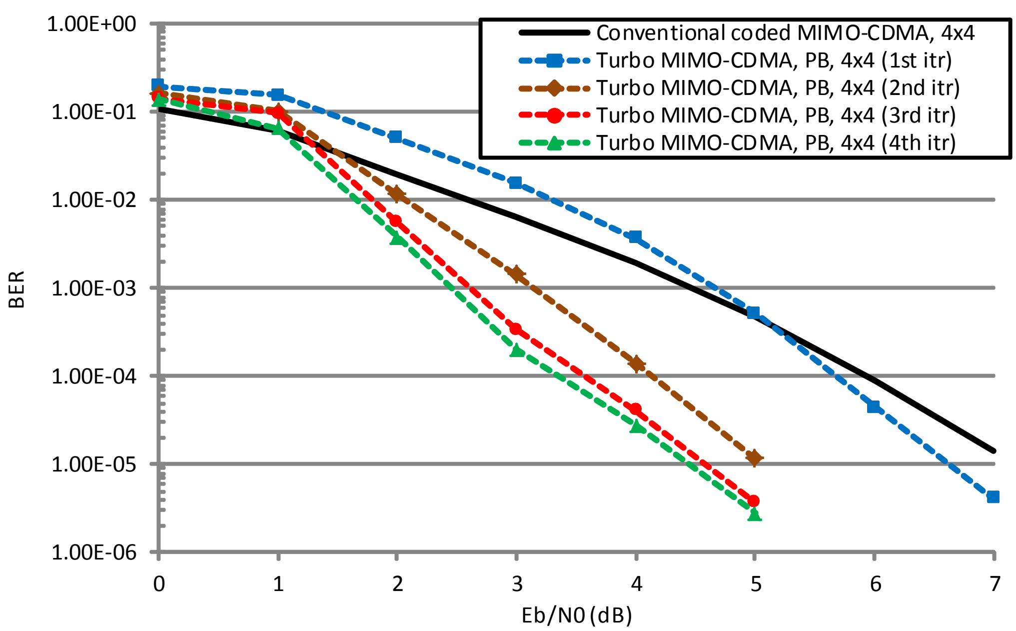

Figure 5.4: BER for the MIMO-CDMA system employing parity bit selected spreading with $Nt = 4$ and $Nr = 4$.

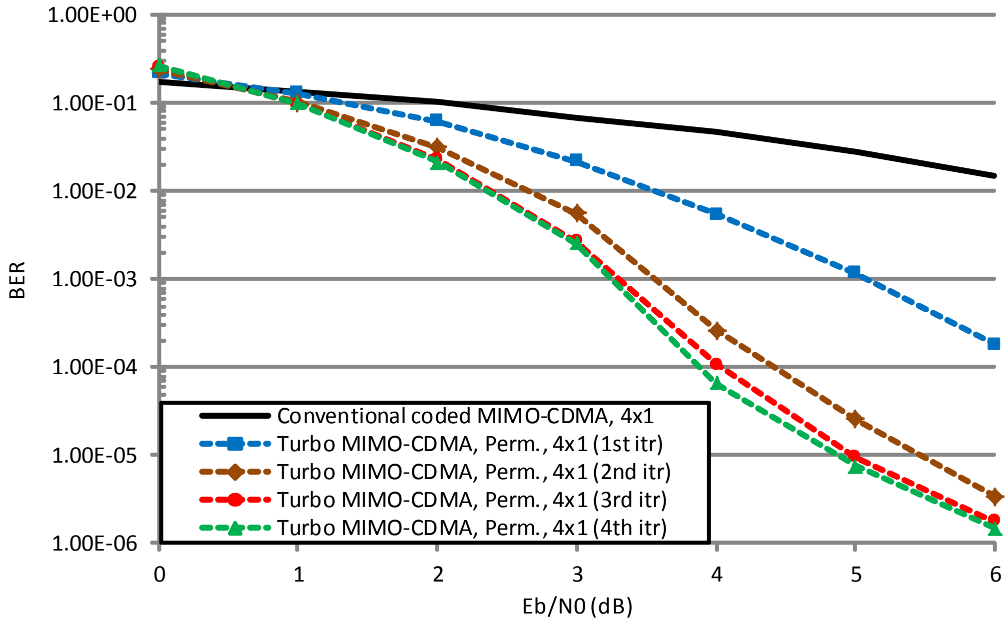

Figure 5.5: BER for the MIMO-CDMA system employing permutation spreading with $Nt = 4$ and $Nr = 1$.

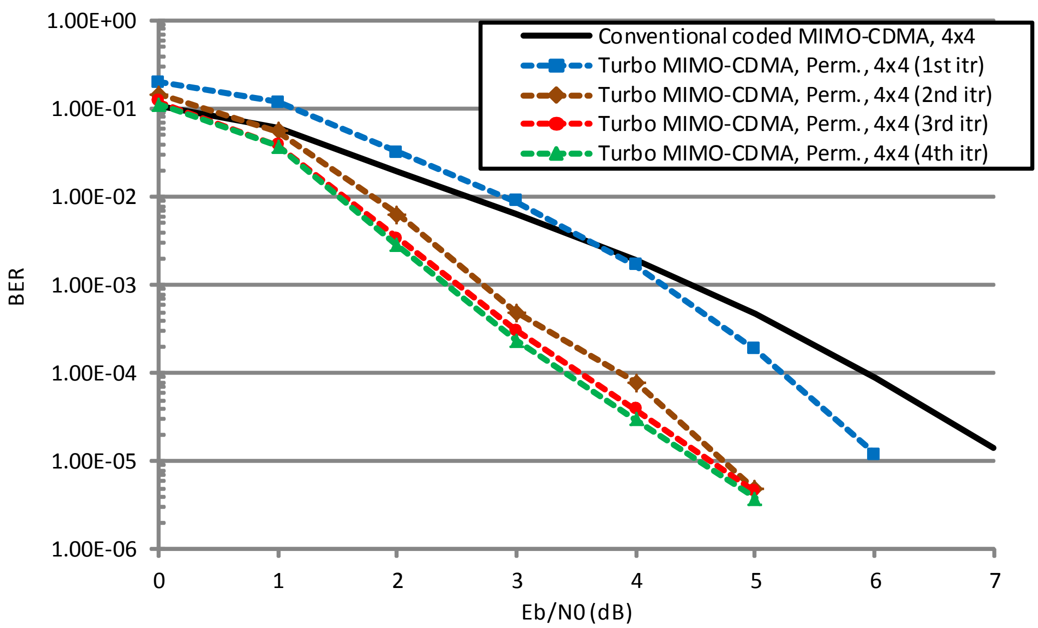

Figure 5.6: BER for the MIMO-CDMA system employing permutation spreading with $Nt = 4$ and $Nr = 4$.

systems with 4 transmit antennas and 4 or 1 receive antennas.

The BER performance of the proposed turbo receiver versus average E_b/N_0 for MIMO-CDMA systems employing parity bit selected spreading with 4 transmit antennas and 1 and 4 receive antennas along with the conventional coded MIMO-CDMA systems are shown in Figures 5.3 and 5.4 respectively. These results show that when parity bit selected spreading is used, the first iteration of the proposed turbo receiver outperforms the conventional coded MIMO-CDMA systems at E_b/N_0 of 11 dB and 5 dB for the systems with 4 and 1 receive antennas respectively. It can be seen that a significant performance improvement is achieved when more iterations are implemented at the receiver. At a BER of 10^{-3}, the fourth iteration of the proposed turbo receiver in MIMO-CDMA systems that employ parity bit selected spreading has around 3 dB gain over the conventional MIMO-CDMA with both 1 and 4 receive antennas.

Figures 5.5 and 5.6, respectively, present the BER performance of the proposed turbo receiver in MIMO-CDMA systems with 4 transmit antennas and 1 and 4 receive antennas when permutation spreading is used. Comparing the simulation results of systems with parity bit selected and permutation spreading shows that when permutation spreading is used, more significant BER performance gain is achieved. This is due to the nature of each spreading code selection technique. In parity bit selected spreading, since all transmit antennas use the same spreading waveform, one decision variable (i.e. matched filter output) per receive antenna contains a signal component which is a sum of all the transmitted signal. This may lead to some interference between the signals transmitted by different antennas that deteriorates the overall BER performance. In the case of permutation spreading, since in each signaling interval different spreading waveform is used by each antenna, there is no interference between the receive signals form each transmit antenna when the channel gains are independent.

The improved BER performance in MIMO-CDMA systems that employ either of two discussed spreading code selection techniques is achieved by creating dependency between the bits carried by different antennas. In other words the information bits transmitted by a transmit antenna encountering deep fades can still be recovered with high probability because of the relationship between the spreading codes used and the information it carries.

5.5 Discussion

In this chapter, we have introduced a novel turbo receiver for MIMO-CDMA systems that employ parity bit selected and permutation spreading. An algorithm for the SISO detector in such systems is developed in which the LLRs of each bit is calculated based on the received signal, the *a priori* information from the previous iteration and the code structure of the MIMO-CDMA system. The proposed turbo receiver consists of a SISO detector for MIMO-CDMA systems followed by a SISO channel decoder. Iteration of LLRs between these two components provides a significant performance improvement in compare to the conventional coded MIMO-CDMA systems. If the MIMO-CDMA system employing parity bit selected and permutation spreading is used in a multiuser scenario, the improved BER performance of the proposed system can be traded off against additional users in a CDMA system.

Chapter 6

Turbo Multiuser Receiver for CDMA Systems Employing Parity Bit Selected Spreading Sequences

6.1 Introduction

In all the previous chapters, just single user scenarios of spread spectrum systems that employ parity bit selected spreading sequences are discussed . In this chapter, we consider multiuser scenario of CDMA systems that employ parity bit selected spreading sequences (CDMA-PB). The performance of the spread spectrum systems in a multiuser environment is hugely affected by the multiple access interference (MAI) and hence, design of an effective multiuser detector (MUD) that can mitigate the effect of MAI is greatly needed.

In this chapter we propose a turbo multiuser receiver for CDMA systems that employ parity bit selected spreading sequences. Considering an uplink channel, data bits of every user are first convolutionally encoded and interleaved before being used to calculate the parity bits and then select the spreading codes in a SS-PB system. In the base station, a turbo multiuser receiver is implemented that consists of a SISO multiuser detector and a SISO channel decoder. The receiver implements a turbo processing by iteratively exchanging soft information on every coded bits between a SISO multiuser detector and a SISO channel decoder. In each iteration, the extrinsic information provided by the SISO channel decoder in the previous iteration is used to estimate MAI from all users which will be subsequently subtracted out from the received signal. The performance of the proposed multiuser receiver is investigated for both synchronous and asynchronous channels.

The turbo multiuser receiver is further extended to MC-CDMA as well as MIMO-CDMA systems where parity bit selected and permutation spreading are used. Simulation results presented in this chapter show a significant improvement in the BER performance of CDMA-PB systems when a turbo multiuser receiver is implemented.

6.2 System Model

We consider a convolutionally coded CDMA system with K users in which each user employs parity bit selected spreading technique to spread its data bits. The block diagram of the transmitter of such a system is shown in Figure 6.1. The binary information data $\{d_k[i]\}$ for user k, $k = 1, \cdots, K$, are convolutionally encoded with code rate R and after passing through an interleaver, the interleaved coded bits $\{m_k[j]\}$ are BPSK modulated to obtain $\{b_k[j]\}$ with the bit duration of T_b. The interleaved coded bits are segmented into blocks of k_0 bits. During the lth signaling interval of T_s ($T_s = k_0 T_b$), the message vector to be transmitted by user k is $\underline{m}_k^l = [m_{k,1}^l, m_{k,2}^l, \cdots, m_{k,k_0}^l]$. This message vector is then input to the parity bit calculator that uses a systematic (n_0, k_0) linear block code to generate parity vector of $\underline{p}_k^l = [p_{k,1}^l, p_{k,2}^l, \cdots, p_{k,(n_0-k_0)}^l]$. The parity bit vector

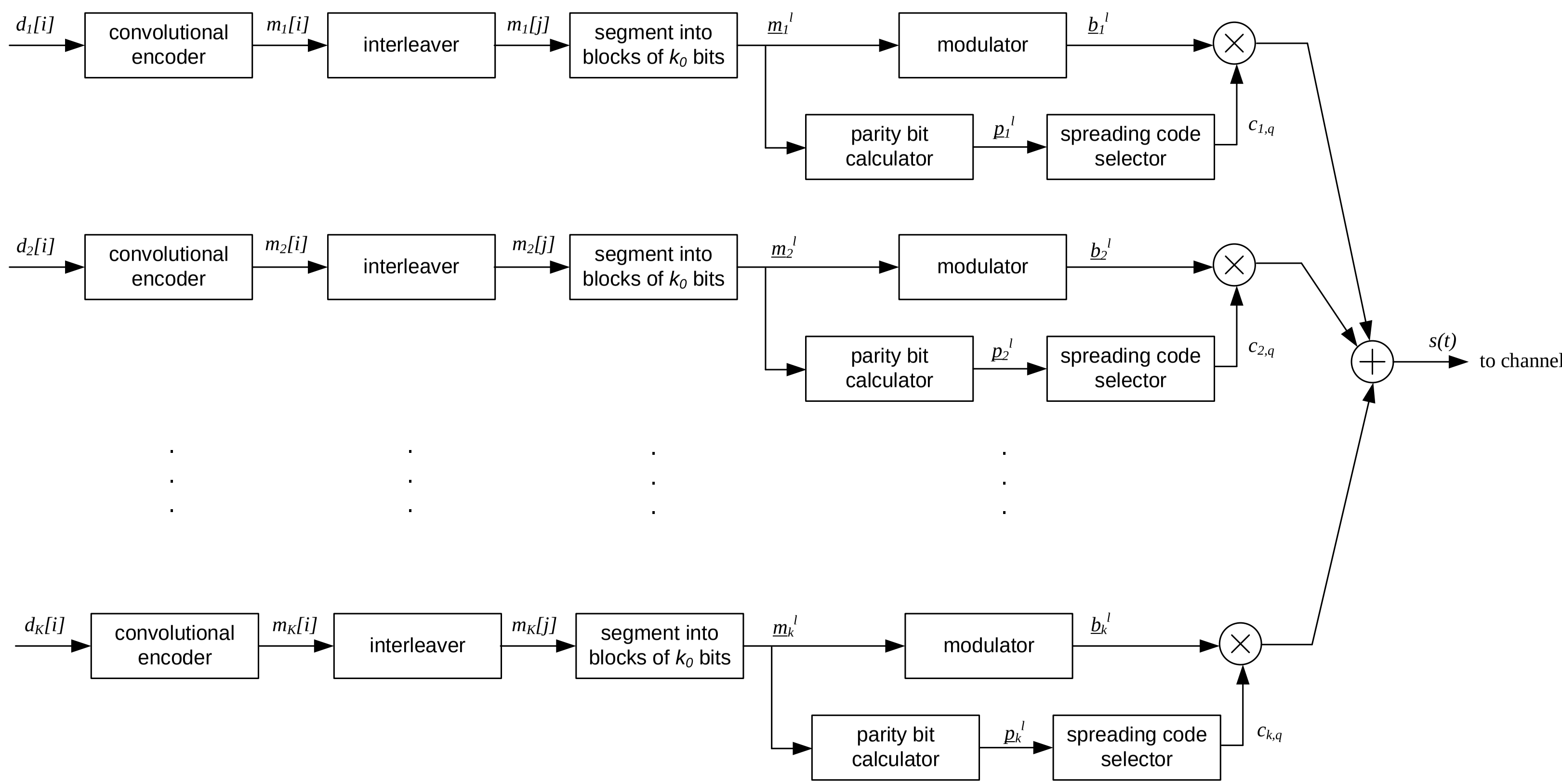

Figure 6.1: Transmitter block diagram of a coded CDMA system employing parity bit selected spreading sequences.

of $\underline{p}_k^l$ is used to select a spreading code from a set of $Q = 2^{(n_0-k_0)}$ mutually orthogonal spreading sequences: $C_k = \{\underline{c}_{k,1}, \cdots, \underline{c}_{k,q}, \cdots, \underline{c}_{k,Q}\}$ where $\underline{c}_{k,q} = [c_{k,q,1}, c_{k,q,2}, \ldots, c_{k,q,N}]$ is the N-chip qth spreading sequence of user k with chip duration of T_c ($T_b = NT_c$) and $c_{k,q,n} \in \{+1, -1\}$. The selected spreading code of $\underline{c}_{k,q_k}$ is then used to spread the modulated signals of all k_0 bits of $\underline{b}_k^l = [b_{k,1}^l, b_{k,2}^l, \cdots, b_{k,k_0}^l]$. On the time interval $lT_s < t < (l+1)T_s$, the transmitted signal from the kth user is

$$S_k(t) = A_k \cos \omega_c t \sum_{i=1}^{k_0} \sum_{n=1}^{N} b_{k,i}^l c_{k,q_k,n} p_{T_c}\left(t - (n-1)T_c - (i-1)T_b\right) \tag{6.1}$$

where

A_k	carrier amplitude of the kth user;
ω_c	carrier frequency;
q_k	index of the spreading code selected for the kth user;
$p_{T_c}(.)$	takes value 1 on the interval $[0, T_c)$ and 0 otherwise.

Assuming transmission through an AWGN channel with coherent detection, the equivalent baseband signal received at the base station is

$$r(t) = \sum_{k=1}^{K} \sum_{i=1}^{k_0} \sum_{n=1}^{N} A_k b_{k,i}^l c_{k,q_k,n} p_{T_c}\left(t - (n-1)T_c - (i-1)T_b - \tau_k\right) + n(t) \tag{6.2}$$

where

τ_k	time delay of the kth user's signal;
$n(t)$	zero-mean white Gaussian noise with variance σ_n^2.

Construction of Sets of Mutually Orthogonal Spreading Codes for Each User

In order to implement a CDMA system employing parity bit selected spreading sequences, each user needs to have a set of mutually orthogonal spreading sequences. We follow the method described in [17] to construct such spreading codes for each user.

Let A be a set of K spreading sequences of length N with desirable auto-correlation and cross-correlation properties: $A = \{\underline{a}_1, \underline{a}_2, \cdots, \underline{a}_K\}$ where $\underline{a}_k = [a_{k,1}, a_{k,2}, \cdots, a_{k,N}]$, and let H be $N \times N$ Hadamard matrix with $h_{i,j}$ as its ijth entry.

Set of Q $(Q < N)$ mutually orthogonal spreading sequences of length N for the kth user is then constructed as

$$\underline{c}_{k,q} = [a_{k,1}h_{q,1}, a_{k,2}h_{q,2}, \cdots, a_{k,N}h_{q,N}], \quad q = 1, 2, \cdots, Q. \tag{6.3}$$

In this way each user has a set of Q mutually orthogonal spreading sequences and low cross-correlation between any two spreading sequences of different users is also maintained. It is also possible to use PN sequences such as m-sequence and Gold sequence that have near orthogonality properties. Using an appropriate multiuser detection at the receiver can remove the interference caused by non-perfectly orthogonal spreading waveforms.

To be able to determine which spreading code is used by each user, the receiver is equipped with KQ matched filters, or in other words K filter banks of Q matched filter each matched to one of the spreading codes of set C_k. The output of the qth matched filter of the kth user on the jth bit interval is

$$z_{k,q}[j] = \int_{jT_b}^{(j+1)T_b} r(t) \sum_{n=1}^{N} c_{k,q,n} p_{T_c}(t - (n-1)T_c) dt \tag{6.4}$$

As the same spreading code is used to spread all the bits of each block of modulated bits of each user, the output of the matched filters should be observed over k_0 bit intervals: $\underline{z}^l_{k,q} = [z^l_{k,q,1}, z^l_{k,q,2}, \cdots, z^l_{k,q,k_0}]$, where $z^l_{k,q,i}$ represents the qth matched filter output of the kth user during the ith bit interval. In the following sections, for notational simplicity, we drop the block index l in all vectors representing a block of information bits whenever possible.

6.3 Turbo Multiuser Detection for Synchronous CDMA-PB Systems

The proposed turbo multiuser receiver introduced in this chapter, conceptually resembles the general structure and follows similar algorithms of other turbo receivers explained in the previous chapters. As the notations are different in the multiuser scenario, in this section we repeat the explanation of the turbo receiver in the case of MUD.

The block diagram of the proposed turbo multiuser receiver is shown in Figure 6.2. It consists of two parts: a SISO multiuser detector that is designed for CDMA-PB systems followed by K parallel single user SISO channel decoders. These two parts are connected via an interleaver and a deinterleaver. The proposed receiver's structure is inspired by the system introduced in [38] where a turbo multiuser detection is proposed for coded CDMA systems.

The SISO multiuser detector observes the received signal (matched filter outputs of all users) and by considering the *extrinsic* information provided by the SISO channel decoders in the previous iteration, delivers the *a posteriori* LLR of a transmitted "+1" and a transmitted "−1" for each coded bit of every user,

$$\Lambda_d\left(b_{k,i}\right) \triangleq \ln \frac{P\left(b_{k,i}=+1|r(t)\right)}{P\left(b_{k,i}=-1|r(t)\right)}, \quad k=1,2,\cdots,K; \;\; i=1,2,\cdots,k_0. \tag{6.5}$$

Using Bayes' rule, we can rewrite (6.5) as

$$\Lambda_d\left(b_{k,i}\right) = \underbrace{\ln \frac{P\left(r(t)|b_{k,i}=+1\right)}{P\left(r(t)|b_{k,i}=-1\right)}}_{\lambda_d\left(b_{k,i}\right)} + \underbrace{\ln \frac{P\left(b_{k,i}=+1\right)}{P\left(b_{k,i}=-1\right)}}_{\lambda_c^p\left(b_{k,i}\right)}. \tag{6.6}$$

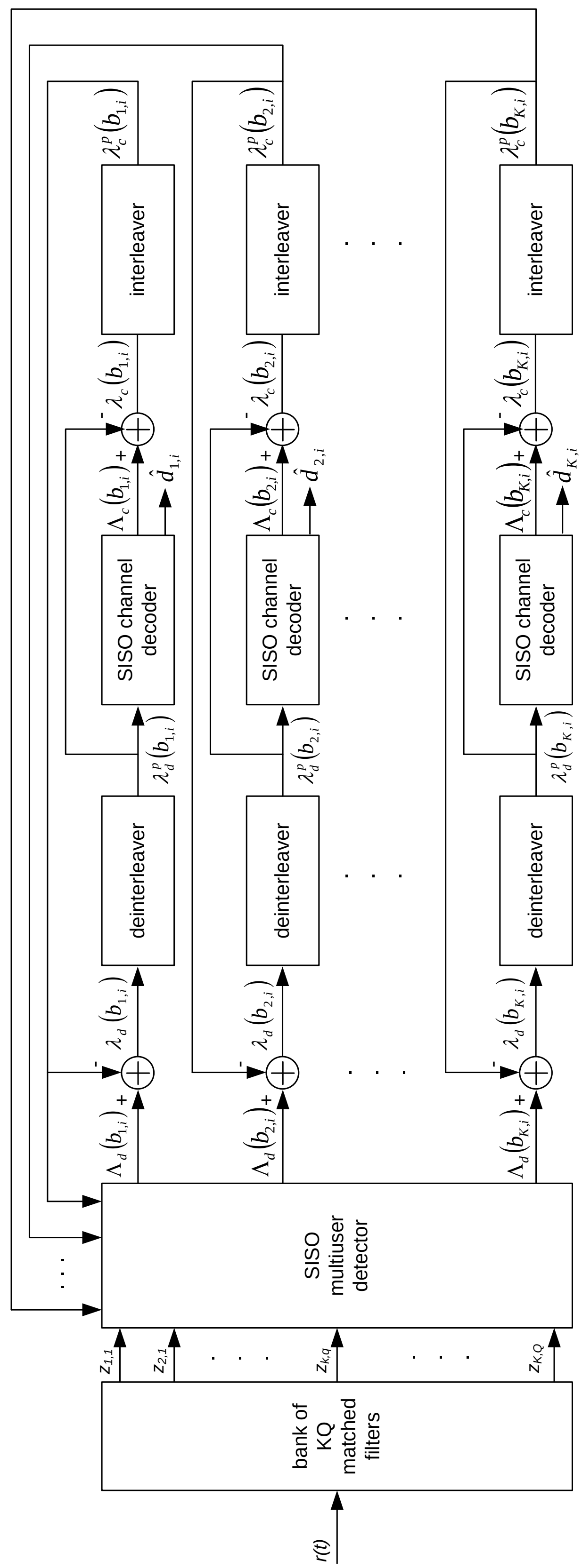

Figure 6.2: Turbo multiuser receiver of a CDMA system employing parity bit selected spreading sequences.

The second term in (6.6), denoted by $\lambda_c^p(b_{k,i})$, represents the *a priori* LLR of the coded bit $b_{k,i}$ calculated by the kth user's SISO channel decoder in the previous iteration (as indicated by a superscript p to show the previous iteration). For the first round of detection (i.e. first iteration) when there is no prior information available about the likelihood of coded bits, we assume all the coded bits can be "+1" or "−1" equiprobably. In other words $\lambda_c^p(b_{k,i}) = 0$ for $k = 1, 2, \cdots, K$ and $i = 1, 2, \cdots, k_0$.

The first term in (6.6), denoted by $\lambda_d(b_{k,i})$, represents the *extrinsic* information provided by the SISO multiuser detector about the ith coded bit of the kth user, $b_{k,i}$. This *extrinsic* information is calculated by the SISO multiuser detector based on the received signal $r(t)$ (or in other words the matched filter outputs), the structure of the SS-PB system and the prior information provided by the SISO channel decoders of all users in the previous iteration about the coded bits other than the ith bit, $\{\lambda_c^p(b_{k,j}), j \neq i\}$. This *extrinsic* information is then passed through a deinterleaver and feed into the SISO channel decoder as *a priori* information in the next iteration.

By making use of the prior information provided by the SISO multiuser detector, $\lambda_d^p(b_{k,i})$, and also considering the trellis structure of the convolutional code used in the transmitter, the kth user's SISO channel decoder calculates the *a posteriori* LLR of each coded bit,

$$\begin{aligned}\Lambda_c(b_{k,i}) &\triangleq \ln \frac{P\left(b_{k,i} = +1 | \{\lambda_d^p(b_{k,i})\}_{i=1}^{k_0}; \text{decoding}\right)}{P\left(b_{k,i} = -1 | \{\lambda_d^p(b_k)\}_{i=1}^{k_0}; \text{decoding}\right)} \\ &= \lambda_c(b_{k,i}) + \lambda_d^p(b_{k,i})\end{aligned} \tag{6.7}$$

where the *a posteriori* probabilities are conditioned on the *extrinsic* information from the SISO multiuser detector in the last iteration, $\{\lambda_d^p(b_{k,i})\}_{i=1}^{k_0}$, and the coding structure shown by "decoding" in (6.7).

It can be seen in (6.7) that the *a posteriori* LLR $\Lambda_c(b_{k,i})$, is the summation of the prior information $\lambda_d^p(b_{k,i})$ and the *extrinsic* information $\lambda_c(b_{k,i})$ that is the information gleaned about $b_{k,i}$ from prior information of other coded bits based on the trellis structure of the channel encoder. The SISO channel decoder also computes the *a posteriori* LLR of the data bits $\Lambda_c(d_{k,i})$, that is used in the last iteration to make decision about the decoded bits. The *extrinsic* information delivered in this stage $\lambda_c(b_{k,i})$ from all user's SISO channel decoders, is then passed an interleaver and fed back to the SISO multiuser detector as the *a priori* information in the next iteration (i.e. $\lambda_c^p(b_{k,i})$).

The SISO channel decoder discussed in this chapter that provides the *a posteriori* information about both coded and data bits is based on the algorithm introduced in [38] which is a slight modification of the well know BCJR algorithm.

For synchronous channel, in equation (6.2), $\tau_k = 0$ for all $k = 1, \cdots, K$ and it is easily seen that a sufficient statistic for demodulating the ith coded bit of the kth user is a matrix $\underline{Z}$ that includes all the KQ matched filter outputs, $z_{k,q,i}$, over k_0 bit intervals:

$$\underline{Z} = \left[\underline{z}_{1,1}^T, \underline{z}_{1,2}^T, \cdots, \underline{z}_{1,Q}^T, \cdots, \underline{z}_{K,1}^T, \cdots, \underline{z}_{K,Q}^T\right]^T. \tag{6.8}$$

Let us define matrix $\underline{B}$ that includes all the coded bits of K users over k_0 bit intervals:

$$\underline{B} = \left[\underline{b}_1^T, \underline{b}_2^T, \cdots, \underline{b}_K^T\right]^T. \tag{6.9}$$

Defining matrix $\underline{Y}$ as a noise and MAI free KQ matched filter outputs over k_0 bit intervals when $\underline{B}$ is sent, matrix $\underline{Z}$ can be written as [11]

$$\underline{Z} = \underline{R}\ \underline{A}\ \underline{Y} + \underline{N} \tag{6.10}$$

where

$$\underline{A} = \text{diag}(\overbrace{A_1, \cdots, A_1}^{Q\text{times}}, \overbrace{A_2, \cdots, A_2}^{Q\text{times}}, \cdots, \overbrace{A_K, \cdots, A_K}^{Q\text{times}}); \tag{6.11}$$

$\underline{N}$ is a $KQ \times k_0$ matrix of Gaussian noise samples at the output of the matched filters; and $\underline{R}$ is correlation matrix which is defined as

$$\underline{R} = \begin{bmatrix} \underline{I} & \underline{\rho}^{(1,2)} & \cdots & \underline{\rho}^{(1,K)} \\ \underline{\rho}^{(2,1)} & \underline{I} & \cdots & \underline{\rho}^{(2,K)} \\ \vdots & \vdots & \ddots & \vdots \\ \underline{\rho}^{(K,1)} & \underline{\rho}^{(K,2)} & \cdots & \underline{I} \end{bmatrix} \tag{6.12}$$

where $\underline{I}$ is the $Q \times Q$ identity matrix and $\underline{\rho}^{(p,q)}$ is the matrix of cross-correlation between spreading sequences of user p and q and it is defined as

$$\underline{\rho}^{(p,q)} = \begin{bmatrix} \rho_{1,1}^{(p,q)} & \cdots & \rho_{1,Q}^{(p,q)} \\ \vdots & \ddots & \vdots \\ \rho_{Q,1}^{(p,q)} & \cdots & \rho_{Q,Q}^{(p,q)} \end{bmatrix} \tag{6.13}$$

in which $\rho_{i,j}^{(p,q)}$ is the cross-correlation between the ith spreading code of user p and the jth spreading code of user q. Identity matrices in $\underline{R}$ come from the fact that spreading sequences of each user are mutually orthogonal; hence, they have zero cross-correlation.

6.3.1 Optimal SISO Multiuser Detector

In this section we explain how the SISO multiuser detector for a CDMA-PB system uses the matched filter outputs in combine with the prior information from the channel decoders in the previous iteration to optimally calculate the *extrinsic* information for every detected bit of every user.

Let us define

$$\mathcal{B}_{+}^{k,i} \triangleq \{\underline{B} : [\underline{B}]_{k,i} = +1; \text{and } [\underline{B}]_{m,n} \in (+1,-1) \text{ for all } m \neq k \text{ } and \text{ } n \neq i\}, \tag{6.14}$$

that consists of all the modulated bit matrices, $\underline{B}$, where the nth bit of the kth user is "+1". Similarly $\mathcal{B}_{-}^{k,i}$ is defined as

$$\mathcal{B}_{-}^{k,i} \triangleq \{\underline{B} : [\underline{B}]_{k,i} = -1; \text{and } [\underline{B}]_{m,n} \in (+1,-1) \text{ for all } m \neq k \textit{ and } n \neq i\}, \tag{6.15}$$

The *extrinsic* information delivered by the SISO multiuser detector $\lambda_d(b_{k,i})$ can be rewritten as

$$\begin{aligned}\lambda_d(b_{k,i}) &= \ln \frac{P(r(t)|b_{k,i}=+1)}{P(r(t)|b_{k,i}=-1)} \\ &= \ln \frac{\sum_{\underline{B}\in\mathcal{B}_{+}^{k,i}} P(\underline{Z}|\underline{B}) \prod_{\substack{m\neq k\\ n\neq i}} P(b_{m,n})}{\sum_{\underline{B}\in\mathcal{B}_{-}^{k,i}} P(\underline{Z}|\underline{B}) \prod_{\substack{m\neq k\\ n\neq i}} P(b_{m,n})}\end{aligned} \tag{6.16}$$

where $P(b_{m,n})$ is the probability that the nth bit of the mth user equals to $b_{m,n}$ for $b_{m,n} \in \{+1,-1\}$. From (6.10) since the noise has a normal distribution, (6.16) can be rewritten as

$$\lambda_d(b_{k,i}) = \ln \frac{\sum\limits_{\underline{B}\in\mathcal{B}_{+}^{k,i}} \exp\left(-\frac{1}{\sigma_n^2}\sum\limits_{j=1}^{k_0}(\underline{Z}_j - \underline{R}\,\underline{A}\,\underline{Y}_j)^T\underline{R}^{-1}(\underline{Z}_j - \underline{R}\,\underline{A}\,\underline{Y}_j)\right)\prod\limits_{\substack{m\neq k\\ n\neq i}} P(b_{m,n})}{\sum\limits_{\underline{B}\in\mathcal{B}_{-}^{k,i}} \exp\left(-\frac{1}{\sigma_n^2}\sum\limits_{j=1}^{k_0}(\underline{Z}_j - \underline{R}\,\underline{A}\,\underline{Y}_j)^T\underline{R}^{-1}(\underline{Z}_j - \underline{R}\,\underline{A}\,\underline{Y}_j)\right)\prod\limits_{\substack{m\neq k\\ n\neq i}} P(b_{m,n})} \tag{6.17}$$

where $\underline{Z}_j$ and $\underline{Y}_j$ represent the jth columns of $\underline{Z}$ and $\underline{Y}$ respectively. Similar to the derivation shown in equation (3.16), the *a priori* probabilities of $P(b_{m,n})$ can be calculated based their LLR's $\lambda_c^p(b_{k,i})$ as follows

$$P(b_{m,n}) = \frac{1}{2}\left[1 + b_{m,n}\tanh\left(\frac{1}{2}\lambda_c^p(b_{m,n})\right)\right]. \tag{6.18}$$

By substituting (6.18) into (6.17) we obtain an equation that shows how an optimal SISO multiuser detector calculates the *extrinsic* information of each coded bits.

6.3.2 Low-Complexity SISO Multiuser Detector Based on Soft Interference Cancellation

It is clear from (6.17) that the optimal SISO multiuser detector has the complexity which is exponential in terms of the number of users K. It makes it quite impossible to employ the optimal detection for the systems with medium to large number of users. In this section we develop a low-complexity approximate SISO multiuser detector for CDMA-PB systems based on soft interference cancellation (SIC) and single user turbo receiver for SS-PB systems.

In the proposed low-complexity SISO multi user detector, at first the soft estimate of the multiple access interference is calculated and subtracted out from the received signal and then K parallel single user turbo receivers for SS-PB systems explained in Chapter 3 are used to separately calculate the extrinsic information of every coded bit of every user.

Let us define $\underline{Y}^{(k)}$ as a part of matrix $\underline{Y}$ that represents signals of user k, in other words: $[\underline{Y}^{(k)}]_{i,j} = [\underline{Y}]_{Q(k-1)+i,j}$. The jth element of the ith row in $\underline{Y}^{(k)}$ can be estimated as

$$\tilde{y}_{i,j}^{(k)} \triangleq E[y_{i,j}^{(k)}] = \sum_{y_{i,j}^{(k)} \in \{-1,0,1\}} y_{i,j}^{(k)} P(y_{i,j}^{(k)}) \tag{6.19}$$

where $E[.]$ denotes the expected value. Let us define S_q as a set of all bit vectors, $\underline{b}_k$, that are associated to the qth spreading code. In other words all the bit vectors of S_q belong to a coset that generate the qth parity bit vector. Since $y_{i,j}^{(k)}$ has a non-zero value only when $\underline{b}_k \in S_i$, (6.19) can be rewritten as

$$\tilde{y}_{i,j}^{(k)} = \sum_{\underline{b}_k \in S_i} b_{k,j} P(\underline{b}_k) \tag{6.20}$$

The probability of vector $\underline{b}_k$, $P(\underline{b}_k)$, is calculated based on the *a priori* probabilities as

$$P(\underline{b}_k) = \prod_{r=1}^{k_0} P(b_{k,r}) = \prod_{r=1}^{k_0} \frac{1}{2}\left[1 + b_{k,r} \tanh\left(\frac{1}{2}\lambda_c^p\left(b_{k,r}\right)\right)\right] \tag{6.21}$$

where the second equality follows from (6.18). By substituting (6.21) into (6.20), we have

$$\tilde{y}_{i,j}^{(k)} = \sum_{\underline{b}_k \in S_i} b_{k,j} \prod_{r=1}^{k_0} \frac{1}{2}\left[1 + b_{k,r} \tanh\left(\frac{1}{2}\lambda_c^p\left(b_{k,r}\right)\right)\right] \tag{6.22}$$

that are used to construct the matrix $\underline{\tilde{Y}}^k$. By having $\underline{\tilde{Y}}^k$ for all $k = 1, \cdots, K$, the estimate matrix of $\underline{\tilde{Y}}$ is formed as follows

$$\underline{\tilde{Y}} \triangleq [\underline{\tilde{Y}}^{(1)^T}, \underline{\tilde{Y}}^{(2)^T}, \cdots, \underline{\tilde{Y}}^{(K)^T}]^T. \tag{6.23}$$

Now that we have the estimate matrix of $\underline{\tilde{Y}}$, MAI caused by every user can be estimated and subtracted out from the matched filter outputs:

$$\underline{\tilde{Z}} \triangleq \underline{Z} - (\underline{R} - \overline{I})\underline{A}\, \underline{\tilde{Y}} \tag{6.24}$$

where $\overline{I}$ is a $KQ \times KQ$ identity matrix and $\underline{\tilde{Z}}$ represents the matched filter outputs after MAI estimation and subtraction. The new matrix of matched filter outputs, $\underline{\tilde{Z}}$, is then divided into K matrices of $\underline{\tilde{Z}}^{(k)}$ for each user. In other words $[\underline{\tilde{Z}}^{(k)}]_{i,j} = [\underline{\tilde{Z}}]_{Q(k-1)+i,j}$.

Following exactly the same algorithms explained in Chapter 3 for a single user SISO detector of a SS-PB system, every user separately and in parallel considers $\underline{\tilde{Z}}^{(k)}$ as the matched filter output and calculates the *extrinsic* information for each coded bit. In the proposed low-complexity multiuser detector, after MAI estimation and cancellation, processing is done separately for each user that leads to a much less complex receiver in compare to the optimal SISO multiuser detector.

6.3.3 Simulation Results

In this section, some simulation results are presented in order to illustrate the performance of the low-complexity turbo multiuser receiver in synchronous CDMA systems that employ parity bit selected spreading codes. All users employ the same $1/2$ rate convolutional code with constraint length of 5 and the generators 23, 35 in octal notation. Each user's coded bits are passed through different random interleavers before being fed into SS-PB modulator. The simulations are performed for SS-PB systems based on linear $(10, 6)$ block encoder whose parity matrix is shown in (2.9). The simulations are performed for AWGN synchronous channels and a coherent detection is assumed in the receiver.

We assume that cross-correlation between any two spreading sequences of different users are the same, in other words:

$$\rho_{i,j}^{(p,q)} = \rho \quad \text{for all } i, j \text{ and } p \neq q. \tag{6.25}$$

In all simulations presented in this section, a cross-correlation of $\rho = 0.3$ is assumed which means that all spreading codes from different users have 30% cross-correlation. In all the scenarios, a perfect power control is in place which means all the received signals from all users have the same power. Figures 6.3 to 6.6 respectively present the BER performance of the low-complexity turbo multiuser receiver in synchronous CDMA-PB systems with 4, 6, 8 and 10 users. The simulations are performed for the first 4 iterations.It

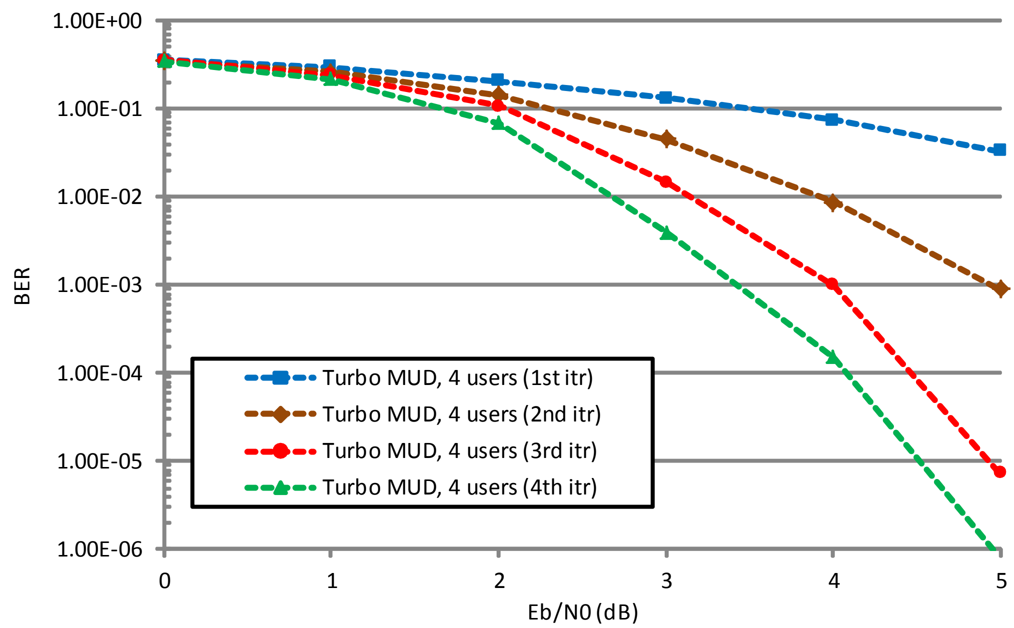

Figure 6.3: Performance of the turbo multiuser receiver that employs iterative SIC in CDMA-PB systems. $K = 4$, $\rho = 0.3$ and PB is based on the linear $(10, 6)$ block encoder.

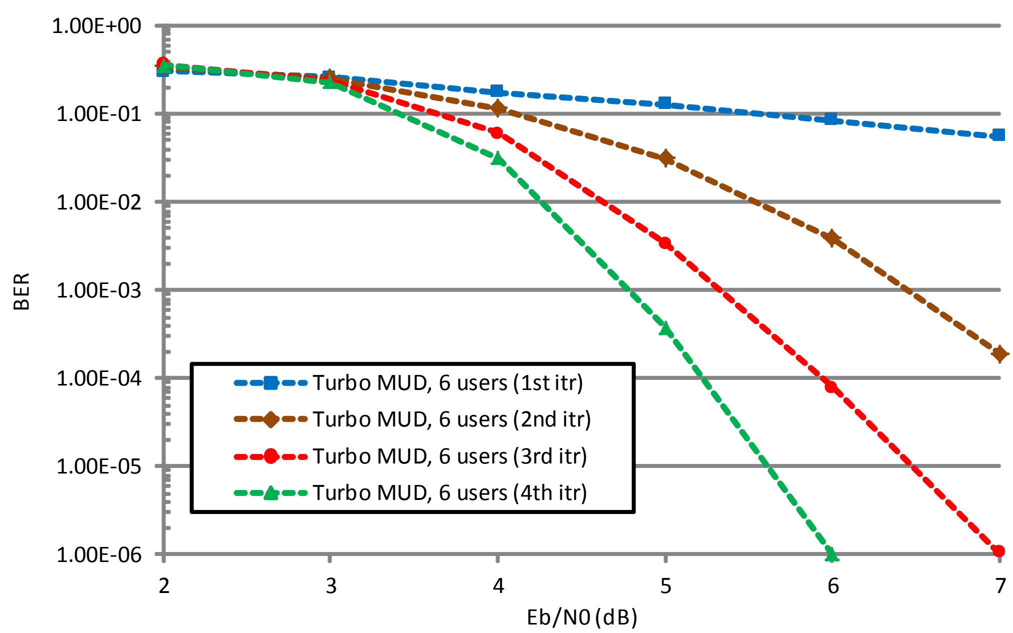

Figure 6.4: Performance of the turbo multiuser receiver that employs iterative SIC in CDMA-PB systems. $K = 6$, $\rho = 0.3$ and PB is based on the linear $(10, 6)$ block encoder.

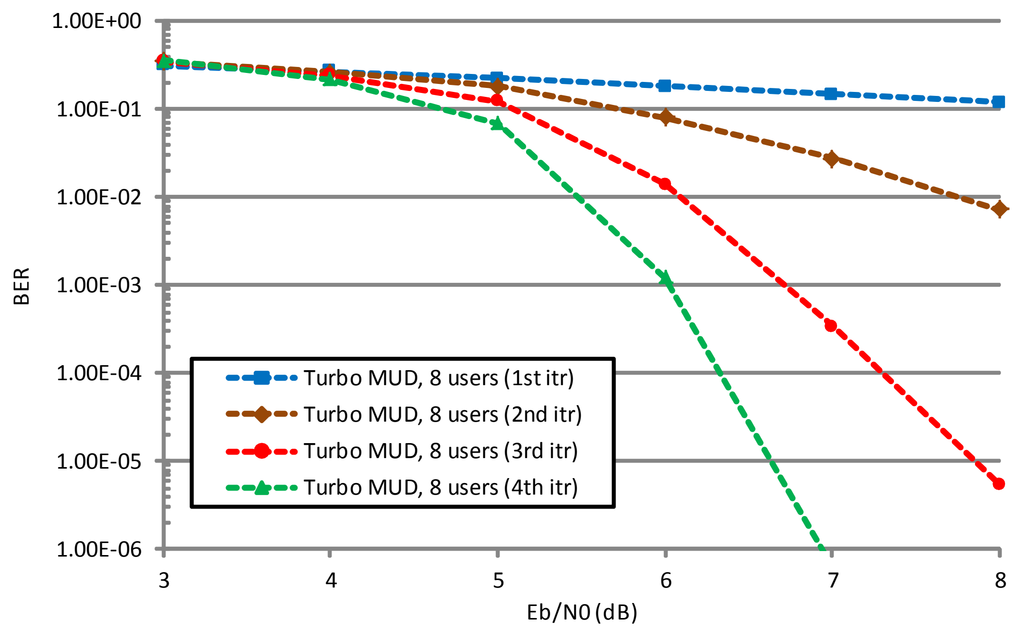

Figure 6.5: Performance of the turbo multiuser receiver that employs iterative SIC in CDMA-PB systems. $K = 8$, $\rho = 0.3$ and PB is based on the linear $(10, 6)$ block encoder.

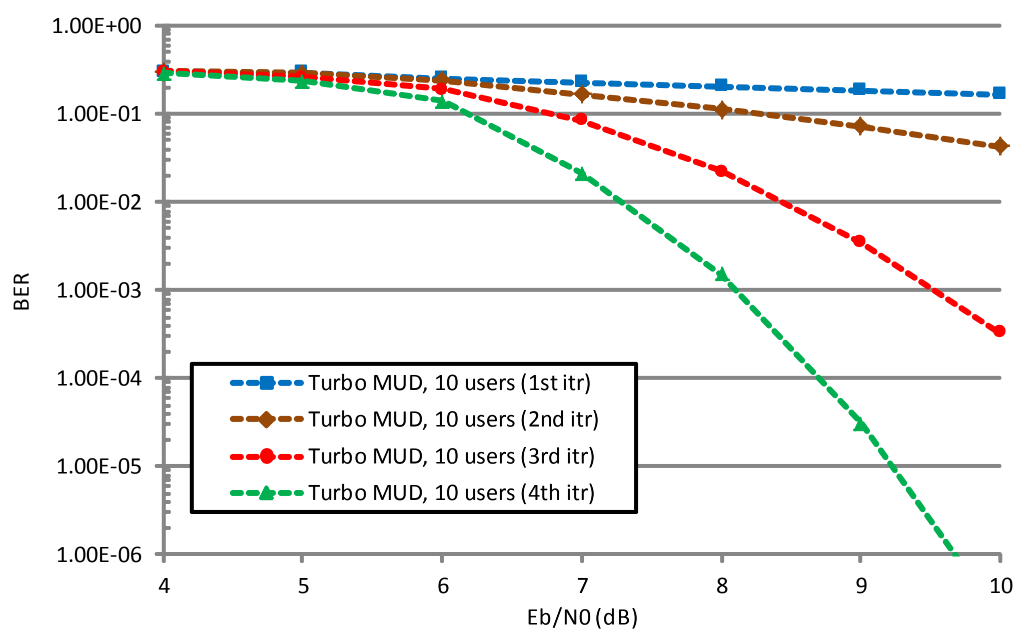

Figure 6.6: Performance of the turbo multiuser receiver that employs iterative SIC in CDMA-PB systems. $K = 10$, $\rho = 0.3$ and PB is based on the linear $(10, 6)$ block encoder.

is seen form these figures that at reasonably high SNR, the proposed low-complexity multiuser receiver offers a significant performance gain over multiple iterations.

6.4 Turbo Multiuser Detection for Asynchronous CDMA-PB Systems

In the asynchronous channel, we lift the restriction that all user's signals are received synchronously at the receiver. In the asynchronous channel, the time delay of every user's signal, τ_k, is uniformly distributed on $[0, T_b)$. For the sake of simplicity, we assume that the channel is chip synchronous, in other words, $\tau_k = n_k T_c$ where n_k is uniformly distributed on the set $\{0, 1, \cdots, N-1\}$. Without loss of generality, we assume $\tau_1 = 0$ and $\tau_1 < \tau_2 < \cdots < \tau_K$.

The output of the vth matched filter of the uth user on the jth bit interval is

$$z_{u,v}[j] = \sum_{k=1}^{K} \sum_{q=1}^{Q} \sum_{m=-1}^{1} A_k x_{k,q}[j] \gamma_{u,v,k,q}^{(m)} + n_{u,v}[j] \tag{6.26}$$

where $x_{k,l}[j]$ is the noise and MAI free qth matched filter output of the kth user in the jth bit interval. Here, $\gamma_{u,v,k,q}^{(-1)}, \gamma_{u,v,k,q}^{(0)}$, and $\gamma_{u,v,k,q}^{(1)}$ are the cross correlation between the uth spreading sequence of the vth user during the jth bit interval and the qth spreading sequence of the kth user during the $(j-1)$th, (j)th, and $(j+1)$th bit intervals, respectively, and $n_{u,v}$ is a zero mean Gaussian random variable with

$$E[n_{u,v}[j] n_{k,l}^*(j+m)] = \begin{cases} \frac{\sigma_n^2 \gamma_{u,v,k,q}^{(m)}}{T_b}, & m = -1, 0, 1 \\ 0, & \text{otherwise.} \end{cases} \tag{6.27}$$

In the case of asynchronous channel, it can be seen that the matched filter output during each bit interval is affected not only by the k_0-long bit vector that the bit comes from, but also the previous and the next block of k_0 bit vectors, in other words, $z^l_{k,q}$ is affected by b^{l-1}_k, b^l_k and b^{l+1}_k.

Considering the same notations as of section 6.3, the matrix of KQ matched filter outputs over k_0 bit intervals, $\underline{Z}$, can be derive from (6.26) as

$$\underline{Z} = \underline{\Gamma}^{(-1)}\underline{A}\ \underline{X}^{(-1)} + \underline{\Gamma}^{(0)}\underline{A}\ \underline{X}^{(0)} + \underline{\Gamma}^{(1)}\underline{A}\ \underline{X}^{(1)} + \underline{N} \tag{6.28}$$

where

- $\underline{Z}$ — $KQ \times k_0$ matrix in which $[\underline{Z}]_{k\circ q,n} = z_{k,q}[j+n]$ with $k \circ q = (k-1)Q + q$ and for $n = 1, \cdots, k_0$;
- $\underline{X}^{(m)}$ — $KQ \times k_0$ matrix where $[\underline{X}^{(m)}]_{k\circ q,n} = x_{k,q}[j+m+n]$, for $m = -1, 0, 1$ and $n = 1, \cdots, k_0$;
- $\underline{\Gamma}^{(m)}$ — $KQ \times KQ$ matrix where $[\underline{\Gamma}^{(m)}]_{k\circ q,u\circ v} = \gamma^{(m)}_{k,q,u,v}$ for $m = -1, 0, 1$;
- $\underline{A}$ — $KQ \times KQ$ diagonal matrix of user's received signal amplitudes as defined in (6.11);
- $\underline{N}$ — $KQ \times k_0$ matrix of noise samples where $[\underline{N}]_{k\circ q,n} = n_{k,q}[j+n]$.

The proposed turbo multiuser detector for asynchronous CDMA-PB systems is based on SIC where the soft MAI is estimated in each iteration and subtracted out from the received signal. Similar to the case of synchronous systems, at first, the Q matched filter outputs of every user are separately observed in order to calculate the *extrinsic* information of every coded bit of every user following the same steps as the single SS-PB turbo receiver. Having the *extrinsic* information of each coded bit and using similar calculation as (6.23), MAI can be estimated and subtracted out from the matched filter outputs.

Denoting $\underline{\tilde{X}}^{(m)}$ as the estimation of $\underline{X}^{(m)}$, the estimated matrix of matched filter outputs, $\underline{\tilde{Z}}^{(m)}$, after MAI subtraction is

$$\underline{\tilde{Z}} = \underline{Z} - \underline{\Gamma}^{(-1)}\underline{A}\ \underline{\tilde{X}}^{(-1)} + \left(\underline{\Gamma}^{(0)} - \overline{I}\right)\underline{A}\ \underline{\tilde{X}}^{(0)} + \underline{\Gamma}^{(1)}\underline{A}\ \underline{\tilde{X}}^{(1)} \tag{6.29}$$

where $\overline{I}$ is a $KQ \times KQ$ identity matrix. When the matched filter outputs are cleared from the estimated MAI, the *extrinsic* information of each user's data bit is separately calculated and fed back for the MAI estimation of the following iteration.

6.4.1 Simulation Results

Some simulation results that illustrate the performance of the proposed turbo receiver in asynchronous CDMA-PB system is presented in this section. The simulation is performed for 4 users in an AWGN channel where the user's signals are chip synchronous but bit asynchronous. All users employ the same 1/2 rate convolutional code with constraint length of 5 and the generators 23, 35 in octal notation. Each user's coded bits are passed through different random interleavers before being fed into SS-PB modulator which is based on the linear $(10, 6)$ block encode.

The spreading sequences are of length $N = 32$ chips (processing gain of 32) and are generated randomly. Time delays of each user are randomly selected with uniform distribution within one bit interval. Following the method explained in section 6.2, 32×32 Hadamard matrix is used in order to create sets of 16 mutually orthogonal spreading sequences for each 4 users. Cross-correlation matrices of $\underline{\Gamma}^{(-1)}$, $\underline{\Gamma}^{(0)}$ and $\underline{\Gamma}^{(1)}$ are computed for the selected spreading codes of every user.

Figure (6.7) presents the BER performance of the proposed turbo multiuser receiver for asynchronous CDMA-PB system over 4 iterations. From this figure, it can be seen that a significant improvement in BER performance is achieved when a turbo multiuser

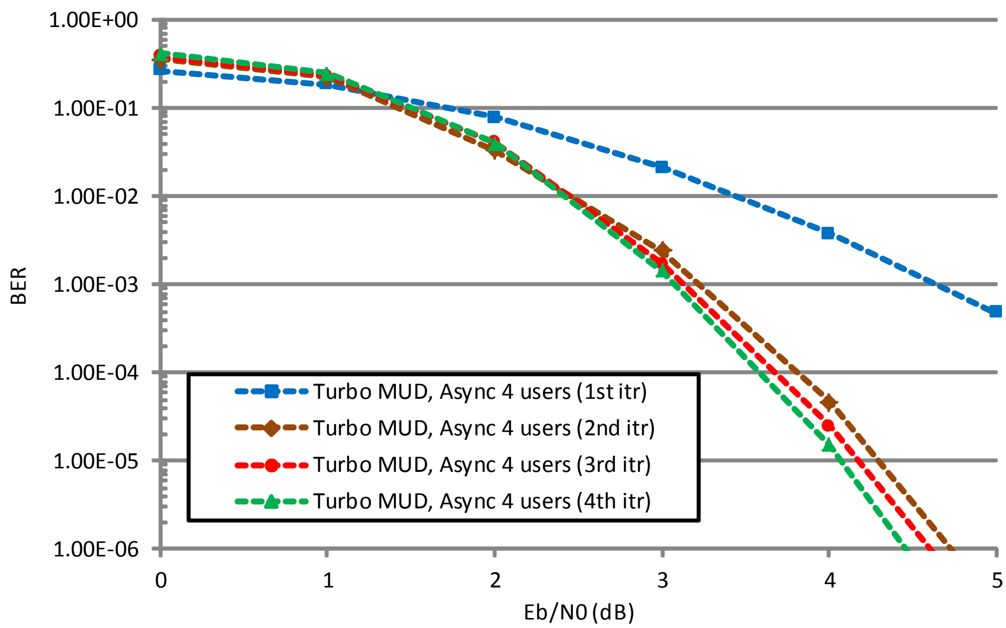

Figure 6.7: Performance of the turbo multiuser receiver in an asynchronous CDMA-PB system that employs parity bit selected spreading sequences. $K = 4$ and PB is based on the linear $(10, 6)$ block encoder.

detection is employed.

6.5 Turbo Multiuser Receiver for MC-CDMA Systems Employing Parity Bit Selected Spreading Sequences

In this section the multiuser scenario of MC-SS-PB systems is considered. Similar to the CDMA-PB systems explained in section 6.2, in MC-CDMA systems that employ parity bit selected spreading code (MC-CDMA-PB), every user is provided by a set of Q mutually orthogonal spreading codes and each one modulates its data bits using MC-SS-PB modulation scheme. We assume that all users share the same subcarriers but spread by different spreading sequences.

An outer convolutional encoder is employed in the transmitter of all K users. The convolutionally encoded bits are then fed into the MC-SS-PB modulator. The transmitter block diagram of a single user coded MC-SS-PB system is shown in Figure 4.1.

Assuming synchronous channel and following the same notations as of Chapter 4, the equivalent baseband received signal of K users in the base station is

$$r(t) = \sum_{k=1}^{K} \sum_{v=1}^{N_c} \mathrm{Re}\left[A_k \alpha_{k,v} \sqrt{\frac{2}{T_s}} b_{k,v} c_{q_k}(t) e^{j2\pi f_v t} \right] + n(t) \tag{6.30}$$

where

A_k	kth user signal amplitude;
$\alpha_{k,v}$	complex fading gain of the kth user in the vth sub-channel;
$b_{k,v}$	kth user's transmitted bit on the vth sub-channel;
f_v	carrier frequency of the vth subcarrier;
N_c	number of subcarriers;
$c_{q_k}(t)$	selected spreading code for the kth user;
$n(t)$	Gaussian noise.

It can be seen that the sufficient statistics to detect every user's data bits are the KQ matched filter outputs of the N_c sub-channels. We can rewrite the outputs of the bank of matched filters in matrix notation as

$$\underline{Z} = \underline{R}\ \underline{A}\ (\underline{Y} \odot \underline{W}) + \underline{N} \tag{6.31}$$

where

$\underline{Z}$ $KQ \times N_c$ matrix of matched filter outputs in which $[\underline{Z}]_{k \circ q, v}$ is the qth matched filter output of the kth user in the vth sub-channel;

$\underline{R}$ $KQ \times KQ$ matrix of cross-correlation where $[\underline{R}]_{i\circ j,m\circ n}$ is the cross-correlation between the ith spreading code of the jth user and the mth spreading code of the nth user;

$\underline{Y}$ $KQ \times N_c$ matrix of noise and MAI free matched filter outputs;

$\underline{N}$ $KQ \times N_c$ matrix of Gaussian noise samples;

$$\underline{A} = \text{diag}(\overbrace{A_1, \cdots, A_1}^{Q\text{times}}, \overbrace{A_2, \cdots, A_2}^{Q\text{times}}, \cdots, \overbrace{A_K, \cdots, A_K}^{Q\text{times}});$$

$$\underline{W} = [\overbrace{\underline{\alpha}_1^T, \cdots, \underline{\alpha}_1^T}^{Q\text{times}}, \overbrace{\underline{\alpha}_2^T, \cdots, \underline{\alpha}_2^T}^{Q\text{times}}, \cdots, \overbrace{\underline{\alpha}_K^T, \cdots, \underline{\alpha}_K^T}^{Q\text{times}}]^T \text{ where } \underline{\alpha}_i = [\alpha_{i,1}, \alpha_{i,2}, \cdots, \alpha_{i,N_c}];$$

and the operator $\odot$ is element by element matrix product, in other words, $[A \odot B]_{i,j} = [A]_{i,j}[B]_{i,j}$.

The proposed turbo multiuser detector for MC-CDMA-PB system is based on the soft interference estimation and cancellation. In each iteration, using the *extrinsic* information provided by the SISO decoder in the previous iteration, MAI caused by all users are estimated and subtracted out from the ouputs of the bank of matched filters. The MAI free matched filter outputs are then used in the next iteration to calculate the *extrinsic* information about every bit of every user.

Following the same derivations as shown in (6.19) to (6.23), estimated matrix of $\underline{\tilde{Y}}$ can be calculated as

$$\tilde{y}_{k\circ q,v} = \sum_{\underline{b}_k \in S_q} b_{k,v} \prod_{j=1}^{N_c} \frac{1}{2} \left[1 + b_{k,j} \tanh\left(\frac{1}{2} \lambda_c^p \left(b_{k,j} \right) \right) \right]. \tag{6.32}$$

By having the estimated matrix of $\underline{\tilde{Y}}$, MAI caused by all users can be estimated and

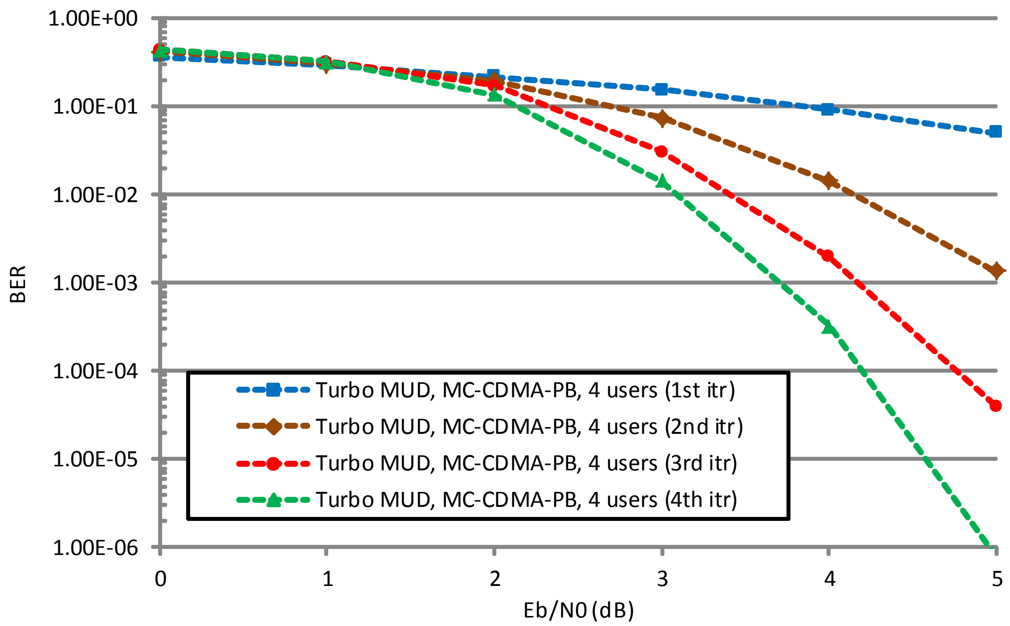

Figure 6.8: Performance of the turbo multiuser receiver for MC-CDMA-PB systems. $K = 4$ and PB is based on the linear $(10, 6)$ block encoder.

subtracted out from the matched filter outputs:

$$\underline{\tilde{Z}} = \underline{Z} - (\underline{R} - \overline{I})\underline{A}\,(\underline{\tilde{Y}} \odot \underline{W}) + \underline{N} \tag{6.33}$$

where $\overline{I}$ is a $KQ \times KQ$ identity matrix and $\underline{\tilde{Z}}$ represents the matched filter outputs after MAI estimation and subtraction. Matrix $\underline{\tilde{Z}}$ is then considered as the MAI free matched filter outputs for the following iteration.

6.5.1 Simulation Results

Simulation is performed for a coded CDMA-PB systems to illustrate the performance of the proposed turbo multiuser receiver. The same system parameters as of the system of section 4.4 is used. All users employ the same $1/2$ rate convolutional code with constraint length of 5 and generators 23, 35 in octal notation. The MC-SS-PB systems is based on linear $(10, 6)$ block encoder whose parity matrix is shown in (2.9).

We assume that every user has a set of mutually orthogonal spreading sequences and also that the cross-correlation between any two spreading sequences of different users are the same. All user's signals are transmitted through the synchronous channel that is assumed to be slowly-varying, frequency selective, Rayleigh channel, where the DS-SS signals in each sub-channel is fading nonselectively and independently. It is assumed that the receiver is able to perfectly estimate the fading gains of each sub-channel.

Figure 6.8 shows the performance of the proposed turbo multiuser receiver for MC-CDMA-PB system with $K = 4$ users. The cross-correlation between any spreading sequence of different users is assumed to be $\rho = 0.3$. From this figure, it can be seen that a significant performance improvement is achieved when turbo multiuser detection is applied in the MC-CDMA-PB systems.

6.6 Turbo Multiuser Receiver for MIMO-CDMA Systems Employing Parity Bit Selected and Permutation Spreading

In this section the multiuser scenario of the MIMO-CDMA systems that employ parity bit selected and permutation spreading is discussed and a turbo multiuser receiver is proposed for such systems. We follow the same system model and notations as of Chapter 5 where the single user turbo receiver for MIMO-CDMA systems is explained.

Let us assume an uplink channel of a synchronous MIMO-CDMA system with K users where each user has N_t transmit antennas and the base station has N_r receive antennas. To be able to employ the parity bit selected or permutation spreading techniques, every user is assigned with a set of Q mutually orthogonal spreading sequences.

All the N_r receive antennas of the base station are equipped with Q matched filters. Let us define $\underline{Z}^{(k)}$, as the $Q \times N_r$ matrix of matched filter outputs of the kth user whose (i,j)th element, $z_{i,j}^{(k)}$ is the ith matched filter output of the jth receive antenna. Then we can rewrite (5.2) and (5.3) in matrix format as

$$\underline{Z}^{(k)} = \underline{Y}^{(k)} \ \underline{H}^{(k)} + \underline{N}^{(k)} \tag{6.34}$$

where $\underline{H}^{(k)}$ is the $N_t \times N_r$ channel gain matrix between the kth user's N_t transmit antennas and N_r base station's receive antennas, $\underline{N}^{(k)}$ is the matrix of Gaussian noise samples and the elements of $\underline{Y}^{(k)}$ is defined as

$$y_{i,j}^{(k)} \triangleq \begin{cases} b_j^{(k)} & \text{if } w_j^{(k)}(t) = c_i(t) \\ 0 & \text{otherwise} \end{cases} \tag{6.35}$$

where $b_j^{(k)}$ is the kth user's transmitted bit by the jth transmit antenna, $w_j^{(k)}(t)$ is the spreading sequence used by the kth user on its jth transmit antennas and $c_i(t)$ is the ith spreading sequence of the set of Q mutually spreading sequences assigned to the kth user. Depending on whether parity bit selected or permutation spreading technique is used, spreading waveforms used by each transmit antenna is selected.

We can extend (6.34) to multiuser scenario as

$$\underline{Z} = \underline{W} \ \underline{R} + \underline{N} \tag{6.36}$$

where

$$\underline{Z} = [\underline{Z}^{(1)^T}, \underline{Z}^{(2)^T}, \cdots, \underline{Z}^{(K)^T}]; \tag{6.37}$$

$$\underline{W} = [\underline{W}^{(1)^T}, \underline{W}^{(2)^T}, \cdots, \underline{W}^{(K)^T}]; \tag{6.38}$$

$$\underline{W}^{(k)} = \underline{Y}^{(k)}\ \underline{H}^{(k)}; \tag{6.39}$$

$$\underline{N} = [\underline{N}^{(1)^T}, \underline{N}^{(2)^T}, \cdots, \underline{N}^{(K)^T}]; \tag{6.40}$$

and $\underline{\mathrm{R}}$ is the $KQ \times KQ$ matrix of cross-correlation whose $(i \circ j, m \circ n)$th element is the cross-correlation between the ith spreading code of the jth user and the mth spreading code of the nth user.

The proposed turbo multiuser receiver is based on iterative soft interference cancellation. Using the *extrinsic* information provided by the channel decoder for every bit of every user in the previous iteration, MAI caused by all users are estimated and subtracted out from the matched filter outputs.

To be able to estimate MAI in each iteration, we first calculate the estimation of matrix $\underline{Y}^{(k)}$ using the *extrinsic* information of the previous iteration. Following the same derivations as shown in (6.19) to (6.23), the estimated matrix of $\underline{\tilde{Y}}^{(k)}$ can be constructed as

$$\tilde{y}_{i,j}^{(k)} = \sum_{\underline{b}^{(k)} \in S_i} b_j^{(k)} \prod_{m=1}^{N_r} \frac{1}{2} \left[1 + b_m^{(k)} \tanh\left(\frac{1}{2} \lambda_c^p \left(b_m^{(k)} \right) \right) \right]. \tag{6.41}$$

where the choice of S_i depends on the spreading code selection technique in use. By having the estimated matrix of $\underline{\tilde{Y}}^{(k)}$, the estimated matrices of $\underline{\tilde{W}}^{(k)}$ and $\underline{\tilde{W}}$ can be consequently calculated as

$$\underline{\tilde{W}}^{(k)} = \underline{\tilde{Y}}^{(k)}\ \underline{H}^{(k)}; \tag{6.42}$$

$$\underline{\tilde{W}} = [\underline{\tilde{W}}^{(1)^T}, \underline{\tilde{W}}^{(2)^T}, \cdots, \underline{\tilde{W}}^{(K)^T}]. \tag{6.43}$$

Using the estimated matrix of $\underline{\tilde{W}}$, MAI caused by all users can be estimated and subtracted out from the outputs of the bank of matched filters:

$$\underline{\tilde{Z}} = \underline{Z} - \underline{\tilde{W}}(\underline{R} - \overline{I}) \tag{6.44}$$

where $\underline{\tilde{Z}}$ is the matched filter outputs after MAI estimation and subtraction and $\overline{I}$ is the $KQ \times KQ$ identity matrix. Matrix $\underline{\tilde{Z}}$ is considered as the matrix of matched filter outputs to be used in the following iteration.

6.6.1 Simulation Results

In this section we present some simulation results that illustrate the performance of the proposed multiuser turbo receiver in MIMO-CDMA systems that employ parity bit selected and permutation spreading. The same system parameters as of section 5.4 are considered. All users employ the same 1/2 rate convolutional code with constraint length of 5 and generators 23, 35 in octal notation. The simulations are performed for the case of synchronous MIMO-CDMA systems. The channel is modeled as slowly-varying, frequency nonselective, Rayleigh fading channel, where the channel gains of different transmit-receive links are uncorrelated. We assume that perfect channel estimation is available at the base station without a power penalty. We assume that every user is assigned with a set of mutually orthogonal spreading sequences and also that the cross-correlation between any two spreading sequence of different users are the same. In all the simulations of this section, a cross-correlation of $\rho = 0.3$ is assumed between any spreading sequence of different users. The simulations are performed for the MIMO-

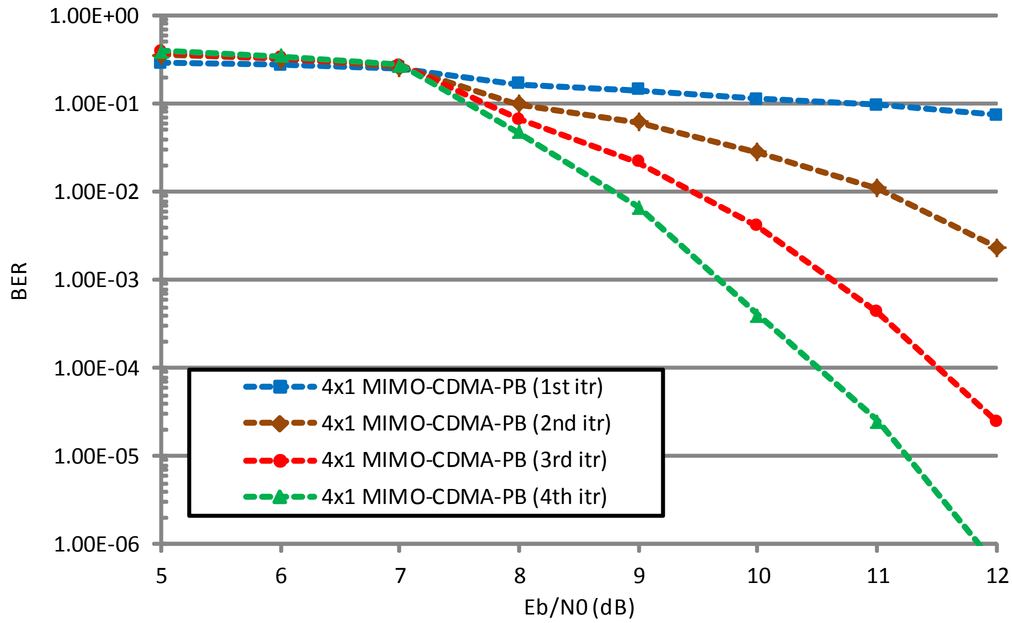

Figure 6.9: Performance of the turbo multiuser receiver in a 4×1 MIMO-CDMA system that employs parity bit selected spreading. $K = 4$, $\rho = 0.3$.

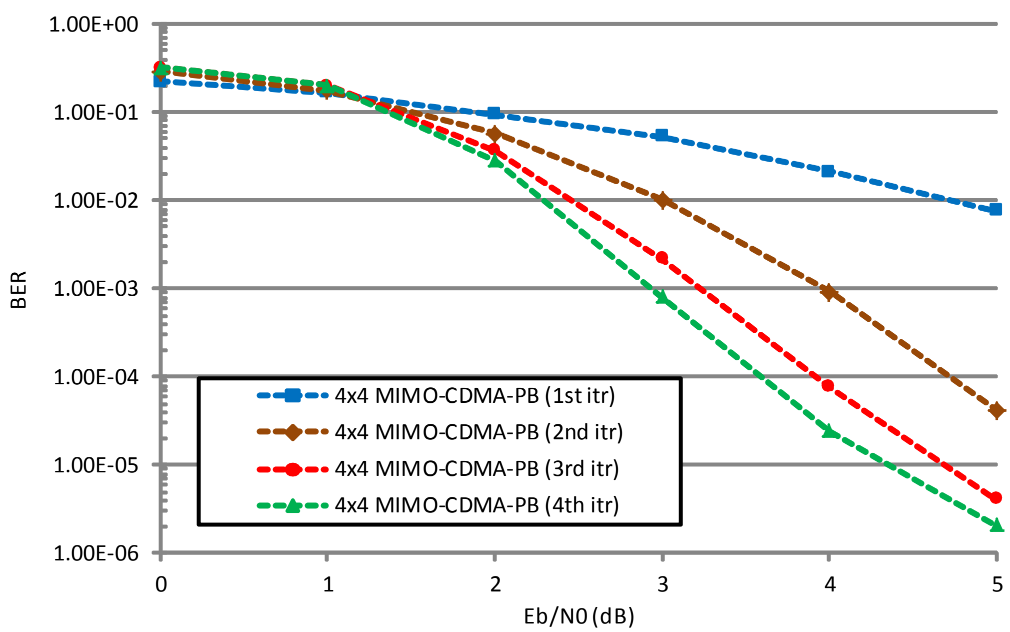

Figure 6.10: Performance of the turbo multiuser receiver in a 4×4 MIMO-CDMA system that employs parity bit selected spreading. $K = 4$, $\rho = 0.3$.

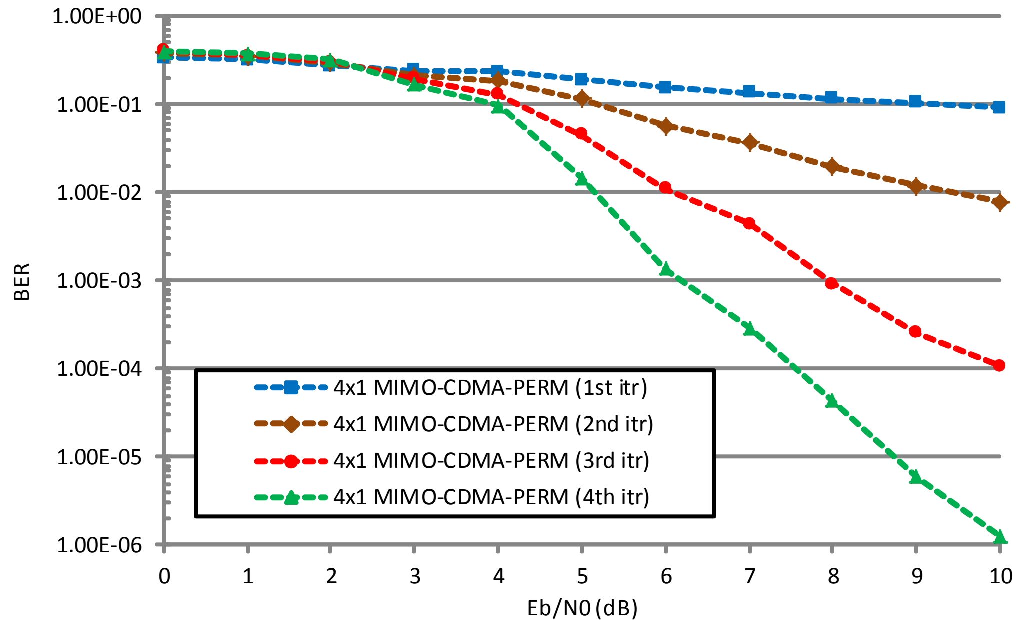

Figure 6.11: Performance of the turbo multiuser receiver in a 4×1 MIMO-CDMA system that employs permutation spreading. $K = 4$, $\rho = 0.3$.

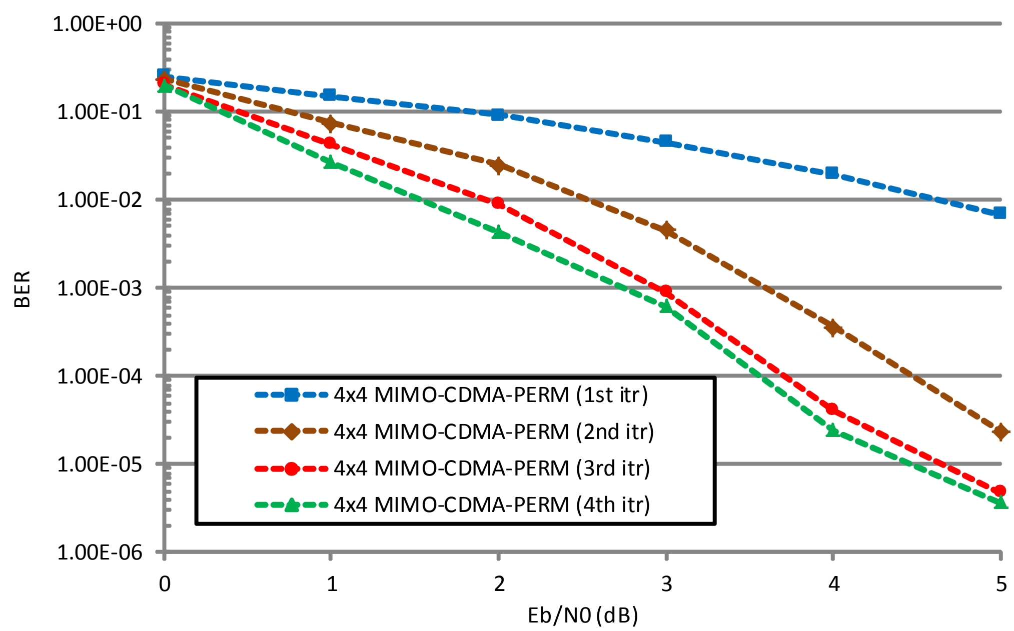

Figure 6.12: Performance of the turbo multiuser receiver in a 4×4 MIMO-CDMA system that employs permutation spreading. $K = 4$, $\rho = 0.3$.

CDMA system with $K = 4$ users where each user has $N_t = 4$ transmit antennas and the base station is equipped with $N_r = 1$ or 4 receive antennas.

The BER performance of the proposed multiuser receiver for the MIMO-CDMA systems that employ parity bit selected spreading sequences with $K = 4$ users and $N_r = 1$ and 4 receive antennas are shown in Figures 6.9 and 6.10 respectively. The parity bit selected spreading technique used in these simulations is based on the system explained in section 2.2.4 and it follows (2.13) to select the spreading sequence for each transmit antenna.

Figures 6.11 and 6.12, respectively, present the BER performance of the proposed turbo multiuser receiver for MIMO-CDMA systems with $N_r = 1$ and 4 receive antennas that employ permutation spreading technique. The selection of spreading waveform for each transmit antenna in such systems follows Table 2.1.

From these figures, it can be seen that in both MIMO-CDMA systems that employ either parity bit selected or permutation spreading technique, a significant BER performance is achieved when the proposed multiuser detection is applied over multiple iterations.

6.7 Discussion

In this chapter, a turbo-based multiuser receiver for different CDMA systems that employ parity bit selected spreading is discussed. The optimum SISO multiuser detection for CDMA-PB systems is derived whose complexity grows exponentially with the number of users and the number of spreading codes used by each user.

A low-complexity SISO multiuser detector for CDMA-PB systems is proposed based on soft interference cancellation. In such systems, the MAI generated by every user is estimated based on the LLRs provided by the SISO channel decoder in the previous iteration for every coded bits of every user. The estimated MAI is then subtracted out from the matched filter outputs in each iteration. This turbo multiuser receiver is proposed and studied for both synchronous and asynchronous channels.

This low-complexity SISO multiuser detector is further developed for MC-CDMA systems and MIMO-CDMA systems that employ parity bit selected and permutation spreading.

Performed simulations for several CDMA systems that employ parity bit selected and permutation spreading demonstrate a significant performance improvement when the proposed turbo multiuser receiver is used over multiple iterations.

Chapter 7

Conclusions and Suggestions for Further Research

7.1 Conclusions

In this thesis, we have proposed turbo receivers for different SS-PB systems. For each specific SS-PB system, an algorithm for the SISO detection is developed in which the LLRs of each detected bit is calculated based on the received signal, the a priori information from the previous iteration and the code structure of the SS-PB system.

In multiuser scenario, a method of allocating the sets of biorthogonal spreading sequences to each user is explained. An optimal algorithm for multiuser detection in SS-PB systems is discussed and it is shown that the optimal detection is prohibitively complex when the number of users increase.

A low complexity multiuser detection is developed in which at first detection is performed separately for each user and then using the LLRs of every bit of every user, MAI is estimated and subtracted out from the received signal. Simulations show a significant performance improvement just after a couple of iterations.

An algorithm for SISO multiuser detection is further developed for MC-CDMA and MIMO-CDMA systems that employ parity bit selected and permutation spreading. Simulation results confirm that even in a multiuser scenario, permutation spreading technique outperforms the parity bit selected technique in MIMO-CDMA systems.

7.2 Suggestions for Further Research

- Throughout this thesis we assumed that a perfect channel estimation is available and that the receiver has the perfect knowledge of the channel gains without loss of any energy. This assumption is not realistic and further investigation is needed to study the effect of channel estimation error on the performance of the proposed systems.

- An adaptive turbo multiuser detector can be developed in which the extrinsic information of each data bit is used in order to estimate the channel gains.

- EXIT chart analysis can be performed on the proposed turbo receivers to better understand and analyze the performance of the system.

- In the case of multiple users of SS-PB systems, an analysis can be done to see the effect of MAI in the performance of SS-PB systems.

- The application of parity bit selected technique in OFDM systems can be investigated where similar to the SS-PB systems, the parity bits are used to select the sub carriers used to transmit the signal.

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