INTERFERENCE CANCELLATION SCHEMES FOR STBC MULTIUSER SYSTEMS

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SUMMARY

Next generation wireless communication systems are expected to provide a variety of services integrating voice, data and video. The rapidly growing demand for these services needs high data rate wireless communication systems with high user capacity. In addition, due to difference in source rate of services, designing a multirate communication system is also an imperative. In order to meet this goal, research efforts are carried out to develop efficient coding and modulation schemes along with sophisticated signal and information processing algorithms to improve the quality and spectral efficiency of wireless links. However, these developments must cope with critical performance limiting challenges that include multipath fading, multiuser interference, power and size of mobile units.

Recently, it has been shown that achievable data rate of wireless communication systems increases dramatically by employing multiple transmit and receive antennas. Employing multiple transmit antennas is feasible in mobile communication system because multiple transmit antennas can be deployed at the base station to improves the downlink performance of the system. Due to less encoding and decoding complexity, STBC is very popular among transmit diversity systems. Most of the STBC schemes relies on that the channel state information available at the receiver. This assumption is not valid in real situations. As a result, channel estimation techniques are more important to implement the STBC schemes. On the other hand, all existing multiuser STBC systems consider equal rate user environment that motivates to explore the system in multirate multiuser environment.

To overcome above-mentioned problems, two novel receivers are proposed. Firstly, An adaptive receiver is proposed to mitigate multiuser interference without any explicit knowledge of channel state information. In addition, the proposed adaptive receiver works without any knowledge of the interferer. Secondly, an iterative interference cancellation receiver for both equal rate and multi rate space-time block coded multiuser systems is presented. Various simulation results demonstrate that both receivers show less computational complexity and better BER performance than that of existing schemes.

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NOMENCLATURE

3G	Third Generation
8-PSK	Eight Phase Shift Keying
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CDMA	Code Division Multiple Access
CP	Cyclic Prefix
dB	Decibel
DSTBC	Differential Space-Time Coding
FDE	Frequency Domain Equalization
FFT	Fast Fourier Transform
FIR	Finite Impulse Response
GSM	Global System for Mobile
IC	Interference Cancellation
IFFT	Inverse Fast Fourier Transform
IIC	Iterative Interference Cancellation
iid	Independent and Identically Distributed
IS-136	US-TDMA, one of the 2^{nd} generation mobile phone systems
IS-54	D-AMPS, digital advanced mobile phone system
LMS	Least Mean Square
LSE	Least Square Error
ML	Maximum Likelihood

MMS	Multi Media Messaging
MMSE	Minimum Mean Square Error
OFDM	Orthogonal Frequency Division Multiplexing
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
QR-RLS	QR decomposition based RLS
RLS	Recursive Least Square
Rx	Receive Antenna
SIR	Signal to Interference Ratio
SNR	Signal to Noise Ratio
ST	Space-Time
STBC	Space-Time Block Coding
STC	Space-Time Coding
STTC	Space-Time Trellis Coding
TDMA	Time Division Multiple Access
Tx	Transmit Antenna
WCDMA	Wideband CDMA
w.r.t.	With respect to
ZF	Zero Forcing

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CHAPTER 1

INTRODUCTION

1.1 Background

The next generation broadband wireless communication systems are expected to provide wireless multimedia services such as high-speed Internet access, multimedia message services (MMS) and mobile computing. The rapidly growing demand for these services needs high data rate wireless communication systems with high user capacity. In order to meet this goal, research efforts are carried out to develop efficient coding and modulation schemes along with sophisticated signal processing techniques to improve the quality and spectral efficiency of wireless communication links. However, these developments must cope with critical performance limiting challenges that include multipath fading, multiuser interference, power and size of mobile units [1,2].

Mobile communication channels are subject to frequency selective and time selective fading that are induced by multipath propagation, phase shifts and Doppler shifts in the signal. The most appropriate way to combat the fading is the exploitation of diversity. The diversity techniques are based on supplying several replicas of the same information bearing signal to the receiver over independently fading channels. There are several ways to provide the independently fading replicas of the information-bearing signal to the receiver. Among those, temporal diversity, frequency diversity and space diversity [3] are three main techniques that are widely used in the communication systems. To provide the temporal diversity, channel coding is used with appropriate interleaving method. The frequency diversity normally introduces redundancy in the frequency domain by transmitting the same information bearing signal over multiple carriers. By deploying multiple transmit and receive antennas, which are separated and/or polarized to create independent fading, space diversity can be achieved.

Recently, it has been shown that achievable data rate of wireless communication systems increases dramatically by employing multiple transmit and receive antennas [4-6]. In these schemes, it is assumed that the complex-valued propagation coefficients between all pairs of transmit and receive antennas are statistically independent and perfectly known at the receiver. Independent channel coefficients are obtained by placing transmit and receive antennas a few wavelengths apart from each other. Because of wide antenna separation, the traditional adaptive array concepts of beam forming and directivity cannot be applied to these systems. Depending on whether multiple antennas are used for transmission or reception, diversity is classified as transmit antenna diversity and receive antenna diversity. In receive antenna diversity schemes, multiple receive antennas are deployed at the receiver to receive multiple copies of transmitted signal which are then properly combined to mitigate the channel fading. In fact, receive antenna diversity schemes have been incorporated with the existing second-generation mobile communication systems such as GSM and IS-136 to improve the mobile to base station transmission (uplink) [7]. Due to the size and power limitations of the mobile unit, it is not feasible to deploy the multiple receive antennas at the mobile unit. As a result, the base station to mobile transmission (downlink) has some bottleneck in the current mobile communication systems. This has motivated the rapidly growing research on transmit antenna diversity.

Transmit antenna diversity is feasible in mobile communication system because multiple transmit antennas can be deployed at the base station and that improves the downlink performance of the system. A number of transmit antenna diversity schemes has been proposed and that can be divided under two main categories such as open loop and closed loop transmit antenna diversities. The open loop scheme transmitter does not require any channel information from the receiver [8-11]. On the other hand, the closed loop scheme transmitter relies on the channel information provided by the receiver via a feedback [8,11,12]. In fast mobility situations, the transmitter may not be capable of capturing the channel variations. As a result, the usage of open loop schemes is motivated for future wireless communication systems, which are characterized by high mobility.

Among open loop transmit diversity schemes space-time coding (STC) is very popular. STC relies on multiple antenna transmissions and suitable signal processing technique at the receiver to provide the coding and diversity gains. The STC includes space-time trellis coding (STTC) [13] and space-time block coding (STBC) [14-17]. Space-time trellis coding (STTC) provides better tradeoff among constellation size, data rate, diversity advantage, coding advantage, and trellis complexity. When the number of transmit antennas is fixed, the decoding complexity of STTC increases exponentially with the diversity order and the transmission rate. To reduce the decoding complexity at the receiver, Alamouti [14] proposed a remarkable scheme called space-time block coding (STBC) for two transmit antennas. It was further extended in [16] for three and four transmit antennas using the theory of orthogonal designs. When the channel state information is available at the receiver, the STBC proposed in [14-17] uses a maximum likelihood detection scheme based on linear processing and is able to achieve the full diversity promised by the transmit and receive antennas. Recently, STC has been adopted in third generation (3G) cellular standards and proposed in many wireless applications [8,10,11,18,19,20].

A differential detection scheme has been proposed in [21] for two transmit antenna system, where channel state information is not required at the receiver. The scheme described in [21] has been further extended to multiple transmit antennas in [22]. Due to the differential detection, the bit error rate (BER) performance of the schemes described in [21] and [22] is 3 dB worse than that of the scheme proposed in [14] and [16], respectively. A new detection schemes for STBC without channel estimation but not fully differential scheme is proposed in [23]. This scheme has more than 3 dB penalty compared to coherent detection in [14]. In [24], a new and general approach to differential modulation for multiple antennas based on group codes was presented. This approach can be applied to any number of antennas, and any signal constellation. A non-differential approach to transmit diversity when the channel is unknown is reported in [25]. But the schemes described in [24] and [25] have both exponential encoding and decoding complexities.

So far, most of the schemes assume that channel is frequency non-selective. Unfortunately, in the broadband wireless communication systems, the channel response over the occupied bandwidth is generally neither flat in frequency nor static across time. In broadband wireless communication systems symbol duration is smaller than the channel delay spread which results the channel to be frequency selective. It is important to investigate the STC in the frequency selective environments. The effects of multipath on the performance of STTC are studied in [26] for a slowly varying channel. Furthermore, it is proved that the presence of multipath does not decrease the diversity order of STTC. Results in [26] suggest that STTC in frequency selective fading channels may achieve higher diversity advantage than those in flat fading channels providing that the channel state information is available at the receiver. However, the scheme works based on ML decoding that is computationally expensive. In this situation, feasible STC designs for frequency selective channels are important.

Since the introduction of STBC by Alamouti in [14], the following transmissions schemes have been proposed to extend the STBC to frequency selective channels. One approach, with lower receiver complexity, is to employ orthogonal frequency division multiplexing (OFDM), which converts the frequency selective channel into a set of flat fading sub channels through inverse fast Fourier transform (IFFT) and cyclic prefix (CP) insertion at the transmitter. CP removal and fast Fourier transform (FFT) are used at the receiver to decode the signals. In each flat fading sub channels, STBC have been applied on each OFDM sub carrier [26-28]. Space-time codes designed for flat-fading channels can be utilized to achieve full multi antenna diversity in frequency selective fading channels [26]. But, the potential diversity gain available in multipath propagation has not been addressed. Recently, in OFDM based systems, it was produced in [30-33] that it is possible to achieve both multi antenna diversity and multipath diversity gains that is equal to the product of number of transmit antennas, number of receive antennas and number of finite impulse response (FIR) channel taps.

Limitation of all multi carrier (OFDM) based STC transmissions is their non-constant modulus, which necessitates power amplifier back off, and thus reduces power efficiency. In addition, multi carrier schemes are more sensitive to inter carrier interference. Above factors motivate the importance of having single carrier STBC transmission schemes [34-40] for frequency selective channels. These single carrier based schemes are based on frequency domain equalization (FDE) of received signal at the receiver that can be employed by taking FFT and then equalizing the received signal block. Finally, equalized received signals go through IFFT block and then decision is made in the time domain using a slicer as described in [34]. A generalized transmission scheme for single carrier STBC is proposed in [38-40] that subsume those in [34-37] as special cases. Furthermore, it is proved in [34] that the diversity order is $N_t N_r (L+1)$, where N_t is the number of transmit antennas, N_r is the number of receive antennas and L is the number of taps in the FIR filter model of the channel.

Co-channel interference is generally recognized as one of the factors that limit the capacity and the transmission quality in wireless communications. An appropriate understanding of the interference is extremely important when analyzing and designing multiuser wireless systems or exploring techniques that mitigate the undesirable effects of co-channel users [41-43]. By exploiting the spatial and temporal structure of the STC, interference from the co-channel users can be suppressed. An interference cancellation scheme for STBC multiuser systems with 2 synchronous co-channel users, each is equipped with 2 transmit antennas, has been presented for flat fading channels in [44]. This scheme is based on minimum mean square error (MMSE) and ML detection at the receiver. Further, this scheme has been extended in [45] for more than two synchronous co-channel users; where each user is equipped with more than 2 transmit antennas. The effectiveness of the interference suppression of these STBC co-channel multiuser systems relies on accurate channel estimates available at the receiver. An interference cancellation method based on zero forcing (ZF) method is presented in [46] for single carrier STBC-FDE [32] two-user system.

Most of above schemes relies on the fact that the channel state information is available at the receiver. This assumption is not valid in real situations. As a result, channel estimation techniques are more important I the implementation of the above schemes. A joint channel estimation and co-channel interference suppression scheme has been presented in [47,48] for wireless time division multiple access (TDMA) systems equipped with multiple transmits and receives antennas in frequency selective channel. This scheme is able to mitigate interference of various origins, including inter-symbol interference and co-channel interference.

An effort has been taken in this thesis to present an overview of interference cancellation schemes in space-time coded multiuser systems and to specifically solve the following problems:

1. Jointly estimate the channel and suppress the interference of STBC multiuser systems by using an adaptive receiver scheme that operates based on least square error (LSE) and recursive least square (RLS) signal processing techniques.
2. Above problem is extended for multirate multiuser systems where an iterative interference cancellation (IIC) scheme is used based on minimum mean square error (MMSE) and maximum likelihood (ML) techniques.

1.2 Outline and Contributions of the Thesis

Chapter 2 presents overview of space-time block coding schemes. Firstly, space-time coding schemes for two transmit antennas and multiple transmit antennas are described. Secondly, Importance of a differential detection scheme for space-time block coding is analyzed and presented. Finally, comparisons of simulation results are given.

Chapter 3 sets up the framework needed to introduce the new interference cancellation schemes for space-time block coded multiuser systems. It starts with a review of statistical and adaptive signal processing concepts such as MMSE, LMS, LSE and RLS. Then, existing interference cancellation techniques for STBC multiuser systems based on MMSE are described. Overview of signal processing techniques and existing interference cancellation schemes simplify the understanding of the derivation of proposed schemes in the following chapters.

An adaptive receiver is presented in chapter 4. In this adaptive receiver, multiuser interference is cancelled without any explicit knowledge of channel state information. The computational complexity of the proposed adaptive receiver is less than that of the MMSE interference cancellation scheme. On the other hand, due to the weight estimation errors, the BER performance of the adaptive receiver is little bit worse than the BER performance of the MMSE interference cancellation scheme. It is also noted that the proposed adaptive receiver works without any explicit knowledge of the channel and interferer. Finally, chapter 4 presents the application of adaptive receiver scheme for multirate multiuser systems.

Iterative interference cancellation scheme (IIC) for both equal rate and multi rate space-time block coded multiuser systems is presented in chapter 5. The proposed IIC scheme is derived based on MMSE interference cancellation and ML decoding. The IIC scheme outperforms the conventional MMSE interference cancellation schemes. In addition, it is shown that the BER performance of IIC multirate systems is better than that of IIC equal rate systems. The simulation results and comparisons show the effectiveness of the proposed scheme.

Chapter 6 summarizes the thesis and discusses the possible extensions and directions of future research. In particular, some aspects and open problems pertaining to the interference cancellation and channel estimation of the STBC systems in frequency selective channels are discussed.

Some results presented in this thesis can be found in the following publications.

V. Mahinthan, B. Kannan, and A. Nallanathan, “Joint channel estimation and interference cancellation in space-time coded multiuser systems,” Proc. IEEE Mobile Wireless Communications Networks, Stockholm, Sweden. Sep. 2002.

V. Mahinthan, B. Kannan, and A. Nallanathan, “Adaptive channel estimation and interference suppression in space-time coded multiuser systems,” Proc. IEEE GLOBECOM, Taipei, Taiwan. Nov. 2002.

V. Mahinthan, B. Kannan, and A. Nallanathan, “Performance of LSE-RLS based interference cancellation scheme for STBC multiuser systems,” IEE Electronic Letters, Dec. 2002, Vol. 38, No. 25, pp. 1729-1730

V. Mahinthan, B. Kannan, and A. Nallanathan, “An adaptive receiver for space-time block coded multiuser systems,” Submitted to the IEEE Transactions on Communications.

V. Mahinthan, B. Kannan, and A. Nallanathan, “An Iterative Interference Cancellation Scheme for STBC Multirate Multiuser Systems,” Submitted to IEEE Communication Letters.

CHAPTER 2

OVERVIEW OF SPACE-TIME BLOCK CODES

Space-time block coding has been introduced recently and has sparked a wide interest among the research communities as it promises to increase transmission rates significantly in wireless communications. In this chapter, the system model and receiver structure of the STBC and DSTBC are reviewed. In addition, advantages and inherent problems of STBC and DSTBC schemes are discussed.

2.1 Introduction

Alamouti [14] proposed a simple transmit diversity scheme called space-time block codes, which fully exploits the spatial diversity offered by two transmit and multiple receive antennas and improves the overall performance of the wireless communication systems. It was further extended in [16] for three and four antennas using the theory of orthogonal designs. When channel state information is available at the receiver, the STBC proposed in [14-16] uses a maximum likelihood detection based on linear processing at the receiver. The differential detection schemes have been proposed in [21] and [22] for the STBC schemes presented in [14] and [16] respectively, where channel state information are not required at the receiver. But the BER performances of the differential detection schemes are 3 dB worse than that of respective coherent schemes.

2.2 Space-Time Block Coding

2.2.1 Two Transmit and One Receive Antenna System

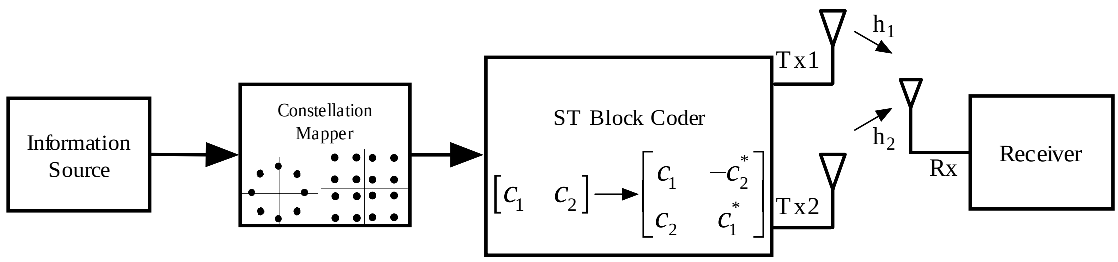

Fig. 2.1 Space-time block coded transmit diversity system for two transmit and one receive antennas

Consider the transmit diversity scheme with 2 transmit and 1 receive antennas as shown in Fig. 2.1, where two transmit antennas are well separated to get independent fading. At a given symbol period T_s two digitally modulated signals are simultaneously transmitted from the two antennas. As given in the Table 2.1, symbols c_1 and c_2 are transmitted from antenna 1 and antenna 2 respectively. During the next symbol period signal $-c_2^*$ is transmitted from antenna 1, and signal c_1^* is transmitted from antenna 2. Let h_1 and h_2 be the channel fading coefficient from first and second transmit antenna to the receive antenna. The received signal over two consecutive symbol periods is defined as r_1 and r_2.

Table 2.1 Space-time block codes for two transmit antenna system

Transmit Antenna	Time t	Time $t+T_s$
Tx 1	c_1	$-c_2^*$
Tx 2	c_2	c_1^*

The received signal can be written as,

$$r_1 = h_1 c_1 + h_2 c_2 + \eta_1 \tag{2.1}$$

$$r_2 = -h_1 c_2^* + h_2 c_1^* + \eta_2 \tag{2.2}$$

where η_1 and η_2 represent the zero mean complex gaussian noise with variance $1/(2SNR)$ per complex dimension. The transmitted symbols have average energy (1/2) per antenna. The channel fading coefficients are assumed to be constant over two consecutive symbol periods.

Defining the received signal vector, $\boldsymbol{r} = \begin{bmatrix} r_1 & r_2^* \end{bmatrix}^T$, the code symbols vector $\boldsymbol{c} = \begin{bmatrix} c_1 & c_2 \end{bmatrix}^T$ and noise vector $\boldsymbol{\eta} = \begin{bmatrix} \eta_1 & \eta_2^* \end{bmatrix}^T$, equations (2.1) and (2.2) can be written as

$$\boldsymbol{r} = \boldsymbol{H}\boldsymbol{c} + \boldsymbol{\eta} \tag{2.3}$$

where channel matrix $\boldsymbol{H}$ is defined as

$$\boldsymbol{H} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix} \tag{2.4}$$

Hence, channel matrix $\boldsymbol{H}$ is orthogonal $\boldsymbol{H}^*\boldsymbol{H} = (|h_1|^2 + |h_2|^2)\boldsymbol{I}_2$.

By multiplying the receive signal by conjugate transpose of the channel matrix,

$$\tilde{\boldsymbol{r}} = \boldsymbol{H}^{*}\boldsymbol{r} = (|h_1|^2 + |h_2|^2)\boldsymbol{c} + \tilde{\boldsymbol{\eta}} \tag{2.5}$$

where $\tilde{\boldsymbol{\eta}} = \boldsymbol{H}^{*}\boldsymbol{\eta}$.

By the maximum likelihood rule, the decoded symbol code vector can be written as,

$$\hat{\boldsymbol{c}} = \arg\min_{\hat{\boldsymbol{c}}\varepsilon\boldsymbol{C}} \left\{ \left\| \tilde{\boldsymbol{r}} - (|h_1|^2 + |h_2|^2)\hat{\boldsymbol{c}} \right\|^2 \right\} \tag{2.6}$$

2.2.2 Two Transmit and Multiple Receive Antenna System

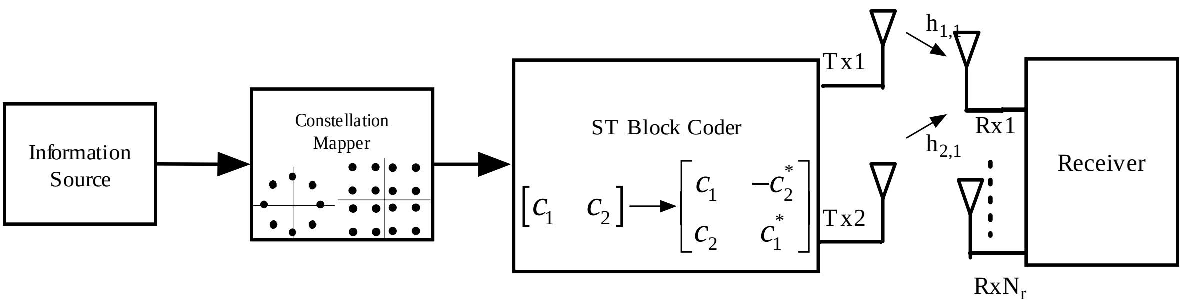

Fig. 2.2 Space-time block coded transmit diversity system for two transmit and multiple receive antennas

Consider the transmit diversity scheme with 2 transmit and N_r receive antennas as shown in Fig. 2.2 where both transmit and receive antennas are well separated to get independent fading. As written in (2.3), the received signal vector $\boldsymbol{r}_m$ at m^{th} receive antenna can be written as,

$$\boldsymbol{r}_m = \boldsymbol{H}_m\boldsymbol{c} + \boldsymbol{\eta}_m \tag{2.7}$$

where $\boldsymbol{\eta}_m$ is the noise vector at the m^{th} receive antenna and

$$\boldsymbol{H}_m = \begin{bmatrix} h_{1,m} & h_{2,m} \\ h_{2,m}^* & -h_{1,m}^* \end{bmatrix} \tag{2.8}$$

Hence, channel matrix $\boldsymbol{H}_m$ is orthogonal $\boldsymbol{H}_m{}^*\boldsymbol{H}_m = (\left|h_{1,m}\right|^2 + \left|h_{2,m}\right|^2)\boldsymbol{I}_2$.

By multiplying the receive signal by conjugate transpose of the channel matrix,

$$\tilde{\boldsymbol{r}}_m = \boldsymbol{H}_m{}^*\boldsymbol{r}_m = (\left|h_{1,m}\right|^2 + \left|h_{2,m}\right|^2)\boldsymbol{c} + \tilde{\boldsymbol{\eta}}_m \tag{2.9}$$

where $\tilde{\boldsymbol{\eta}}_m = \boldsymbol{H}_m{}^*\boldsymbol{\eta}_m$.

By the maximum likelihood rule, the decoded symbol code vector can be written as,

$$\hat{\boldsymbol{c}} = \arg\min_{\hat{c}\varepsilon\boldsymbol{C}} \left\{ \sum_{m=1}^{N_r} \left\| \tilde{\boldsymbol{r}}_m - (\left|h_{1,m}\right|^2 + \left|h_{2,m}\right|^2)\hat{\boldsymbol{c}} \right\|^2 \right\} \tag{2.10}$$

In the case of N_r receive antennas, $2N_r$ diversity can be achieved as shown in (2.10).

2.2.3 Multiple Transmit and Multiple Receive Antenna System

The space-time block coding transmission matrix $\boldsymbol{G}_{N_t}$ for more than two transmit antenna system is designed based on the classical mathematical framework of orthogonal designs [16]. However, STBC can be constructed using orthogonal designs for few sporadic values of N_t (number of transmit antennas). It is shown in [16] that STBC for any signal constellations and any numbers of transmit antennas can be generated by the generalization of orthogonal designs.

As stated in the literature, a $p \times N_t$ transmission matrix $\boldsymbol{G}_{N_t}$ defines the STBC. The entries of the real transmission matrix are positive or negative values of c_l where

$l = 1, 2....L$. On the other hand, the entries of the complex transmission matrix are linear combination of the variables c_l and their conjugates. The rate S of the code is defined as $S = L / p$.

2.2.3.1 Orthogonal Designs for Real Signal Constellations

The real orthogonal designs only exist for small number of dimensions. A real orthogonal design of size N_t is an $N_t \times N_t$ orthogonal transmission matrix which exists only for $N_t = 2, 4, 8$. Some examples are given in Table 2.2. The limitations of providing transmit diversity through linear processing based on orthogonal square transmission matrix can be subdued by the generalized orthogonal designs [16]. The non-square orthogonal transmission matrix of generalized orthogonal designs for $N_t = 3, 5, 6, 7$ is given in Table 2.3.

Table 2.2 Real orthogonal transmission matrices for $N_t = 2, 4, 8$

$N_t = 2$	$N_t = 4$	$N_t = 8$
$\begin{pmatrix} c_1 & c_2 \\ -c_2 & c_1 \end{pmatrix}$	$\begin{pmatrix} c_1 & c_2 & c_3 & c_4 \\ -c_2 & c_1 & -c_4 & c_3 \\ -c_3 & c_4 & c_1 & -c_2 \\ -c_4 & -c_3 & c_2 & c_1 \end{pmatrix}$	$\begin{pmatrix} c_1 & c_2 & c_3 & c_4 & c_5 & c_6 & c_7 & c_8 \\ -c_2 & c_1 & c_4 & -c_3 & c_6 & -c_5 & -c_8 & c_7 \\ -c_3 & -c_4 & c_1 & c_2 & c_7 & c_8 & -c_5 & -c_6 \\ -c_4 & c_3 & -c_2 & c_1 & c_8 & -c_7 & c_6 & -c_5 \\ -c_5 & -c_6 & -c_7 & -c_8 & c_1 & c_2 & c_3 & c_4 \\ -c_6 & c_5 & -c_8 & c_7 & -c_2 & c_1 & -c_4 & c_3 \\ -c_7 & c_8 & c_5 & -c_6 & -c_3 & c_4 & c_1 & -c_2 \\ -c_8 & -c_7 & c_6 & c_5 & -c_4 & -c_3 & c_2 & c_1 \end{pmatrix}$

2.2.3.2 Orthogonal Designs for Complex Signal Constellations

The complex orthogonal designs only exist for $N_t = 2$. The STBC proposed by Alamouti [14] uses the complex orthogonal design for $N_t = 2$. To overcome the barrier of complex orthogonal designs, a generalized complex orthogonal design is proposed in [16]. Furthermore, it is proved that a half rate generalized complex orthogonal transmission matrices exist for any numbers of transmit antennas. Some examples are given in Table 2.4. Specially, ¾ rate generalized complex orthogonal transmission matrices are provided for three and four transmit antennas in Table 2.5.

Table 2.3 Real generalized orthogonal transmission matrices for $N_t = 3,5,6,7$

$N_t = 3$	$N_t = 5$
$\begin{pmatrix} c_1 & c_2 & c_3 \\ -c_2 & c_1 & -c_4 \\ -c_3 & c_4 & c_1 \\ -c_4 & -c_3 & c_2 \end{pmatrix}$	$\begin{pmatrix} c_1 & c_2 & c_3 & c_4 & c_5 \\ -c_2 & c_1 & c_4 & -c_3 & c_6 \\ -c_3 & -c_4 & c_1 & c_2 & c_7 \\ -c_4 & c_3 & -c_2 & c_1 & c_8 \\ -c_5 & -c_6 & -c_7 & -c_8 & c_1 \\ -c_6 & c_5 & -c_8 & c_7 & -c_2 \\ -c_7 & c_8 & c_5 & -c_6 & -c_3 \\ -c_8 & -c_7 & c_6 & c_5 & -c_4 \end{pmatrix}$
$N_t = 6$	$N_t = 7$
$\begin{pmatrix} c_1 & c_2 & c_3 & c_4 & c_5 & c_6 \\ -c_2 & c_1 & c_4 & -c_3 & c_6 & -c_5 \\ -c_3 & -c_4 & c_1 & c_2 & c_7 & c_8 \\ -c_4 & c_3 & -c_2 & c_1 & c_8 & -c_7 \\ -c_5 & -c_6 & -c_7 & -c_8 & c_1 & c_2 \\ -c_6 & c_5 & -c_8 & c_7 & -c_2 & c_1 \\ -c_7 & c_8 & c_5 & -c_6 & -c_3 & c_4 \\ -c_8 & -c_7 & c_6 & c_5 & -c_4 & -c_3 \end{pmatrix}$	$\begin{pmatrix} c_1 & c_2 & c_3 & c_4 & c_5 & c_6 & c_7 \\ -c_2 & c_1 & c_4 & -c_3 & c_6 & -c_5 & -c_8 \\ -c_3 & -c_4 & c_1 & c_2 & c_7 & c_8 & -c_5 \\ -c_4 & c_3 & -c_2 & c_1 & c_8 & -c_7 & c_6 \\ -c_5 & -c_6 & -c_7 & -c_8 & c_1 & c_2 & c_3 \\ -c_6 & c_5 & -c_8 & c_7 & -c_2 & c_1 & -c_4 \\ -c_7 & c_8 & c_5 & -c_6 & -c_3 & c_4 & c_1 \\ -c_8 & -c_7 & c_6 & c_5 & -c_4 & -c_3 & c_2 \end{pmatrix}$

Table 2.4 – Unit and Half rate complex generalized orthogonal transmission matrices

$N_t = 2$	$N_t = 3$	$N_t = 4$
$\begin{pmatrix} c_1 & c_2 \\ -c_2^* & c_1^* \end{pmatrix}$	$\begin{pmatrix} c_1 & c_2 & c_3 \\ -c_2 & c_1 & -c_4 \\ -c_3 & c_4 & c_1 \\ -c_4 & -c_3 & c_2 \\ c_1^* & c_2^* & c_3^* \\ -c_2^* & c_1^* & -c_4^* \\ -c_3^* & c_4^* & c_1^* \\ -c_4^* & -c_3^* & c_2^* \end{pmatrix}$	$\begin{pmatrix} c_1 & c_2 & c_3 & c_4 \\ -c_2 & c_1 & -c_4 & c_3 \\ -c_3 & c_4 & c_1 & -c_2 \\ -c_4 & -c_3 & c_2 & c_1 \\ c_1^* & c_2^* & c_3^* & c_4^* \\ -c_2^* & c_1^* & -c_4^* & c_3^* \\ -c_3^* & c_4^* & c_1^* & -c_2^* \\ -c_4^* & -c_3^* & c_2^* & c_1^* \end{pmatrix}$

Table 2.5 – ¾ Rate complex generalized orthogonal transmission matrices

$N_t = 3$	$N_t = 4$
$\begin{pmatrix} c_1 & c_2 & \frac{c_3}{\sqrt{2}} \\ -c_2^* & c_1^* & \frac{c_3}{\sqrt{2}} \\ \frac{c_3^*}{\sqrt{2}} & \frac{c_3^*}{\sqrt{2}} & \frac{(-c_1-c_1^*+c_2-c_2^*)}{2} \\ \frac{c_3^*}{\sqrt{2}} & -\frac{c_3^*}{\sqrt{2}} & \frac{(c_1-c_1^*+c_2+c_2^*)}{2} \end{pmatrix}$	$\begin{pmatrix} c_1 & c_2 & \frac{c_3}{\sqrt{2}} & \frac{c_3}{\sqrt{2}} \\ -c_2^* & c_1^* & \frac{c_3}{\sqrt{2}} & -\frac{c_3}{\sqrt{2}} \\ \frac{c_3^*}{\sqrt{2}} & \frac{c_3^*}{\sqrt{2}} & \frac{(-c_1-c_1^*+c_2-c_2^*)}{2} & \frac{(c_1-c_1^*-c_2-c_2^*)}{2} \\ \frac{c_3^*}{\sqrt{2}} & -\frac{c_3^*}{\sqrt{2}} & \frac{(c_1-c_1^*+c_2+c_2^*)}{2} & -\frac{(c_1+c_1^*+c_2-c_2^*)}{2} \end{pmatrix}$

2.2.3.3 STBC System from Orthogonal Designs

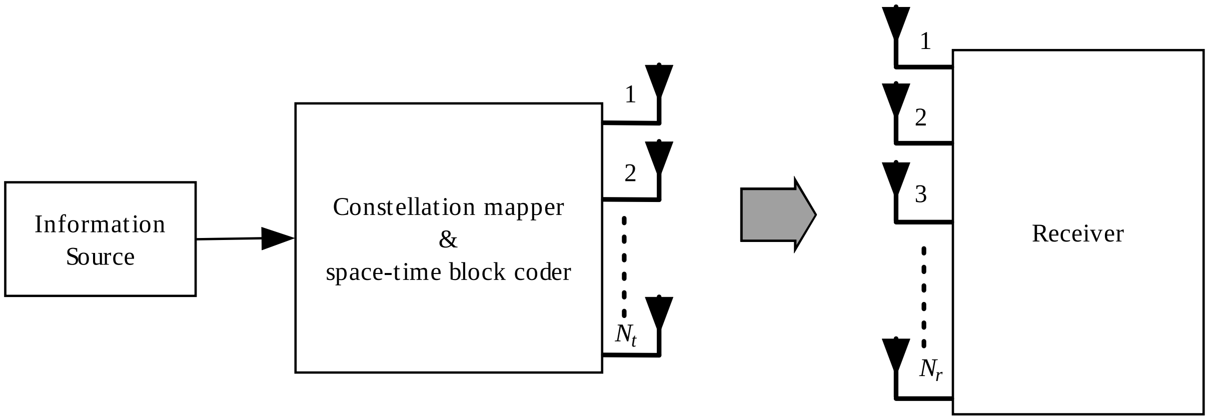

Fig. 2.3 Space-time block coded transmit diversity system for multiple transmit and multiple receive antennas

Consider a transmit and receive diversity system as shown Fig. 2.3 where the transmitter is equipped with N_t transmit antennas and receiver is equipped with N_r receive antennas. Both transmit antennas and receive antennas are well separated from each other to obtain independent channel fading. Further, the channel fading is assumed to be quasi static so that the path gains are constant over a frame and vary from one frame to another.

Consider the system using the STBC defined by the transmission matrix $\boldsymbol{G}_{N_t}$. At each time slot $t(=1,2...p)$, signals $C_{t,j},(j=1,2,...N_t)$ are taken from transmission matrix $\boldsymbol{G}_{N_t}$ and simultaneously transmitted from the N_t transmit antennas. The flat fading path gain $h_{j,k}$ is defined as from j^{th} transmit antenna of the transmitter to the $k^{th}(k=1,2,...N_r)$ receive antenna of the receiver respectively.

Based on the system mentioned above, the received signal at k antenna during t^{th} time slot, $r_{t,k}$ is given by

$$r_{t,k} = \sum_{j=1}^{N_t} h_{j,k} C_{t,j} + \eta_{t,k} \tag{2.11}$$

where the noise samples $\eta_{t,k}$ are independent samples of a zero mean complex Gaussian random variable with variance $1/(2SNR)$ per complex dimension. The average energy of the symbols transmitted from each antenna is normalized to be $1/N_t$. Therefore the average power of the received signal at each antenna is 1.

Assuming coherent detection and perfect channel information is available at the receiver. The decision metric can be computed by ML rule [49],

$$\sum_{t=1}^{p} \sum_{k=1}^{N_r} \left\| r_{t,k} - \sum_{j=1}^{N_t} h_{j,k} C_{t,j} \right\|^2 \tag{2.12}$$

over all codewords $C_{1,1}C_{1,2}...C_{1,N_t}C_{2,1}C_{2,1}.....C_{2,N_t}..........C_{p,1}C_{p,2}...C_{p,N_t}$ and decision can be taken in favor of the codeword that minimizes the decision metric given in (2.12).

2.3 Differential Space-Time Block Coding

Space-time block coding schemes given in section 2.2 assume perfect channel knowledge is available at the receiver to decode the received signal. However, in real situation the channel knowledge is not known perfectly. This problem is subdued in two ways. One method is employing appropriate channel estimation techniques with the training symbols or pilot symbols. This method will be covered in the next chapters. Another method is differential detection scheme where the symbols are decoded without the channel knowledge. Differential detection schemes are widely used for single transmit antenna systems. Furthermore, the differential detection schemes are used in the IEEE IS-54 standard. This motivates the generalization of the differential detection schemes for the multiple transmit antenna schemes that might be used in next generation communication systems such as wideband code division multiple access (WCDMA) [56] and CDMA 2000 [56].

A partial differential solution for Alamouti scheme [14] is proposed in [23], where symbols are decoded at the receiver without the channel knowledge. In the mean time, known symbols should be transmitted at the beginning of the transmission of a frame. The scheme given in [23] can be characterized as joint channel and symbol estimation. In this scheme [23], detected symbols at time slot t are used to detect the symbols at time slot $t+1$. This joint channel and symbol estimation can lead to error propagation. A truly differential detection scheme has been proposed in [21] for two transmit antenna system and it is further extended to multiple transmit antennas in [22]. Detail description of the scheme given in [21] is presented and their performance is analyzed in the following sections.

2.3.1 Encoding of DSTBC

The transmitter starts to transmit the symbols as Alamouti codes for first and second symbol duration of the frame. These arbitrary symbols c_1 and c_2 at time 1 and symbols

$-c_2^*$ and c_1^* at time 2 are transmitted from transmit antennas 1 and 2 respectively. These two transmissions do not convey any information. Instead, it will be used to encode and decode subsequent symbols. Consider c_{2t-1} and c_{2t} are transmitted at time $2t-1$ from transmit antennas 1 and 2 respectively and $-c_{2t}^*$ and c_{2t-1}^* are transmitted at time $2t$ from transmit antennas 1 and 2 respectively. Subsequently, the symbols c_{2t+1} and c_{2t+2} are encoded by differential encoder as given in the following equation:

$$\begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix} = A(Bits_{2t+1})\begin{pmatrix} c_{2t-1} & c_{2t} \end{pmatrix} + B(Bits_{2t+1})\begin{pmatrix} -c_{2t}^* & c_{2t-1}^* \end{pmatrix}, \tag{2.13}$$

where $Bits_{2t+1}$ is a block of bits that arrives at the encoder at time $2t+1$. $A(Bits)$ and $B(Bits)$ are defined by the mapping functions which are given as,

$$A(Bits) = (c_{2t+1}c_{2t-1}^* + c_{2t+2}c_{2t}^*) \tag{2.14}$$

$$B(Bits) = (-c_{2t+1}c_{2t}^* + c_{2t+2}c_{2t-1}^*), \tag{2.15}$$

where c_{2t-1} and c_{2t} are predefined and depending on signal constellation. c_{2t+1} and c_{2t+2} are mapped from gray coded signal constellations [50].

For example, consider the BPSK modulation scheme with constellation points $\left[-\frac{1}{\sqrt{2}} \quad \frac{1}{\sqrt{2}}\right]$. Let $c_{2t-1} = c_{2t} = \frac{1}{\sqrt{2}}$. Then the mapping $M(Bits) = \begin{pmatrix} A(Bits) & B(Bits) \end{pmatrix}$ maps the two input bits from the set $\nu = \{(1,0),(0,1),(-1,0),(0,-1)\}$ as given below,

$$M(00) = \begin{pmatrix} -1 & 0 \end{pmatrix}$$

$$M(01) = \begin{pmatrix} 0 & 1 \end{pmatrix}$$

$$M(10) = \begin{pmatrix} 0 & -1 \end{pmatrix}$$

$$M(11) = \begin{pmatrix} 1 & 0 \end{pmatrix}$$

By using the mapped information of $A(Bits)$ and $B(Bits)$ in the equation (2.13), c_{2t+1} and c_{2t+2} are calculated to transmit at time $2t+1$ from antennas 1 and 2 respectively. At time $2t+2$, $-c_{2t+2}^*$ and c_{2t+1}^* are transmitted from antennas 1 and 2 respectively. This process is continued until the end of the frame.

2.3.2 Decoding of DSTBC

Consider one receive antenna for notational simplicity. Let $r_{2t-1}, r_{2t}, r_{2t+1}$ and r_{2t+2} be the received signals, they can be written as in the vector form,

$$\begin{pmatrix} r_{2t-1} & r_{2t}^* \end{pmatrix} = \begin{pmatrix} c_{2t-1} & c_{2t} \end{pmatrix} \boldsymbol{\Lambda}(h_1, h_2) + \boldsymbol{N}_{2t-1} \tag{2.16}$$

$$\begin{pmatrix} r_{2t+1} & r_{2t+2}^* \end{pmatrix} = \begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix} \boldsymbol{\Lambda}(h_1, h_2) + \boldsymbol{N}_{2t+1} \tag{2.17}$$

$$\begin{pmatrix} r_{2t} & -r_{2t-2}^* \end{pmatrix} = \begin{pmatrix} -c_{2t}^* & c_{2t-1}^* \end{pmatrix} \boldsymbol{\Lambda}(h_1, h_2) + \boldsymbol{N}_{2t} \tag{2.18}$$

where $\boldsymbol{\Lambda}(h_1, h_2) = \begin{pmatrix} h_1 & h_2^* \\ h_2 & -h_1^* \end{pmatrix}$, $\boldsymbol{N}_{2t-1} = \begin{pmatrix} \eta_{2t-1} & \eta_{2t}^* \end{pmatrix}$, $\boldsymbol{N}_{2t+1} = \begin{pmatrix} \eta_{2t+1} & \eta_{2t+2}^* \end{pmatrix}$ and $\boldsymbol{N}_{2t} = \begin{pmatrix} \eta_{2t} & -\eta_{2t-1}^* \end{pmatrix}$.

From the equations (2.16) and (2.17),

$$\begin{pmatrix} r_{2t-1} & r_{2t}^* \end{pmatrix}\begin{pmatrix} r_{2t+1} & r_{2t+2}^* \end{pmatrix}^* = \begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix}\boldsymbol{\Lambda}(h_1,h_2)\boldsymbol{\Lambda}^*(h_1,h_2)\begin{pmatrix} c_{2t-1}^* & c_{2t}^* \end{pmatrix}$$

$$+\begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix}\boldsymbol{\Lambda}(h_1,h_2)\boldsymbol{N}_{2t-1}^*$$

$$+\boldsymbol{N}_{2t+1}\boldsymbol{\Lambda}^*(h_1,h_2)\begin{pmatrix} c_{2t-1} & c_{2t} \end{pmatrix}^* + \boldsymbol{N}_{2t+1}\boldsymbol{N}_{2t-1}^* .$$

It follows that

$$r_{2t+1}r_{2t-1}^* + r_{2t}r_{2t+2}^* = (|h_1|^2 + |h_2|^2)(c_{2t+1}c_{2t-1}^* + c_{2t+2}c_{2t}^*)$$

$$+\begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix}\boldsymbol{\Lambda}(h_1,h_2)\boldsymbol{N}_{2t-1}^*$$

$$+\boldsymbol{N}_{2t+1}\boldsymbol{\Lambda}^*(h_1,h_2)\begin{pmatrix} c_{2t-1} & c_{2t} \end{pmatrix}^* + \boldsymbol{N}_{2t+1}\boldsymbol{N}_{2t-1}^* . \quad (2.19)$$

Let us define

$$\Re_1 = r_{2t+1}r_{2t-1}^* + r_{2t}r_{2t+2}^* \quad (2.20)$$

$$\mathbb{N}_1 = \begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix}\boldsymbol{\Lambda}(h_1,h_2)\boldsymbol{N}_{2t-1}^* + \boldsymbol{N}_{2t+1}\boldsymbol{\Lambda}^*(h_1,h_2)\begin{pmatrix} c_{2t-1} & c_{2t} \end{pmatrix}^* + \boldsymbol{N}_{2t+1}\boldsymbol{N}_{2t-1}^* \quad (2.21)$$

then $\Re_1$ can be written as,

$$\Re_1 = (|h_1|^2 + |h_2|^2)A(Bits) + \mathbb{N}_1 . \quad (2.22)$$

Next, the receiver needs the second term in the right side of (2.13).

From the equations (2.17) and (2.18),

$$\begin{pmatrix} r_{2t+1} & r_{2t+2}^* \end{pmatrix}\begin{pmatrix} r_{2t} & r_{2t-1}^* \end{pmatrix}^* = \begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix}\boldsymbol{\Lambda}(h_1,h_2)\boldsymbol{\Lambda}^*(h_1,h_2)\begin{pmatrix} -c_{2t} & c_{2t-1} \end{pmatrix}$$

$$+\begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix}\boldsymbol{\Lambda}(h_1,h_2)\boldsymbol{N}_{2t}^*$$

$$+\boldsymbol{N}_{2t+1}\boldsymbol{\Lambda}^*(h_1,h_2)\begin{pmatrix} -c_{2t}^* & c_{2t-1}^* \end{pmatrix}^* + \boldsymbol{N}_{2t+1}\boldsymbol{N}_{2t}^* .$$

It follows that

$$r_{2t+1}r_{2t}^* - r_{2t-1}r_{2t+2}^* = (|h_1|^2 + |h_2|^2)(-c_{2t+1}c_{2t}^* + c_{2t+2}c_{2t-1}^*)$$

$$+\begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix} \boldsymbol{\Lambda}(h_1, h_2)\boldsymbol{N}_{2t}^*$$

$$+\boldsymbol{N}_{2t+1}\boldsymbol{\Lambda}^*(h_1, h_2)\begin{pmatrix} -c_{2t}^* & c_{2t-1}^* \end{pmatrix}^* + \boldsymbol{N}_{2t+1}\boldsymbol{N}_{2t}^* . \quad (2.23)$$

Let us define

$$\Re_2 = r_{2t+1}r_{2t}^* - r_{2t-1}r_{2t+2}^* \quad (2.24)$$

$$\mathbb{N}_2 = \begin{pmatrix} c_{2t+1} & c_{2t+2} \end{pmatrix} \boldsymbol{\Lambda}(h_1, h_2)\boldsymbol{N}_{2t}^* + \boldsymbol{N}_{2t+1}\boldsymbol{\Lambda}^*(h_1, h_2)\begin{pmatrix} -c_{2t}^* & c_{2t-1}^* \end{pmatrix}^* + \boldsymbol{N}_{2t+1}\boldsymbol{N}_{2t}^* . \quad (2.25)$$

Then $\Re_2$ can be written as,

$$\Re_2 = (|h_1|^2 + |h_2|^2)B(Bits) + \mathbb{N}_2 . \quad (2.26)$$

From the equations (2.22) and (2.26),

$$\begin{pmatrix} \Re_1 & \Re_2 \end{pmatrix} = (|h_1|^2 + |h_2|^2)\begin{pmatrix} A(Bits) & B(Bits) \end{pmatrix} + \begin{pmatrix} \mathbb{N}_1 & \mathbb{N}_2 \end{pmatrix} \quad (2.27)$$

To compute $\begin{pmatrix} A(Bits) & B(Bits) \end{pmatrix}$, the receiver now computes vector ν, that is closest to $\begin{pmatrix} \Re_1 & \Re_2 \end{pmatrix}$. Once the vector is computed, the inverse mapping of $M(Bits)$ is applied and the transmitted bits are decoded.

DSTBC for multiple transmit antennas can be extended from above derivations that is given in [22] based on the theory of generalized orthogonal designs.

2.4 Simulation

In this section, BER performance of the STBC and DSTBC schemes are obtained using Monte-Carlo simulations. Alamouti codes for two transmit antenna diversity and half rate complex generalized orthogonal designs for three and four transmit antenna diversity are considered. The transmission matrices for the simulation are given in

Table 2.4. The channel is considered as Rayleigh quasi-static flat fading. The amplitude of fading from each transmit antenna to each receive antenna is assumed to be mutually uncorrelated and i.i.d. Furthermore, the transmitted power is shared equally between transmit antennas in order to ensure the same total power as one transmit antenna scheme. In the STBC cases, perfect channel knowledge is assumed at the receiver to decode the signals.

2.4.1 Simulation Results

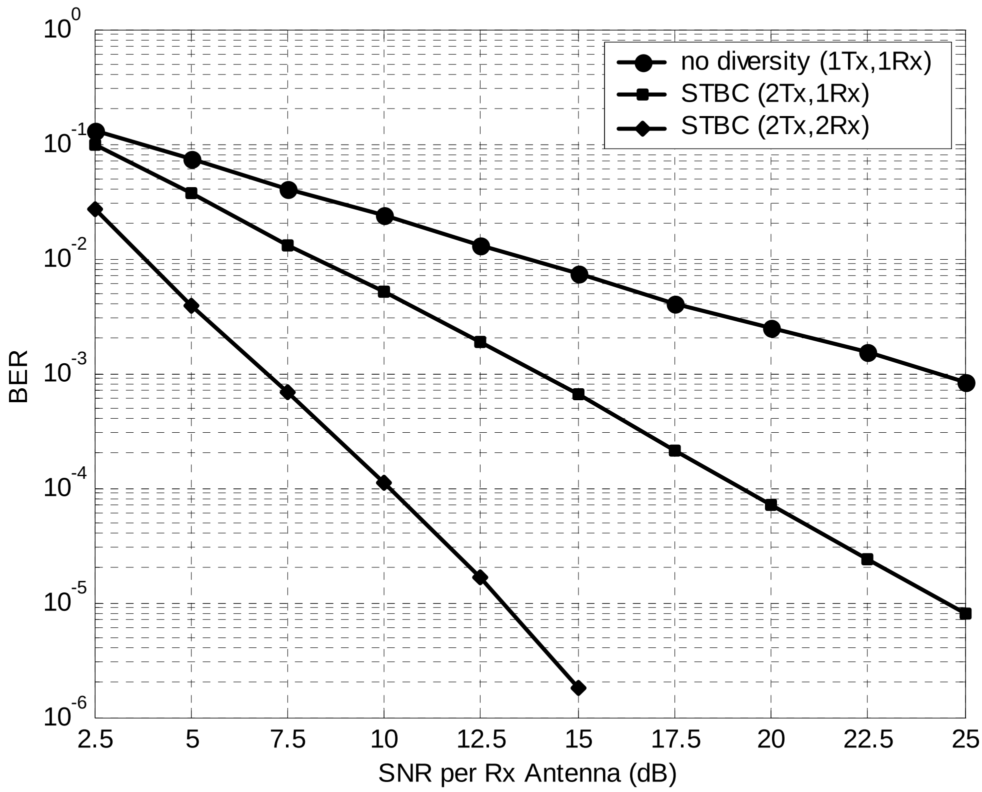

Fig. 2.4: The BER performance of STBC scheme for two transmit antennas using BPSK modulation scheme

In Fig. 2.4, BER performance of the STBC scheme (Alamouti scheme) for both two transmit one receive and two transmit two receive antenna systems are compared with conventional one transmit and one receive antenna scheme. As expected, the BER performance is improved with diversity order.

Fig. 2.5 provides simulation results for transmission of 1 bits/s/Hz using one, two, three and four transmit antennas. One and two transmit diversity schemes use the BPSK signal constellations. On the other hand, three and four transmit diversity schemes use the QPSK signal constellations and the transmission matrix, $\boldsymbol{G}_3$ and $\boldsymbol{G}_4$ of Table 2.4 respectively. Since $\boldsymbol{G}_3$ and $\boldsymbol{G}_4$ are half rate codes, the total transmission rate in each case is 1 bits/s/Hz.

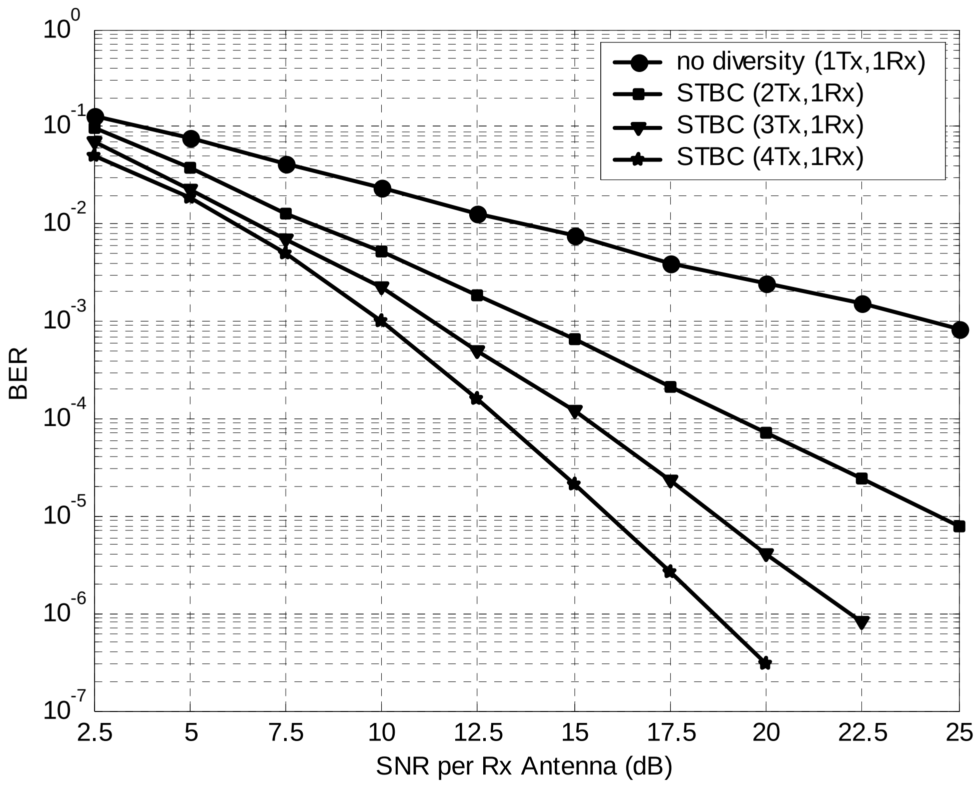

Fig. 2.5: The BER performance comparison of STBC scheme for two, three and four transmit antennas with no diversity scheme

The above results show that diversity gains can be achieved significantly by increasing the number of transmit antennas with slightly high decoding complexity.

The error performance of DSTBC scheme is compared with the corresponding coherent STBC schemes in Fig. 2.6, Fig. 2.7 and Fig. 2.8 for two, three and four transmit antennas respectively. In these simulations, BPSK signal constellation is used. Due to the power of noise terms $\mathbb{N}_1$ and $\mathbb{N}_2$ are more than two times of the corresponding noise power of the coherent scheme, the performance of the differential scheme is at least 3 dB less than that of coherent scheme.

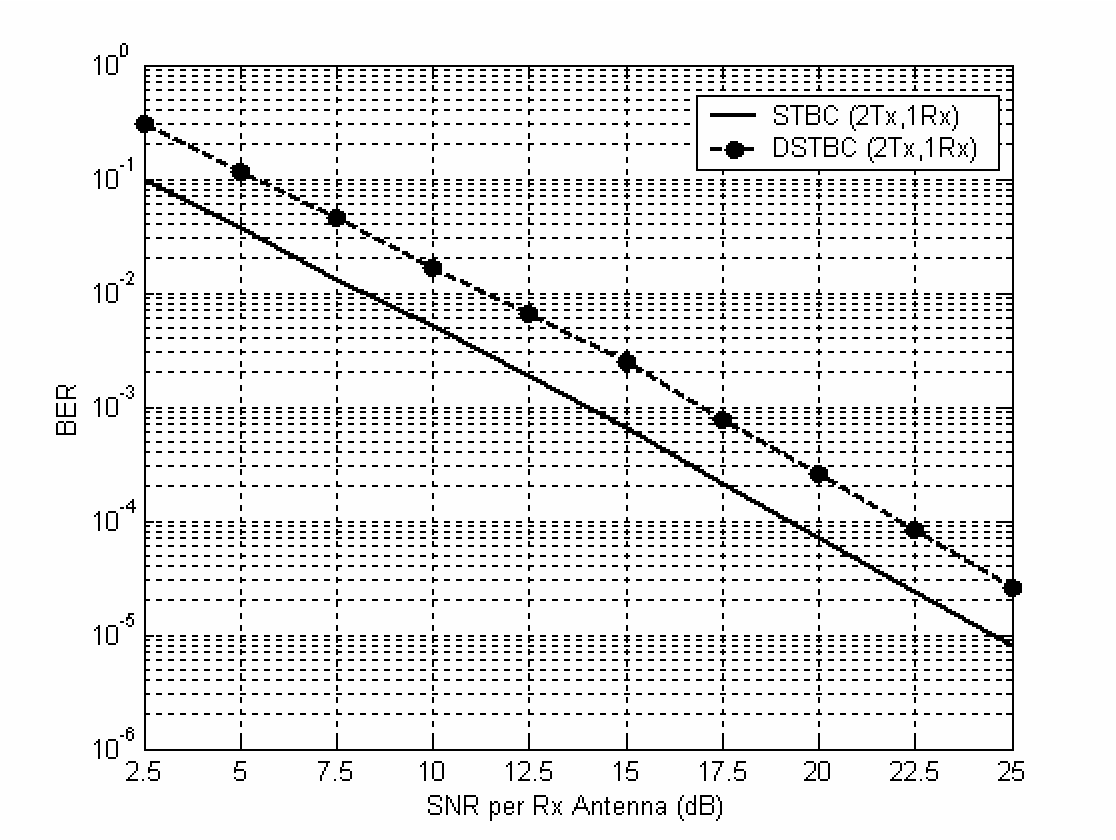

Fig. 2.6 The BER performance comparison of STBC and DSTBC schemes for two transmit antennas.

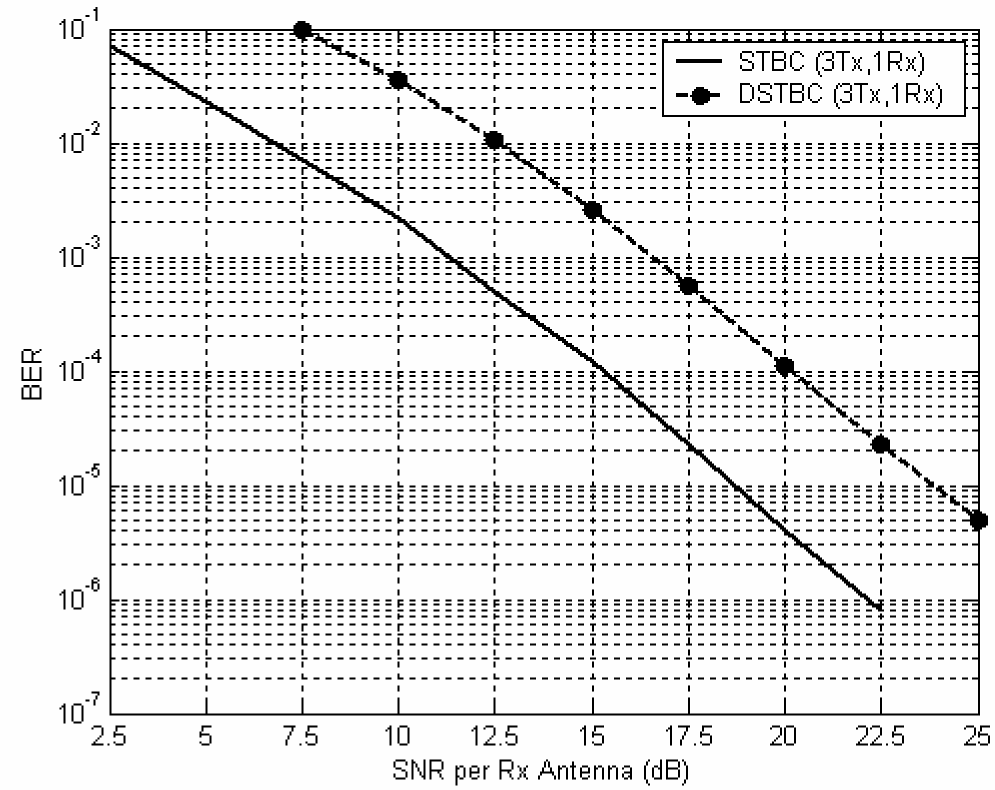

Fig. 2.7 The BER performance comparison of STBC and DSTBC schemes for three transmit antennas

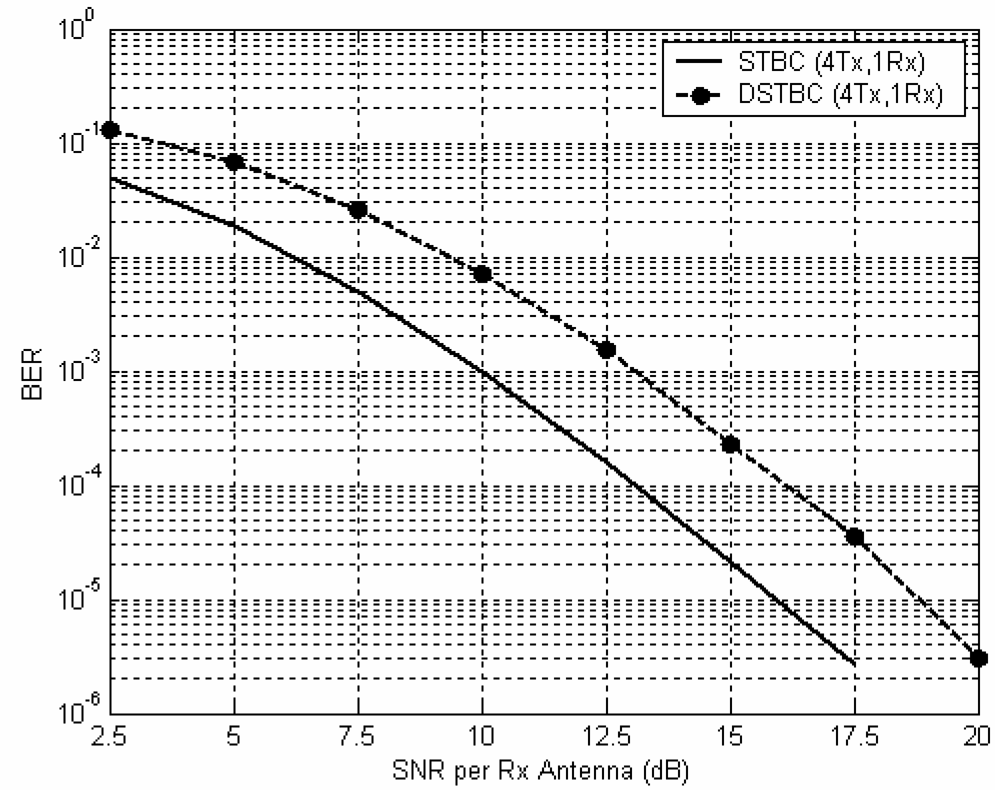

Fig. 2.8 The BER performance comparison of STBC and DSTBC schemes for four transmit antennas

2.5 Summary

Space-time block coding system has been presented in section 2.2. STBC scheme for two transmit one receive and two transmit multiple receive antennas are presented in section 2.2.1 and section 2.2.2 respectively. It is further extended for multiple transmit antennas in section 2.2.3 using the theory of generalized orthogonal designs for real and complex signal constellations. A differential space-time block coding has been given in section 2.3. Subsequently, simulation results are given in section 2.4.

CHAPTER 3

OVERVIEW OF INTERFERENCE SUPPRESSION IN STBC MULITUSER SYSTEMS

Since the wireless channel is a shared medium, interference mitigation in wireless communication systems is important to enhance the system performance. In this chapter, statistical and adaptive signal processing techniques are studied and applied to mitigate the interference in the STBC multiuser systems. After the introduction, section 3.2 derives the LMS and RLS algorithms based on MMSE and LSE, respectively. MMSE interference suppression from the literature is reviewed in section 3.3 which will be used in our proposed adaptive interference cancellation and suppression techniques that will be covered in chapter 4 and chapter 5.

3.1 Introduction

Signal processing algorithms nowadays are generally implemented using digital hardware operating on digital signals. The basic foundation of this modern approach based on discrete-time system theory [51, 52]. Since Widrow and Hoff [53] introduced the concept of the least mean square (LMS) adaptive algorithm based on minimum mean square estimation (MMSE) for estimating unknown process in 1959, adaptive signal processing or adaptive filtering has maintained great vitality for many years and matured to the point where it has now become an indispensable part of statistical signal processing.

Much effort has been made to develop the adaptive signal processing techniques in [49,51,54]. Numerous research works being carried out in this field indicate that there is no unique perfect solution to the adaptive filtering problem. So far, two aspects in adaptive filtering are shown to be fruitful. Many researchers have continued in their efforts to develop algorithms that can converge fast. Others have concentrated on computational efficiency. With regard to adaptive filtering problems, there are many recursive algorithms available, each of which offers certain desirable features. The least mean square (LMS) algorithm is very simple but the rate of convergence is very slow compared to other algorithms and depends on the input process. On the other hand, recursive least square (RLS) algorithm converges fast and behaves robustly against input process at the expense of computational complexity.

3.2 Statistical and Adaptive Signal Processing Techniques

3.2.1 Minimum Mean Square Estimation (MMSE) and Wiener-Hopf Equations

Consider the statistical filtering problem described in Fig. 3.1. The filter input is denoted by $r_1(n), r_2(n), \ldots r_N(n)$ and the impulse response of the filter is denoted by complex weights $w_0, w_1, \ldots w_N$. The output of the linear filter at discrete time n is defined by the sum of weighted inputs

$$y(n) = \sum_{k=1}^{N} w_k^* r_k(n), \qquad n = 0,1,2,...... \tag{3.1}$$

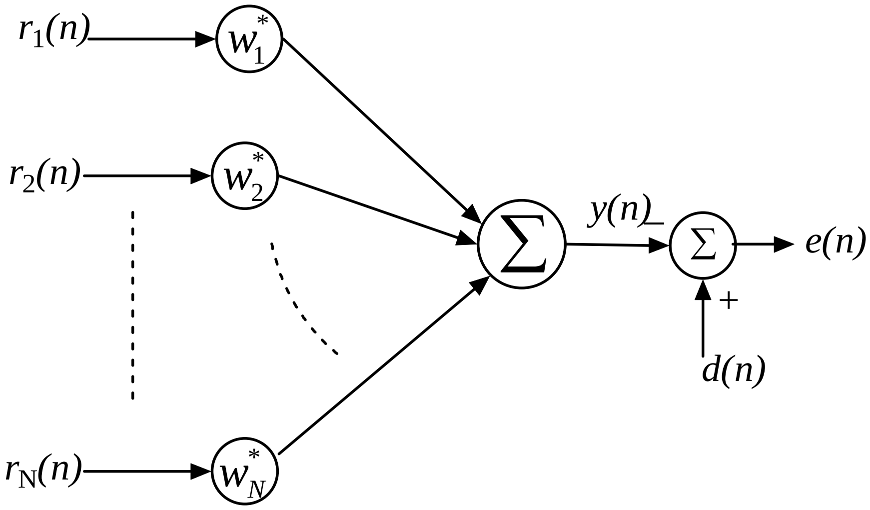

Fig. 3.1 Linear filter for MMSE

The filter input and the desired response are single realizations of jointly wide-sense stationary stochastic processes. The estimation of $d(n)$ has error, which is defined by the difference

$$e(n) = d(n) - y(n) = d(n) - \sum_{k=1}^{N} w_k^* r_k(n)\,. \tag{3.2}$$

The estimation error $e(n)$ is a sample of a random process. The mean square value of $e(n)$ is minimized in the design of an optimum filter. The cost function of the mean square error is defined by,

$$J = E[e(n)e^*(n)] \tag{3.3}$$

$$= E[|e(n)|^2]\,,$$

where E denotes the expectation operator. The requirement to design the optimum filter is to minimize the value of J. The correlation between the filter output $y(n)$ and the estimation error $e(n)$ can be written as,

$$E\left[y(n)e^{*}(n)\right]=E\left[\sum_{k=1}^{N}w_{k}^{*}r_{k}(n)e^{*}(n)\right]$$

$$=\sum_{k=1}^{N}w_{k}^{*}E[r_{k}(n)e^{*}(n)]. \tag{3.4}$$

By applying the theory of principle of orthogonality described in [49], the correlation between the optimum filter output $y_{o}(n)$ and the optimum filter estimation error $e_{o}(n)$ should be zero, that can be written as,

$$E[y_{o}(n)e_{o}^{*}(n)]=\sum_{k=1}^{N}w_{ok}^{*}E[r_{k}(n)e_{o}^{*}(n)]=0, \tag{3.5}$$

where w_{ok} is the k^{th} weight of the optimum filter.

Therefore, $E[r_{k}(n)e_{o}^{*}(n)]=0$ (3.6)

By expanding the equation (3.6),

$$E[\boldsymbol{r}(n)e_{o}^{*}(n)]=E[\boldsymbol{r}(n)(d^{*}(n)-\boldsymbol{w}_{o}\boldsymbol{r}^{*}(n))]=\boldsymbol{0}$$

$$E[\boldsymbol{r}(n)d^{*}(n)]=\boldsymbol{w}_{o}E[\boldsymbol{r}(n)\boldsymbol{r}^{*}(n)], \tag{3.7}$$

where received signal vector is $\boldsymbol{r}(n)=\left[r_{1}(n)\quad r_{2}(n)\quad\cdots\quad r_{N}(n)\right]^{T}$ and weight vector is $\boldsymbol{w}_{o}=\left[w_{o1}\quad w_{o2}\quad\cdots\quad w_{oN}\right]^{T}$. Signal correlation matrix $\boldsymbol{R}$ and cross correlation vector $\boldsymbol{P}$ are defined as $\boldsymbol{R}=E[\boldsymbol{r}(n)\boldsymbol{r}^{*}(n)]$ and $\boldsymbol{P}=E[\boldsymbol{r}(n)d^{*}(n)]$ respectively. By rewriting the equation (3.7), Wiener-Hopf equations can be formulated in the matrix form as,

$$\boldsymbol{P}=\boldsymbol{R}\boldsymbol{w}_{o}. \tag{3.8}$$

If correlation matrix $\boldsymbol{R}$ is nonsingular, optimum weights for the filter can be derived from the equation (3.8) and that can be written as,

$$\boldsymbol{w}_o = \boldsymbol{R}^{-1}\boldsymbol{P} . \tag{3.9}$$

3.2.2 Least Mean Square (LMS) Algorithm

The LMS algorithm is a stochastic gradient algorithm and a replacement for the steepest descent algorithm that uses a deterministic gradient approach in recursive computation of the wiener filter weights [49]. Further, It does not need the matrix inversion or expectation of pertinent functions to compute the weights of the filter. Moreover, the LMS algorithm is most popular linear adaptive filtering algorithm because of its simplicity.

By expanding the cost function (3.3),

$$J(n) = \sigma_d^2 - \boldsymbol{w}^*(n)\boldsymbol{P} - \boldsymbol{P}^*\boldsymbol{w}(n) + \boldsymbol{w}^*(n)\boldsymbol{R}\boldsymbol{w}(n) , \tag{3.9}$$

where σ_d^2 is variance of the desired response $d(n)$. The gradient vector $\nabla J(n)$ for given value of $\boldsymbol{w}(n)$ is,

$$\nabla J(n) = -2\boldsymbol{P} + 2\boldsymbol{R}\boldsymbol{w}(n) . \tag{3.10}$$

Accordingly, the steepest decent algorithm is described in [49],

$$\boldsymbol{w}(n+1) = \boldsymbol{w}(n) - \frac{1}{2}\mu\nabla J(n) , \tag{3.11}$$

where μ is step size parameter.

By substituting the (3.10) in (3.11), the steepest decent algorithm for tap weight vector can be written as,

$$\boldsymbol{w}(n+1)=\boldsymbol{w}(n)+\mu[\boldsymbol{P}-\boldsymbol{R}\boldsymbol{w}(n)]\,. \tag{3.12}$$

Table 3.1: LMS Algorithm

Initialization:	
	μ =step size parameter
	$0<\mu<\frac{2}{\lambda_{\max}}$; $\lambda_{\max}$ is maximum Eigen value of the input process
	If prior knowledge of the tap-weight vector $\hat{\boldsymbol{w}}(n)$ is available
	$\hat{\boldsymbol{w}}(0)=\hat{\boldsymbol{w}}(n)$
	else
	$\hat{\boldsymbol{w}}(0)=0$
Data:	
	$\boldsymbol{r}(n)$ = input vector
	$d(n)$ = desired response
Computation:	
	For n=0,1,2,3… Compute
	$e(n)=d(n)-\hat{\boldsymbol{w}}^{*}(n)\boldsymbol{r}(n)$
	$\hat{\boldsymbol{w}}(n+1)=\hat{\boldsymbol{w}}(n)+\mu\boldsymbol{r}(n)e^{*}(n)$

If it were possible to measure the gradient vector $\nabla J(n)$ at each iteration n, and if the step size is chosen suitably, then the tap weight vector computed by using the steepest descent algorithm would converge to the optimum Wiener solution. But in reality, exact measurement of gradient vector is not possible. To overcome this problem, the simplest choice of estimators is to use instantaneous estimates for $\boldsymbol{R}$ and $\boldsymbol{P}$ that are based on samples values of the tap input vector and desired response, defined respectively,

$$\hat{\boldsymbol{R}}(n) = \boldsymbol{r}(n)\boldsymbol{r}^*(n) \tag{3.13}$$

and $$\hat{\boldsymbol{P}}(n) = \boldsymbol{r}(n)\boldsymbol{d}^*(n)\ . \tag{3.14}$$

Substituting the estimates of $\hat{\boldsymbol{R}}$ and $\hat{\boldsymbol{P}}$ in (3.12), the new recursive relationship for updating the weight vector can be written as,

$$\hat{\boldsymbol{w}}(n+1) = \hat{\boldsymbol{w}}(n) + \mu\boldsymbol{r}(n)[d^*(n) - \boldsymbol{r}^*(n)\hat{\boldsymbol{w}}(n)] \tag{3.15}$$

$$\hat{\boldsymbol{w}}(n+1) = \hat{\boldsymbol{w}}(n) + \mu\boldsymbol{r}(n)e^*(n)\ .$$

This is called LMS algorithm that is summarized in Table 3.1.

The rate of convergence of the LMS algorithm depends on the Eigen value spread of the input process and selection parameter μ.

3.2.3 Linear Least Square Estimation

The method of least squares is viewed as an alternative approach to Wiener filter theory that is covered in section 3.2.1. Basically, Wiener filters are derived from ensemble averaging with the underlying environment that is assumed to be wide-sense stationary. On the other hand, least square method is based on time averaging that is called deterministic approach. The result of the filter depends on the number of samples used for computations.

Consider the linear filter model shown in Fig. 3.2 where $r(i), r(i-1)$ and $r(i-N)$ are input signals observed at time $i, i-1$ and $i-N$ respectively. The desired signal at time i is given by $d(i)$. The $w_0, w_1, ...w_{N-1}$ are tap weights of the linear filter. The estimation error can be written as,

$$e(i) = d(i) - y(i), \tag{3.16}$$

where

$$y(i) = \sum_{k=0}^{N-1} w_k^* r(i-k). \tag{3.17}$$

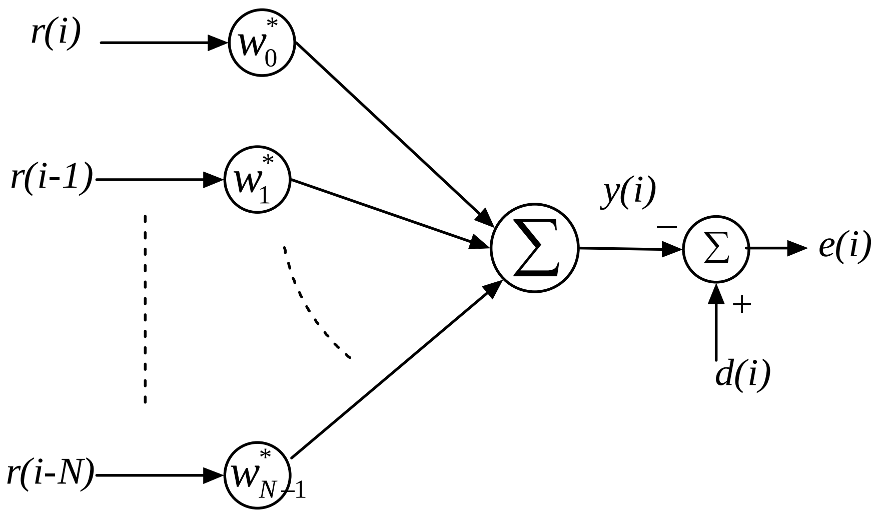

Fig. 3.2 Linear filter model for LSE

Substituting Eq. (3.17) into Eq. (3.16) yields

$$e(i) = d(i) - \sum_{k=0}^{N-1} w_k^* r(i-k). \tag{3.18}$$

In the method of least squares, the tap weights w_k of the filter should be optimized to minimize the cost function that is given by sum of error squares as,

$$\varepsilon(w_0, w_1, ...w_{N-1}) = \sum_{i=i_1}^{i_2} |e(i)|^2, \tag{3.19}$$

where i_1 and i_2 are error minimization index limits that depend on size of data window. For minimization of cost function $\varepsilon(w_0, w_1, ...w_{N-1})$ with respect to the tap weights $w_0, w_1, ...w_{N-1}$ of the filter, derivative of the gradient vector $\nabla_k \varepsilon$ should statisfy

$$\nabla_k \varepsilon = 0, \qquad k = 0,1,\dots,N-1. \tag{3.20}$$

Let $e_{\min}(i)$ denote the minimum estimation error that is given by the optimum filter, then the Eq. (3.20) can be derived using principle of orthogonality,

$$\sum_{i=N}^{M} r(i-k)e_{\min}^{*}(i) = 0, \qquad k = 0,1,\dots,N-1. \tag{3.21}$$

Minimum estimation error $e_{\min}(i)$ can be written as,

$$e_{\min}(i) = d(i) - \sum_{t=0}^{N-1} \hat{w}_t^{*} r(i-t), \tag{3.22}$$

where t is used as dummy summation index and $\hat{w}_t$ is defined as optimum weight of the filter at index t.

Substituting Eq. (3.22) into Eq. (3.21) and rearranging,

$$\sum_{t=0}^{N-1} \hat{w}_t \sum_{i=N}^{M} r(i-t)r^{*}(i-t) = \sum_{i=N}^{M} r(i-t)d^{*}(i), \qquad k = 0,1,\dots,N-1. \tag{3.23}$$

Let us define time average correlation function

$$\phi(t,k) = \sum_{i=N}^{M} r(i-k)r^{*}(i-t), \qquad 0 \le (t,k) \le N-1 \tag{3.24}$$

and time average cross correlation function,

$$z(-k) = \sum_{i=N}^{M} r(i-k)d^{*}(i), \qquad 0 \le k \le N-1. \tag{3.25}$$

Using the time average correlation and cross correlation functions, the Eq. (3.22) can be rewritten as,

$$\sum_{t=0}^{N-1} \hat{w}_t \phi(t,k) = z(-k), \qquad k = 0,1,\dots,N-1. \tag{3.26}$$

The normal equations for linear least square filters can be formulated by writing Eq. (3.26) in matrix form and it is given by,

$$\boldsymbol{\phi}\hat{\boldsymbol{w}} = \boldsymbol{z}\,, \tag{3.27}$$

where optimum tap weight vector $\hat{\boldsymbol{w}} = [\hat{w}_0, \hat{w}_1, ...\hat{w}_{N-1}]^T$, (3.28)

time average cross correlation vector $\boldsymbol{z} = [z(0), z(-1), ...z(-M+1)]^T$ (3.29)

and time average correlation matrix

$$\boldsymbol{\phi} = \begin{bmatrix} \phi(0,0) & \phi(1,0) & \cdots & \phi(N-1,0) \\ \phi(0,1) & \phi(1,1) & \cdots & \phi(N-1,1) \\ \vdots & \vdots & \ddots & \vdots \\ \phi(0,N-1) & \phi(1,N-1) & \cdots & \phi(N-1,N-1) \end{bmatrix}. \tag{3.30}$$

3.2.4. Recursive Least Square Algorithm

In this section, adaptive implementation of the method of least square which is known as recursive least square algorithm (RLS) is introduced. The RLS algorithm is developed by exploiting a relation between method of least square and matrix algebra known as matrix inversion lemma that is also known as *Woodbary's Identity*. Significant feature of the RLS algorithm is that the rate of convergence of the algorithm is much faster than that of LMS algorithm. However this improvement of RLS filter is achieved at the expense of computational complexity. In contrast, nowadays computing power of digital computer is improved tremendously which helps to implement the RLS like algorithm in the real world applications. Another attractive performance of the RLS algorithm is that the rate of convergence of the algorithm does not depend on the Eigen value spread of the input process.

In the derivation of RLS algorithm, the cost function $\varepsilon(n)$ is redefined with forgetting factor λ that is given by,

$$\varepsilon(n) = \sum_{i=1}^{n} \lambda^{n-i} \left|e(i)\right|^2 , \tag{3.31}$$

where n is the variable length of the input data and $\lambda \le 1$. When $\lambda = 1$, above cost function is same as the cost function of linear least square problem given by Eq. (3.19). When $\lambda < 1$ the cost function given by (3.31) is not equal to (3.19). A regulating term is added to Eq. (3.31). The new cost function for recursive solution is given by,

$$\varepsilon(n) = \sum_{i=1}^{n} \lambda^{n-i} \left|e(i)\right|^2 + \delta\lambda^n \left\|\boldsymbol{w}(n)\right\|^2 . \tag{3.32}$$

Expanding the Eq. (3.32), the effect of regulating term $\delta\lambda^n \left\|\boldsymbol{w}(n)\right\|^2$ in the cost function is equivalent to the reformulation of time average correlation matrix $\boldsymbol{\phi}(n)$ and cross correlation matrix $\boldsymbol{z}(n)$ that are written as,

$$\boldsymbol{\phi}(n) = \sum_{i=1}^{n} \lambda^{n-i} \boldsymbol{r}(i)\boldsymbol{r}^*(i) + \delta\lambda^n \boldsymbol{I} \tag{3.33}$$

and $$\boldsymbol{z}(n) = \sum_{i=1}^{n} \lambda^{n-i} \boldsymbol{r}(i) d^*(i) \tag{3.34}$$

respectively. The Eq. (3.33) can be further expanded by isolating the corresponding term $i = n$,

$$\boldsymbol{\phi}(n) = \lambda\left[\sum_{i=1}^{n-1} \lambda^{n-1-i} \boldsymbol{r}(i)\boldsymbol{r}^*(i) + \delta\lambda^{n-1}\boldsymbol{I}\right] + \boldsymbol{r}(n)\boldsymbol{r}^*(n)$$

$$\boldsymbol{\phi}(n) = \lambda\boldsymbol{\phi}(n-1) + \boldsymbol{r}(n)\boldsymbol{r}^*(n) , \tag{3.35}$$

where $\boldsymbol{\phi}(n-1)$ is the old value of the correlation matrix, the matrix product $\boldsymbol{r}(n)\boldsymbol{r}^*(n)$ plays a major role in updating the correlation matrix. Similarly, the Eq. (3.34) can be expanded as,

$$\boldsymbol{z}(n) = \lambda\boldsymbol{z}(n-1) + \boldsymbol{r}(n) d^*(n) . \tag{3.36}$$

To obtain the weight vector, the normal equations (3.27) can be rearranged as,

$$\hat{\boldsymbol{w}}(n) = \boldsymbol{\phi}^{-1}(n)\boldsymbol{z}(n) . \tag{3.37}$$

Let $\boldsymbol{A}$ and $\boldsymbol{B}$ be two positive definite $M \times M$ matrices related by

$$\boldsymbol{A} = \boldsymbol{B}^{-1} + \boldsymbol{C}\boldsymbol{D}^{-1}\boldsymbol{C}^{*} , \tag{3.38}$$

where $\boldsymbol{D}$ is positive definite $N \times M$ matrix and $\boldsymbol{C}$ is an $M \times N$ matrix. From the matrix inversion lemma, the inverse of the matrix $\boldsymbol{A}$ can be expressed as,

$$\boldsymbol{A}^{-1} = \boldsymbol{B} - \boldsymbol{B}\boldsymbol{C}(\boldsymbol{D} + \boldsymbol{C}^{*}\boldsymbol{B}\boldsymbol{C})^{-1}\boldsymbol{C}^{*}\boldsymbol{B} . \tag{3.39}$$

Applying the matrix inversion lemma given in (3.39) to find out the inverse of the correlation matrix $\boldsymbol{\phi}(n)$ that is assumed to be invertible or non-singular.

Let define $\boldsymbol{A} = \boldsymbol{\phi}(n)$, $\boldsymbol{B}^{-1} = \lambda\boldsymbol{\phi}(n-1)$, $\boldsymbol{C} = \boldsymbol{r}(n)$ and $\boldsymbol{D} = 1$, then substituting these definitions into the Eq. (3.39), the inverse correlation can be,

$$\boldsymbol{\phi}^{-1}(n) = \lambda^{-1}\boldsymbol{\phi}^{-1}(n-1) - \frac{\lambda^{-2}\boldsymbol{\phi}^{-1}(n-1)\boldsymbol{r}(n)\boldsymbol{r}^{*}(n)\boldsymbol{\phi}^{-1}(n-1)}{1 + \lambda^{-1}\boldsymbol{r}^{*}(n)\boldsymbol{\phi}^{-1}(n-1)\boldsymbol{r}(n)} . \tag{3.40}$$

Let $$\boldsymbol{P}(n) = \boldsymbol{\phi}^{-1}(n) \tag{3.41}$$

and $$\boldsymbol{k}(n) = \frac{\lambda^{-1}\boldsymbol{P}(n-1)\boldsymbol{r}(n)}{1 + \lambda^{-1}\boldsymbol{r}^{*}(n)\boldsymbol{P}(n-1)\boldsymbol{r}(n)} . \tag{3.42}$$

Using equations (3.41) and (3.42), Eq. (3.40) can be written as,

$$\boldsymbol{P}(n) = \lambda^{-1}\boldsymbol{P}(n-1) - \lambda^{-1}\boldsymbol{k}(n)\boldsymbol{r}^{*}(n)\boldsymbol{P}(n-1) \tag{3.43}$$

and by rearranging the Eq. (3.42),

$$\boldsymbol{k}(n) = \boldsymbol{P}(n)\boldsymbol{r}(n) . \tag{3.44}$$

For the convenience, the Eq. (3.37) can be written as,

$$\hat{\boldsymbol{w}}(n) = \boldsymbol{\phi}^{-1}(n)\boldsymbol{z}(n) = \boldsymbol{P}(n)\boldsymbol{z}(n) . \tag{3.45}$$

Above given equation are utilized to derive the following recursive equation for weight vector $\hat{\boldsymbol{w}}(n)$,

$$\hat{\boldsymbol{w}}(n) = \hat{\boldsymbol{w}}(n-1) + \boldsymbol{k}(n)\xi^*(n)\,, \tag{3.46}$$

where the priori estimation error $\xi^*(n)$ is defined as,

$$\xi^*(n) = d(n) - \hat{\boldsymbol{w}}^*(n-1)\boldsymbol{r}(n)\,. \tag{3.47}$$

The RLS algorithm is summarized in the Table 3.2.

Table 3.2: RLS Algorithm

Initialization:
$\hat{\boldsymbol{w}}(0) = 0$
$\boldsymbol{P}(0) = \delta^{-1}\boldsymbol{I}$,
and
$\delta = \begin{cases} \text{small positive constant for high SNR} \\ \text{large positive constant for low SNR} \end{cases}$
Data:
$\boldsymbol{r}(n)$ = input vector
$d(n)$ = desired response
Computation:
For n=0,1,2,3… Compute
$\boldsymbol{k}(n) = \dfrac{\boldsymbol{P}(n-1)\boldsymbol{r}(n)}{\lambda + \boldsymbol{r}^*(n)\boldsymbol{P}(n-1)\boldsymbol{r}(n)}$,
$\xi^*(n) = d(n) - \hat{\boldsymbol{w}}^*(n-1)\boldsymbol{r}(n)$,
$\hat{\boldsymbol{w}}(n+1) = \hat{\boldsymbol{w}}(n) + \mu\boldsymbol{r}(n)e^*(n)$,
and
$\boldsymbol{P}(n) = \lambda^{-1}\boldsymbol{P}(n-1) - \lambda^{-1}\boldsymbol{k}(n)\boldsymbol{r}^*(n)\boldsymbol{P}(n-1)$

3.3 Interference Suppression Using MMSE

In this section, an interference suppression and ML decoding scheme for STBC multiuser systems given in [44] is described. The scheme mitigates the interference based on MMSE. Inherent problems of the MMSE interference suppression scheme are discussed in this section.

Fig. 3.3 shows the system model of two synchronous co-channel user scheme, in which each user is equipped with two transmit antennas and uses STBC. Both users communicate with the same base station that is equipped with two receive antennas. The second receive antenna is used to suppress the interference from the co-channel interferer without sacrifying the diversity order given by STBC.

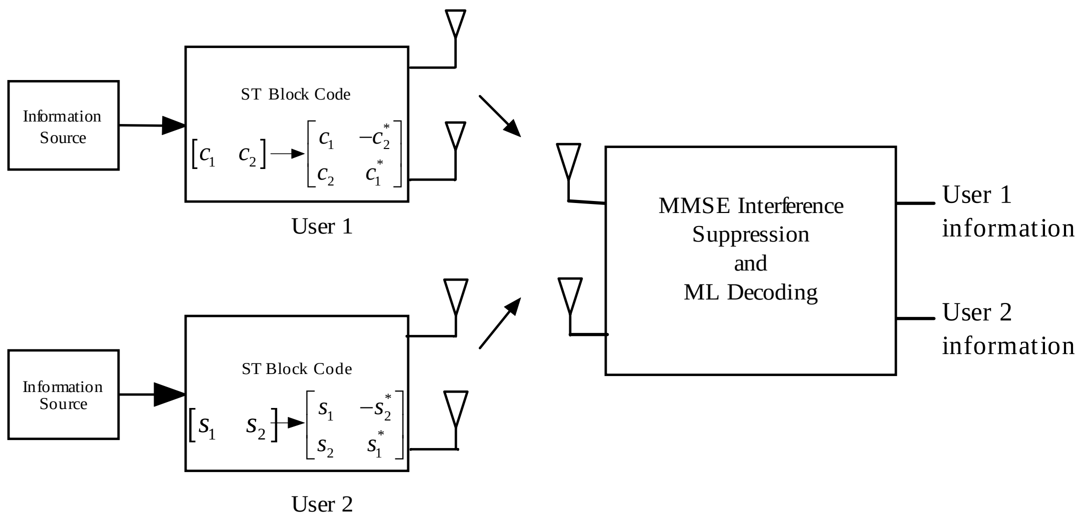

Fig. 3.3: MMSE Interference suppression system for two users with STBC

3.3.1 System Model

The flat fading path gain $h_{i,j}$ is defined as from i^{th} transmit antenna of first user to j^{th} receive antenna of the receiver. Similarly, $g_{i,j}$ is defined path gain between i^{th} transmit antenna of second user and j^{th} receive antenna of the receiver. The symbols

$\{c_1, c_2\}$ and $\{s_1, s_2\}$ are transmitted from the first and second users respectively. Let the received signals from the first receive antenna at first and second symbol intervals $r_{1,1}$ and $r_{1,2}$ respectively.

$$r_{1,1} = h_{1,1}c_1 + h_{2,1}c_2 + g_{1,1}s_1 + g_{2,1}s_2 + \eta_{1,1} \tag{3.48}$$

$$r_{1,2} = -h_{1,1}c_2^* + h_{2,1}c_1^* - g_{1,1}s_2^* + g_{2,1}s_1^* + \eta_{1,2} \tag{3.49}$$

The noise $\eta_{i,j}$ is modeled as a zero mean complex gaussian process with variance $(1/2\gamma)$ per complex dimension. Similarly, let the received signal from the second antenna over two consecutive symbol periods $r_{2,1}$ and $r_{2,2}$ respectively.

$$r_{2,1} = h_{1,2}c_1 + h_{2,2}c_2 + g_{1,2}s_1 + g_{2,2}s_2 + \eta_{2,1} \tag{3.50}$$

$$r_{22} = -h_{12}c_2^* + h_{22}c_1^* - g_{12}s_2^* + g_{22}s_1^* + \eta_{22} \tag{3.51}$$

Now, the overall signal vector can be written as,

$$\boldsymbol{r} = \begin{bmatrix} r_{1,1} \\ r_{1,2}^* \\ r_{2,1} \\ r_{2,2}^* \end{bmatrix} = \begin{bmatrix} h_{1,1} & h_{2,1} & g_{1,1} & g_{2,1} \\ h_{2,1}^* & -h_{1,1}^* & g_{2,1}^* & -g_{1,1}^* \\ h_{1,2} & h_{2,2} & g_{1,2} & g_{2,2} \\ h_{2,2}^* & -h_{1,2}^* & g_{2,2}^* & -g_{2,1}^* \end{bmatrix} \begin{bmatrix} c_1 \\ c_2 \\ s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} \eta_{1,1} \\ \eta_{1,2}^* \\ \eta_{2,1} \\ \eta_{2,2}^* \end{bmatrix}. \tag{3.52}$$

Above equation can further be simplified to,

$$\boldsymbol{r = H.\tilde{C} + \eta}, \tag{3.53}$$

where $\boldsymbol{H} = \begin{bmatrix} h_{1,1} & h_{2,1} & g_{1,1} & g_{2,1} \\ h_{2,1}^* & -h_{1,1}^* & g_{2,1}^* & -g_{1,1}^* \\ h_{1,2} & h_{2,2} & g_{1,2} & g_{2,2} \\ h_{2,2}^* & -h_{1,2}^* & g_{2,2}^* & -g_{2,1}^* \end{bmatrix}$, $\tilde{\boldsymbol{C}} = \begin{bmatrix} c_1 \\ c_2 \\ s_1 \\ s_2 \end{bmatrix}$ and $\boldsymbol{\eta} = \begin{bmatrix} \eta_{1,1} & \eta_{1,2}^* & \eta_{1,2} & \eta_{2,2}^* \end{bmatrix}^T$.

3.3.2 MMSE Interference Suppression

For notational simplicity, $\boldsymbol{r}$ is written as,

$$\boldsymbol{r} = \begin{bmatrix} r_{11} & r_{12}^* & r_{12} & r_{22}^* \end{bmatrix}^T = \begin{bmatrix} r_1 & r_2 & r_3 & r_4 \end{bmatrix}^T. \tag{3.54}$$

To perform the interference suppression, the mean squared error due to the co-channel interference and the noise in the decoded symbols c_1 and c_2 should be minimized. The cost function of the MSE can be written as,

$$J(\boldsymbol{w}\ \boldsymbol{\alpha}) = E\left\{\left\|\sum_{i=1}^{4} w_i^* r_i - \left(\alpha_1^* c_1 + \alpha_2^* c_2\right)\right\|^2\right\}$$

$$= E\left\{\left\|\boldsymbol{w}^* \boldsymbol{r} - \boldsymbol{\alpha}^* \boldsymbol{c}\right\|^2\right\}, \tag{3.55}$$

where, $\boldsymbol{w} = \begin{bmatrix} w_1 & w_2 & w_3 & w_4 \end{bmatrix}^T$ and $\boldsymbol{\alpha} = \begin{bmatrix} \alpha_1 & \alpha_2 \end{bmatrix}^T$.

The weights $\boldsymbol{w}$ and $\boldsymbol{\alpha}$ are selected to minimize the cost function $J(\boldsymbol{w}\ \boldsymbol{\alpha})$. In order to avoid the situation where optimal weights getting zero values, one of the weights α_1 or α_2 is set to one. First, let us assume that α_1=1. In this, case the error criterion is

$$J(\boldsymbol{w}\ \boldsymbol{\alpha}) = E\left\{\left\|\boldsymbol{w}^* \boldsymbol{r} - \alpha_2^* c_2 - c_1\right\|^2\right\} = E\left\{\left\|\tilde{\boldsymbol{w}}_1^* \tilde{\boldsymbol{r}}_1 - c_1\right\|^2\right\}$$

$$= J(\tilde{\boldsymbol{w}}_1), \tag{3.56}$$

where $\tilde{\boldsymbol{w}}_1 = \begin{bmatrix} \boldsymbol{w}^T & -\alpha_2 \end{bmatrix}^T$ and $\tilde{\boldsymbol{r}}_1 = \begin{bmatrix} \boldsymbol{r}^T & c_2 \end{bmatrix}^T$.

Then, $\tilde{\boldsymbol{r}}_1$ can be rewritten as,

$$\tilde{\boldsymbol{r}}_1 = \begin{bmatrix} \boldsymbol{H} & \boldsymbol{0} \\ \boldsymbol{0}^T & 1 \end{bmatrix} \begin{bmatrix} \tilde{\boldsymbol{C}} \\ c_2 \end{bmatrix} + \begin{bmatrix} \boldsymbol{\eta} \\ 0 \end{bmatrix} = \tilde{\boldsymbol{H}} \tilde{\boldsymbol{C}}_1 + \tilde{\boldsymbol{\eta}}, \tag{3.57}$$

where $\tilde{\boldsymbol{H}} = \begin{bmatrix} \boldsymbol{H} & \boldsymbol{0} \\ \boldsymbol{0}^T & 1 \end{bmatrix}$ and $\tilde{\boldsymbol{C}}_1 = \begin{bmatrix} \tilde{\boldsymbol{C}} \\ c_2 \end{bmatrix}$.

To minimize the cost function $J(\tilde{\boldsymbol{w}}_1)$, $\tilde{\boldsymbol{w}}_1$ should be selected to optimum value.

Expanding (3.56), it can be re-written as,

$$J(\tilde{\boldsymbol{w}}_1) = E\left\{(\tilde{\boldsymbol{w}}_1^* \tilde{\boldsymbol{r}}_1 - c_1)(\tilde{\boldsymbol{w}}_1^* \tilde{\boldsymbol{r}}_1 - c_1)^*\right\}$$

$$= \tilde{\boldsymbol{w}}_1^* \boldsymbol{R}_1 \tilde{\boldsymbol{w}}_1 + E\left\{c_1 c_1^*\right\} - \boldsymbol{Z}_1^* \tilde{\boldsymbol{w}}_1^* - \tilde{\boldsymbol{w}}_1^* \boldsymbol{Z}_1$$

$$= \tilde{\boldsymbol{w}}_1^* \boldsymbol{R}_1 \tilde{\boldsymbol{w}}_1 + E_S - \boldsymbol{Z}_1^* \tilde{\boldsymbol{w}}_1^* - \tilde{\boldsymbol{w}}_1^* \boldsymbol{Z}_1 , \qquad (3.58)$$

where E_S is symbol energy, $\boldsymbol{R}_1 = E\left\{\tilde{\boldsymbol{r}}_1 \tilde{\boldsymbol{r}}_1^*\right\}$ and $\boldsymbol{Z}_1 = E\left\{\tilde{\boldsymbol{r}}_1 c_1^*\right\}$.

Defining $\boldsymbol{V}_1 = \begin{bmatrix}1 & 0 & 0 & 0\end{bmatrix}^T$ and $\boldsymbol{V}_2 = \begin{bmatrix}0 & 1 & 0 & 0\end{bmatrix}^T$, $\boldsymbol{R}_1$ can be shown as,

$$\boldsymbol{R}_1 = \tilde{\boldsymbol{H}} E\left\{\tilde{\boldsymbol{C}}\tilde{\boldsymbol{C}}^*\right\} \tilde{\boldsymbol{H}}^* + E\left\{\tilde{\boldsymbol{\eta}}\tilde{\boldsymbol{\eta}}^*\right\}$$

$$= E_S \tilde{\boldsymbol{H}} \begin{bmatrix} \boldsymbol{I}_4 & \boldsymbol{V}_2 \\ \boldsymbol{V}_1^T & 1 \end{bmatrix} \tilde{\boldsymbol{H}}^* + N_0 \begin{bmatrix} \boldsymbol{I}_4 & \boldsymbol{0} \\ \boldsymbol{0}^T & 0 \end{bmatrix}$$

$$= E_S \begin{bmatrix} \boldsymbol{M} & \boldsymbol{h}_1 \\ \boldsymbol{h}_2^* & 1 \end{bmatrix}, \qquad (3.60)$$

where $\boldsymbol{M} = \boldsymbol{H}\boldsymbol{H}^* + \frac{1}{\gamma}\boldsymbol{I}_4 = E\left\{\boldsymbol{r}.\boldsymbol{r}^*\right\}; \boldsymbol{h}_2 = \begin{bmatrix} h_{2,1} & -h_{1,1}^* & h_{2,2} & -h_{1,2}^* \end{bmatrix}^T$ (second column of $\boldsymbol{H}$) and $\boldsymbol{h}_1 = \begin{bmatrix} h_{1,1} & h_{2,1}^* & h_{1,2} & h_{2,2}^* \end{bmatrix}^T$ (first column of $\boldsymbol{H}$).

$\boldsymbol{Z}_1$ can be shown as,

$$\boldsymbol{Z}_1 = E\left\{\tilde{\boldsymbol{r}}_1 c_1^*\right\} = E_S \tilde{\boldsymbol{H}} \begin{bmatrix} \boldsymbol{V}_1 \\ 0 \end{bmatrix}$$

$$= E_S \begin{bmatrix} h_{1,1} & -h_{2,1}^* & h_{2,1} & -h_{1,2}^* & 0 \end{bmatrix}^T = E_S \begin{bmatrix} \boldsymbol{h}_1 \\ 0 \end{bmatrix}. \qquad (3.61)$$

Differentiating (3.58) w.r.t. $\tilde{\boldsymbol{w}}_1$ the Wiener-Hopf equations for interference suppression can be written as,

$$\boldsymbol{R}_1\tilde{\boldsymbol{w}}_1 = \boldsymbol{Z}_1 . \tag{3.62}$$

By substituting (3.60) and (3.61) in (3.62),

$$\begin{bmatrix} \boldsymbol{M} & \boldsymbol{h}_1 \\ \boldsymbol{h}_2^* & 1 \end{bmatrix}\begin{bmatrix} \tilde{\boldsymbol{w}}_1 \\ -\alpha_2 \end{bmatrix} = \begin{bmatrix} \mathbf{h}_1 \\ 0 \end{bmatrix}. \tag{3.63}$$

By solving Eq. (3.63),

$$\tilde{\boldsymbol{w}}_1 = \left(\boldsymbol{M} - \boldsymbol{h}_2\boldsymbol{h}_2^*\right)^{-1}\boldsymbol{h}_1 \text{ and} \tag{3.64}$$

$$\alpha_2 = \boldsymbol{h}_2^*\left(\boldsymbol{M} - \boldsymbol{h}_2\boldsymbol{h}_2^*\right)^{-1}\boldsymbol{h}_1 . \tag{3.65}$$

Since $\boldsymbol{h}_1$ and $\boldsymbol{h}_2$ are orthogonal (i.e. $\boldsymbol{h}_1^*\boldsymbol{h}_2 = 0$)

$$\alpha_2 = 0 \text{ and} \tag{3.66}$$

$$\boldsymbol{w}_1 = \boldsymbol{M}^{-1}\boldsymbol{h}_1 . \tag{3.67}$$

Minimization of MSE criterion $J(\tilde{\boldsymbol{w}}_1)$ results in MMSE interference canceller that minimizes the mean square error in c_1 without any regard of c_2.

Similarly, it can be shown to decode c_2,

$$\alpha_1 = 0 \text{ and} \tag{3.68}$$

$$\boldsymbol{w}_2 = \boldsymbol{M}^{-1}\boldsymbol{h}_2 . \tag{3.69}$$

When $\alpha_2 = 1$, let the cost function is $J(\tilde{\boldsymbol{w}}_2)$. By minimizing the cost function $J(\tilde{\boldsymbol{w}}_2)$, the interference canceller will minimize the MSE in c_2 without any regard of c_1.

From (3.66), (3.67), (3.68) and (3.69), it can be seen that MMSE interference canceller for signals from the first user will consist of two different linear weights $\boldsymbol{w}_1$ and $\boldsymbol{w}_2$ for

c_1 and c_2, respectively. Finally, the ML decoding rule to decode the signals c_1 and c_2 is given by the equation (3.70).

$$\hat{\boldsymbol{c}} = \arg\min_{\hat{\boldsymbol{c}} \varepsilon \boldsymbol{C}} \left\{ \left\| \boldsymbol{w}_1^* \boldsymbol{r} - \hat{c}_1 \right\|^2 + \left\| \boldsymbol{w}_2^* \boldsymbol{r} - \hat{c}_2 \right\|^2 \right\} \tag{3.70}$$

Hence, the decoding simplicity of the STBC is maintained with the MMSE interference suppression.

CHAPTER 4

AN ADAPTIVE RECEIVER FOR STBC MULITUSER SYSTEMS

The MMSE interference suppression for the STBC co-channel multiuser systems that is reviewed in chapter 3 assumes that the channel knowledge is available at the receiver perfectly. However, the channel has to be estimated in real situations. To overcome this problem, an adaptive receiver for STBC multiuser systems is proposed. The proposed receiver jointly estimates the tap weight vector of the receiver and mitigates the interference.

4.1 Introduction

STBC multiuser system with M synchronous co-channel users is considered where each user is equipped with N_t transmit antennas. In this scenario, there will be MN_t interfering signals arriving at the receiver. There should be $N_t(M-1)+1$ receiving antennas at the receiver to mitigate the interference from the $M-1$ co-channel users by the classical interference cancellation techniques while maintaining the transmit diversity order N_t. In [44], it has been shown that, by exploiting the temporal and spatial structure, only M receive antennas are required to suppress the interference from $M-1$ co-channel users while maintaining the transmit diversity order N_t.

A least square error (LSE) based scheme for interference suppression is introduced while RLS scheme updates the weights adaptively using training symbols. It is shown that the proposed scheme estimates the weights and suppresses the interference without any explicit knowledge of the channel and interferer.

4.2 System Model

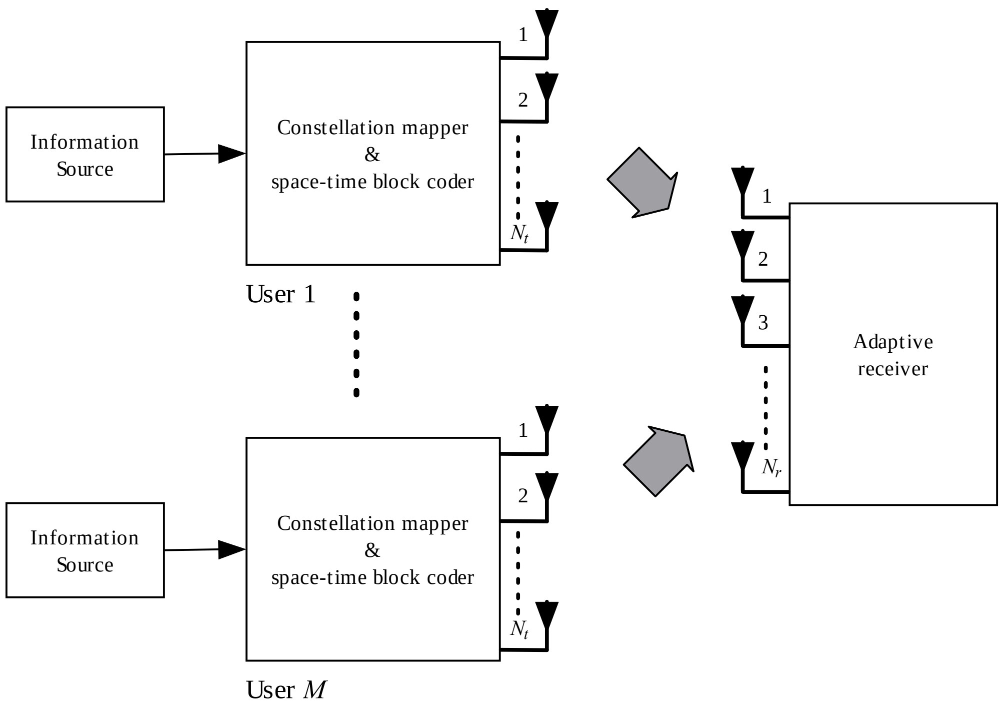

Fig. 4.1: The block diagram of space-time block coded multiuser system

Consider a multi-user environment with M synchronous co-channel users where each user is equipped with N_t transmitting antennas as shown in Fig. 4.1. The receiver is equipped with $N_r (\geq M)$ receive antennas to suppress the interference from the $M-1$ co-channel interferes without sacrifying the diversity order obtained by STBC. The space-time block code is defined by a $p \times N_t$ transmission matrix $\boldsymbol{G}_{N_t}$ that is explained in the section 2.2.3. The entries of the transmission matrix for the i^{th} user are linear combinations of the variables $c_{i,l}$ where $l = 1, 2, ...L$ and their conjugates. The rate S of the code is defined as $S = L / p$. The relationships between number of N_t, p, S and L are given in Table 4.1.

TABLE 4.1: Transmission table

N_t	p	L	S
2	2	2	1
3	8	4	½
4	8	4	½

In this paper, the unit rate code proposed in [14] for two transmit antennas and the half rate codes proposed in [16] for three and four transmit antennas are considered. The transmission matrix of the STBC is given in Table 2.4. The flat fading path gain from j^{th} transmit antenna of the i^{th} user to the $k^{th} (k = 1, 2, ...M)$ receive antenna of the receiver is defined by $h^{i}_{j,k}$. It is assumed that the receiver is equipped with $M (N_r = M)$ receive antennas.

4.2.1 Unit Rate STBC for Two Transmit Antenna Systems

The received signal at the k^{th} receive antenna during two successive symbol duration for a two transmit antenna system can be written as,

$$r_{k,1} = \sum_{i=1}^{M} (h_{1,k}^{i} c_{i,1} + h_{2,k}^{i} c_{i,2}) + \eta_{k,1} \tag{4.1}$$

and $$r_{k,2} = \sum_{i=1}^{M} (- h_{1,k}^{i} c_{i,2}^{*} + h_{2,k}^{i} c_{i,1}^{*}) + \eta_{k,2} , \tag{4.2}$$

where the noise $\eta_{k,1}$ and $\eta_{k,2}$ are modeled as a zero mean complex gaussian random variables with variance (1/2γ) per complex dimension where γ is SNR at the receiver. The received signal could be written in matrix form as,

$$\boldsymbol{r} = \begin{bmatrix} r_{11} \\ r_{12}^{*} \\ \vdots \\ r_{M,1} \\ r_{M,2}^{*} \end{bmatrix} = \begin{bmatrix} \boldsymbol{H}_1^1 & \cdots & \boldsymbol{H}_1^M \\ \vdots & \ddots & \vdots \\ \boldsymbol{H}_M^1 & \cdots & \boldsymbol{H}_M^M \end{bmatrix} \begin{bmatrix} \boldsymbol{C_1} \\ \vdots \\ \boldsymbol{C_M} \end{bmatrix} + \begin{bmatrix} \eta_{11} \\ \eta_{12}^{*} \\ \vdots \\ \eta_{M,1} \\ \eta_{M,2}^{*} \end{bmatrix}, \tag{4.3}$$

$$\boldsymbol{r} = \boldsymbol{H} \cdot \tilde{\boldsymbol{C}} + \boldsymbol{\eta} , \tag{4.4}$$

where $\boldsymbol{r} = \begin{bmatrix} r_{1,1} & r_{1,2}^{*} & \cdots & r_{M,1} & r_{M,2}^{*} \end{bmatrix}^T = \begin{bmatrix} r_1 & r_2 & \cdots & \cdots & r_{2M} \end{bmatrix}^T$,

$$\boldsymbol{H} = \begin{bmatrix} \boldsymbol{H}_1^1 & \cdots & \boldsymbol{H}_1^M \\ \vdots & \ddots & \vdots \\ \boldsymbol{H}_M^1 & \cdots & \boldsymbol{H}_M^M \end{bmatrix}, \ \tilde{\boldsymbol{C}} = \begin{bmatrix} \boldsymbol{C_1} \\ \vdots \\ \boldsymbol{C_M} \end{bmatrix}, \ \boldsymbol{\eta} = \begin{bmatrix} \eta_{1,1} & \eta_{1,2}^{*} & \cdots & \eta_{M,1} & \eta_{M,2}^{*} \end{bmatrix}^T,$$

$$\boldsymbol{H}_k^i = \begin{bmatrix} h_{1,k}^i & h_{2,k}^i \\ h_{2,k}^{i\ *} & -h_{1,k}^{i\ *} \end{bmatrix} \text{ and } \boldsymbol{C_i} = \begin{bmatrix} c_{i,1} \\ c_{i,2} \end{bmatrix}.$$

4.2.2 Half Rate STBC for Three and Four Transmit Antenna Systems

The received signal could be written in matrix form as,

$$\boldsymbol{r} = \boldsymbol{H} \cdot \tilde{\boldsymbol{C}} + \boldsymbol{\eta}\,, \tag{4.5}$$

where $\boldsymbol{r} = \begin{bmatrix} \boldsymbol{r}_1 & \cdots & \boldsymbol{r}_M \end{bmatrix}^T = \begin{bmatrix} r_1 & r_2 & \cdots & \cdots & r_{pM} \end{bmatrix}^T$,

$$\boldsymbol{r}_m = \begin{bmatrix} r_{m,1} & r_{m,2} & r_{m,3} & r_{m,4} & r_{m,5}^* & r_{m,6}^* & r_{m,7}^* & r_{m,8}^* \end{bmatrix},$$

$$\tilde{\boldsymbol{C}} = \begin{bmatrix} \boldsymbol{C_1} \\ \vdots \\ \boldsymbol{C_M} \end{bmatrix}, \boldsymbol{C}_i = \begin{bmatrix} c_{i,1} \\ c_{i,2} \\ c_{i,3} \\ c_{i,4} \end{bmatrix},$$

$$\boldsymbol{\eta} = \begin{bmatrix} \boldsymbol{\eta}_1 & \cdots & \boldsymbol{\eta}_M \end{bmatrix}^T, \boldsymbol{\eta}_m = \begin{bmatrix} \eta_{m,1} & \eta_{m,2} & \eta_{m,3} & \eta_{m,4} & \eta_{m,5}^* & \eta_{m,6}^* & \eta_{m,7}^* & \eta_{m,8}^* \end{bmatrix}$$

$$\boldsymbol{H} = \begin{bmatrix} \boldsymbol{H}_1^1 & \cdots & \boldsymbol{H}_1^M \\ \vdots & \ddots & \vdots \\ \boldsymbol{H}_M^1 & \cdots & \boldsymbol{H}_M^M \end{bmatrix} \text{ and } \boldsymbol{H}_k^i = \begin{bmatrix} \boldsymbol{h}_k^i \\ \boldsymbol{h}_k^{i*} \end{bmatrix},$$

where, $\boldsymbol{h}_k^i = \begin{bmatrix} h_{1,k}^i & h_{2,k}^i & h_{3,k}^i & 0 \\ h_{2,k}^i & -h_{1,k}^i & 0 & -h_{3,k}^i \\ h_{3,k}^i & 0 & -h_{1,k}^i & h_{2,k}^i \\ 0 & h_{3,k}^i & -h_{2,k}^i & -h_{1,k}^i \end{bmatrix}$ and $\boldsymbol{h}_k^i = \begin{bmatrix} h_{1,k}^i & h_{2,k}^i & h_{3,k}^i & h_{4,k}^i \\ h_{2,k}^i & -h_{1,k}^i & h_{4,k}^i & -h_{3,k}^i \\ h_{3,k}^i & -h_{4,k}^i & -h_{1,k}^i & h_{2,k}^i \\ h_{4,k}^i & h_{3,k}^i & -h_{2,k}^i & -h_{1,k}^i \end{bmatrix}$ for three and four transmit antenna systems respectively.

4.3 Adaptive Receiver

4.3.1 Interference Suppression Based on LSE

For notational simplicity, the received signal vector is written as,

$$\boldsymbol{r} = \begin{bmatrix} r_1 & r_2 & \cdots & \cdots & r_{pM} \end{bmatrix}^T . \tag{4.6}$$

Let us consider the first user. To perform the interference suppression, the least squared error due to the co-channel interference and the noise in the decoded symbol $c_{1,l}$ for all l should be minimized. The cost function of LSE can be written as,

$$\boldsymbol{J}(\boldsymbol{w},\boldsymbol{\alpha}) = \frac{1}{n}\sum_{k=1}^{n}\left\|\sum_{i=1}^{pM} w_i^* r_i(k) - \sum_{l=1}^{L}\alpha_l^* c_{1,l}(k)\right\|^2 = \frac{1}{n}\sum_{k=1}^{n}\left\|\boldsymbol{w}^*\boldsymbol{r}(k) - \boldsymbol{\alpha}^*\boldsymbol{C}_1(k)\right\|^2 , \tag{4.7}$$

where, $\boldsymbol{w} = \begin{bmatrix} w_1 & w_2 & \cdots & w_{pM} \end{bmatrix}^T$ and $\boldsymbol{\alpha} = \begin{bmatrix} \alpha_1 & \cdots & \alpha_L \end{bmatrix}^T$.

The weights $\boldsymbol{w}$ and $\boldsymbol{\alpha}$ are selected to minimize the cost function $\boldsymbol{J}(\boldsymbol{w},\boldsymbol{\alpha})$. In order to avoid the situation where optimal weights getting zero values, one of the weights is set to one. To decode the first symbol $c_{1,1}$, α_1 is set to 1. In this case the error criterion is

$$J(\boldsymbol{w}_1,\boldsymbol{\alpha}_1) = \frac{1}{n}\sum_{k=1}^{n}\left\|\boldsymbol{w}_1^*\boldsymbol{r}(k) - \tilde{\boldsymbol{\alpha}}_1^*\tilde{\boldsymbol{C}}_1(k) - c_{1,1}(k)\right\|^2$$

$$= \frac{1}{n}\sum_{k=1}^{n}\left\|\tilde{\boldsymbol{w}}_1^*\tilde{\boldsymbol{r}}_1(k) - c_{1,1}(k)\right\|^2 = J(\tilde{\boldsymbol{w}}_1), \tag{4.8}$$

where $\tilde{\boldsymbol{\alpha}}_1 = \begin{bmatrix} \alpha_2 & \cdots & \alpha_L \end{bmatrix}^T$, $\tilde{\boldsymbol{C}}_1(k) = \begin{bmatrix} c_{1,2}(k) & \cdots & c_{1,L}(k) \end{bmatrix}^T$, $\tilde{\boldsymbol{w}}_1 = \begin{bmatrix} \boldsymbol{w}_1^T & -\tilde{\boldsymbol{\alpha}}_1^T \end{bmatrix}^T$

and $\tilde{\boldsymbol{r}}_1(k) = \begin{bmatrix} \boldsymbol{r}^T(k) & \tilde{\boldsymbol{C}}_1(k) \end{bmatrix}^T$.

Then, $\tilde{\boldsymbol{r}}_1$ can be rewritten as,

$$\tilde{\boldsymbol{r}}_1(k) = \begin{bmatrix} \boldsymbol{H} & \boldsymbol{0}_{pM\times L-1} \\ \boldsymbol{0}_{L-1\times LM} & \boldsymbol{I}_{L-1} \end{bmatrix} \begin{bmatrix} \tilde{\boldsymbol{C}}(k) \\ \tilde{\boldsymbol{C}}_1(k) \end{bmatrix} + \begin{bmatrix} \boldsymbol{\eta}(k) \\ \boldsymbol{0}_{L-1\times 1} \end{bmatrix} = \tilde{\boldsymbol{H}}\hat{\boldsymbol{C}}_1(k) + \tilde{\boldsymbol{\eta}}(k) . \tag{4.9}$$

where $\tilde{\boldsymbol{H}} = \begin{bmatrix} \boldsymbol{H} & \boldsymbol{0}_{pM\times L-1} \\ \boldsymbol{0}_{L-1\times LM} & \boldsymbol{I}_{L-1} \end{bmatrix}$, $\hat{\boldsymbol{C}}_1(k) = \begin{bmatrix} \tilde{\boldsymbol{C}}(k) \\ \tilde{\boldsymbol{C}}_1(k) \end{bmatrix}$ and $\tilde{\boldsymbol{\eta}}(k) = \begin{bmatrix} \boldsymbol{\eta}(k) \\ \boldsymbol{0}_{L-1\times 1} \end{bmatrix}$.

We need to select $\tilde{\boldsymbol{w}}_1$ such that $J(\tilde{\boldsymbol{w}}_1)$ is minimized. Thus,

$$J(\tilde{\boldsymbol{w}}_1) = \frac{1}{n}\sum_{k=1}^{n}\left\{(\tilde{\boldsymbol{w}}_1^*\tilde{\boldsymbol{r}}(k) - c_{1,1}(k))(\tilde{\boldsymbol{w}}_1^*\tilde{\boldsymbol{r}}(k) - c_{1,1}(k))^*\right\}$$

$$= \tilde{\boldsymbol{w}}_1^*\boldsymbol{R}_1\tilde{\boldsymbol{w}}_1 + \frac{1}{n}\sum_{k=1}^{n}\left\{c_{1,1}(k).c_{1,1}^*(k)\right\} - \boldsymbol{Z}_1^*\tilde{\boldsymbol{w}}_1 - \tilde{\boldsymbol{w}}_1^*\boldsymbol{Z}_1$$

$$= \tilde{\boldsymbol{w}}_1^*\boldsymbol{R}_1\tilde{\boldsymbol{w}}_1 + E_s - \boldsymbol{Z}_1^*\tilde{\boldsymbol{w}}_1 - \tilde{\boldsymbol{w}}_1^*\boldsymbol{Z}_1 \tag{4.10}$$

where E_s is the symbol energy, $\boldsymbol{R}_1 = \frac{1}{n}\sum_{k=1}^{n}\left\{\tilde{\boldsymbol{r}}_1(k).\tilde{\boldsymbol{r}}_1^*(k)\right\}$ and $\boldsymbol{Z}_1 = \frac{1}{n}\sum_{k=1}^{n}\left\{\tilde{\boldsymbol{r}}_1(k).c_{1,1}^*(k)\right\}$.

Defining $\boldsymbol{V}_2 = \begin{bmatrix} \boldsymbol{0}_{L-1\times 1} & \boldsymbol{I}_{L-1} & \boldsymbol{0}_{L-1\times LM-L} \end{bmatrix}^T$ and $\boldsymbol{V}_1 = \begin{bmatrix} 1 & \boldsymbol{0}_{1\times pM-1} \end{bmatrix}^T$, it can easily shown that

$$\boldsymbol{R}_1 = \frac{1}{n}\tilde{\boldsymbol{H}}\sum_{k=1}^{n}\left\{\hat{\boldsymbol{C}}_1(k)\hat{\boldsymbol{C}}_1^*(k)\right\}\tilde{\boldsymbol{H}}^* + \frac{1}{n}\sum_{k=1}^{n}\tilde{\boldsymbol{\eta}}(k)\tilde{\boldsymbol{\eta}}^*(k)$$

$$= E_s\tilde{\boldsymbol{H}}\begin{bmatrix} \boldsymbol{I}_{pM} & \boldsymbol{V}_2 \\ \boldsymbol{V}_2^T & \boldsymbol{I}_{L-1} \end{bmatrix}\tilde{\boldsymbol{H}}^* + N_0\begin{bmatrix} \boldsymbol{I}_{pM} & \boldsymbol{0}_{pM\times L-1} \\ \boldsymbol{0}_{L-1\times pM} & \boldsymbol{0}_{L-1\times L-1} \end{bmatrix}$$

$$= E_s\begin{bmatrix} \boldsymbol{M} & \boldsymbol{h}_2 \\ \boldsymbol{h}_2^* & \boldsymbol{I}_{L-1} \end{bmatrix}, \tag{4.11}$$

where $\boldsymbol{M} = \boldsymbol{H}\boldsymbol{H}^* + \frac{1}{\gamma}\boldsymbol{I}_{pM} = \frac{1}{n}\sum_{k=1}^{n}\boldsymbol{r}(k).\boldsymbol{r}^*(k)$; $\boldsymbol{h}_2$ is equal to $\boldsymbol{H}$ excluding the first column and $\boldsymbol{h}_1$ is the first column of $\boldsymbol{H}$. And

$$\boldsymbol{Z}_1 = \frac{1}{n}\sum_{k=1}^{n}\left\{\tilde{\boldsymbol{r}}_1(k)c_{1,1}^*(k)\right\} = E_s\tilde{\boldsymbol{H}}\begin{bmatrix}\boldsymbol{V}_1 \\ \boldsymbol{0}_{L-1\times 1}\end{bmatrix} = E_s\begin{bmatrix}\boldsymbol{h}_1 \\ \boldsymbol{0}_{L-1\times 1}\end{bmatrix}. \quad (4.12)$$

Differentiating (4.10) w.r.t. $\tilde{\boldsymbol{w}}_1$ the normal equations of the LSE can be written as

$$\boldsymbol{R}_1\tilde{\boldsymbol{w}}_1 = \boldsymbol{Z}_1. \quad (4.13)$$

By substituting (4.11) and (4.12) in (4.13),

$$\begin{bmatrix}\boldsymbol{M} & \boldsymbol{h}_2 \\ \boldsymbol{h}_2^* & \boldsymbol{I}_{L-1}\end{bmatrix}\begin{bmatrix}\boldsymbol{w}_1 \\ -\boldsymbol{\alpha}_2\end{bmatrix} = \begin{bmatrix}\boldsymbol{h}_1 \\ \boldsymbol{0}_{L-1\times 1}\end{bmatrix}. \quad (4.14)$$

By solving the equation (4.14),

$$\boldsymbol{w}_1 = \left[\boldsymbol{M} \quad -\boldsymbol{h}_2\boldsymbol{h}_2^*\right]^{-1}\boldsymbol{h}_1 \text{ and} \quad (4.15)$$

$$\boldsymbol{\alpha}_2 = \boldsymbol{h}_2^*\left[\boldsymbol{M} \quad -\boldsymbol{h}_2\boldsymbol{h}_2^*\right]^{-1}\boldsymbol{h}_1. \quad (4.16)$$

Since $\boldsymbol{h}_1$ and $\boldsymbol{h}_2$ are orthogonal (i.e. $\boldsymbol{h}_1^*\boldsymbol{h}_2 = \boldsymbol{0}$)

$$\boldsymbol{\alpha}_2 = \boldsymbol{0} \text{ and} \quad (4.17)$$

$$\boldsymbol{w}_1 = \boldsymbol{M}^{-1}\boldsymbol{h}_1. \quad (4.18)$$

Minimization of LSE criterion $J(\tilde{\boldsymbol{w}}_1)$ results in an interference canceller that minimizes the least square error in $c_{1,1}(k)$ without any regard of $c_{1,2}(k)$, $c_{1,3}(k)\ldots$ $c_{1,L}(k)$. Similarly it can be shown that, for $l = 1, 2....L$,

$$\boldsymbol{w}_l = \boldsymbol{M}^{-1}\boldsymbol{h}_l \quad (4.19)$$

From (4.19), we can show that the LSE interference suppression will consist of L different linear weight $\boldsymbol{w}_l$ for $c_{1,l}$. Hence the decoding simplicity of the space-time block codes is still maintained in our proposed LSE interference suppression scheme.

4.3.2 Adaptive Weight Estimation

Evaluation of $\boldsymbol{M}^{-1}$ in LSE is computationally expensive and perfect channel knowledge should be available at the receiver. To overcome this, the RLS algorithm is used to estimate the weights and to suppress the interference. The derivation of RLS algorithm from LSE is given in section 3.2.4.

The computation starts with known initial conditions $\boldsymbol{P}(0)$ and $\hat{\boldsymbol{w}}(0)$. The information from the received training signal $\boldsymbol{r}(n)$ and known desired symbols $d(n)$ is used to update the weights $\hat{\boldsymbol{w}}(n)$. The summary of the RLS algorithm is shown in Table 4.2.

Table 4.2: RLS Algorithm for weight estimation

Initialization:

$$\hat{\boldsymbol{w}}(0) = 0$$

$\boldsymbol{P}(0) = \delta^{-1}\boldsymbol{I}$, where δ is small positive constant

Data:

$\boldsymbol{r}(n)$ = input vector

$d(n)$ = desired response

Computation:

For n=0,1,2,3… Compute

$$\boldsymbol{k}(n) = \frac{\boldsymbol{P}(n-1)\boldsymbol{r}(n)}{\lambda + \boldsymbol{r}^*(n)\boldsymbol{P}(n-1)\boldsymbol{r}(n)},$$

$$\xi^*(n) = d(n) - \hat{\boldsymbol{w}}^*(n-1)\boldsymbol{r}(n),$$

$$\hat{\boldsymbol{w}}(n+1) = \hat{\boldsymbol{w}}(n) + \mu\boldsymbol{r}(n)e^*(n),$$

and

$$\boldsymbol{P}(n) = \lambda^{-1}\boldsymbol{P}(n-1) - \lambda^{-1}\boldsymbol{k}(n)\boldsymbol{r}^*(n)\boldsymbol{P}(n-1)$$

where $(\lambda < 1)$ is called the forgetting factor.

Our proposed adaptive receiver is shown in Fig. 4.2 where the weight $\boldsymbol{w}_l$, for $l = 1, 2, ...L$, are estimated using the RLS algorithm.

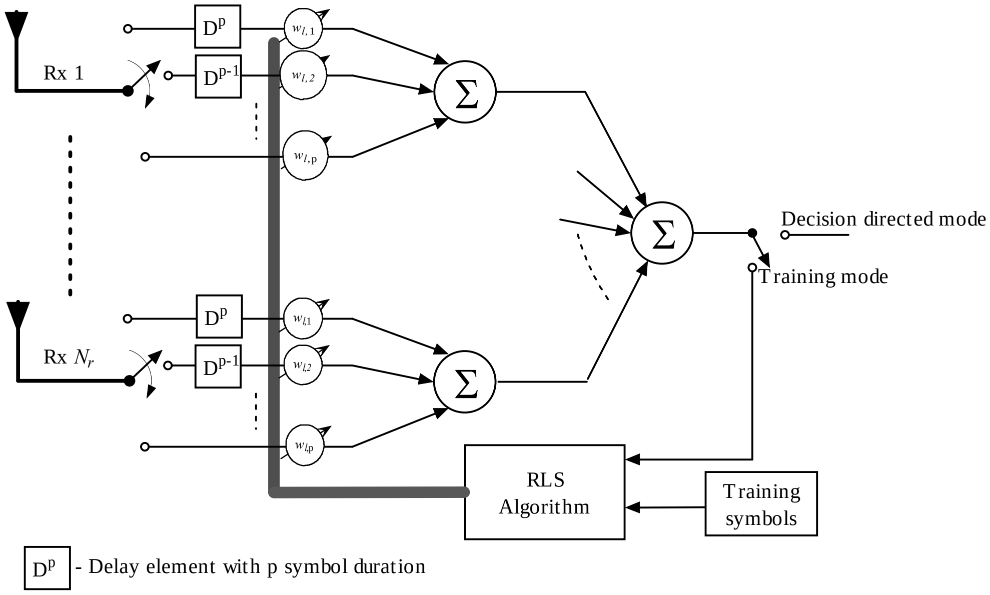

Fig. 4.2: The Structure of proposed adaptive receiver

Furthermore, to estimate the weights, the RLS algorithm uses the training symbols that are accommodated in the initial part of each transmitted frame of symbols. Generally, the RLS algorithm converges in about $2N$ iterations, where N is the number of taps in the interference canceller and is equal to pM in the proposed scheme. The structure of the transmitted frame of symbols is given in Fig. 4.3.

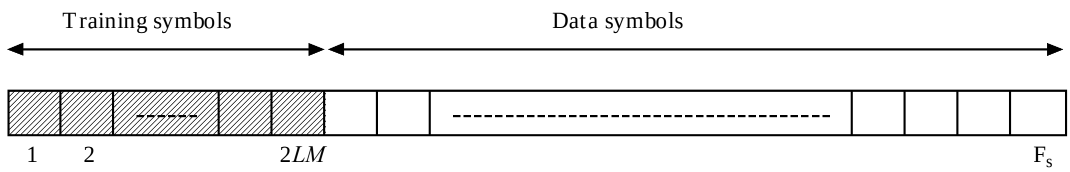

Fig. 4.3: Structure of transmitted frame of symbols

According to Table 4.1, $2pM$ iterations need $2LM$ training symbols to estimate the weights. The estimated weights are utilized in the decision directed mode to decode the remaining message of a frame using the following ML decoding rule:

$$\hat{\boldsymbol{c}} = \arg\min_{\hat{\boldsymbol{c}}\varepsilon\boldsymbol{C}} \left\{ \sum_{l=1}^{L} \left\| \boldsymbol{w}_l^* \boldsymbol{r} - \hat{c}_{1,l} \right\|^2 \right\} \tag{4.20}$$

4.4 Simulation Results

The simulation results for the proposed adaptive receiver are provided in this section. In these simulations, M space-time block coded synchronous co-channel users, where each user is equipped with N_t transmit antennas, are considered. The receiver is equipped with $N_r = M$ receive antennas. The channel is assumed to be quasi-static flat fading. Further, the constants δ, λ and μ (step size of the LMS algorithm) take the values 0.0001, 0.99 and 0.001 respectively.

To study the convergence behavior of the RLS and the LMS algorithms, the ensemble-averaged learning curves of the algorithms are shown in Fig. 4.4 for a unit rate STBC system for two users. The complexity of the LMS algorithm is less than the RLS algorithm; but conversely, the RLS algorithm converges much faster than the LMS algorithm. The fast convergence will reduce the number of training symbols needed to estimate the tap weights. As a result, the RLS algorithm is a suitable candidate for the adaptive receiver. The RLS converges in $2N$ iterations, where N is the number of taps in the interference canceller.

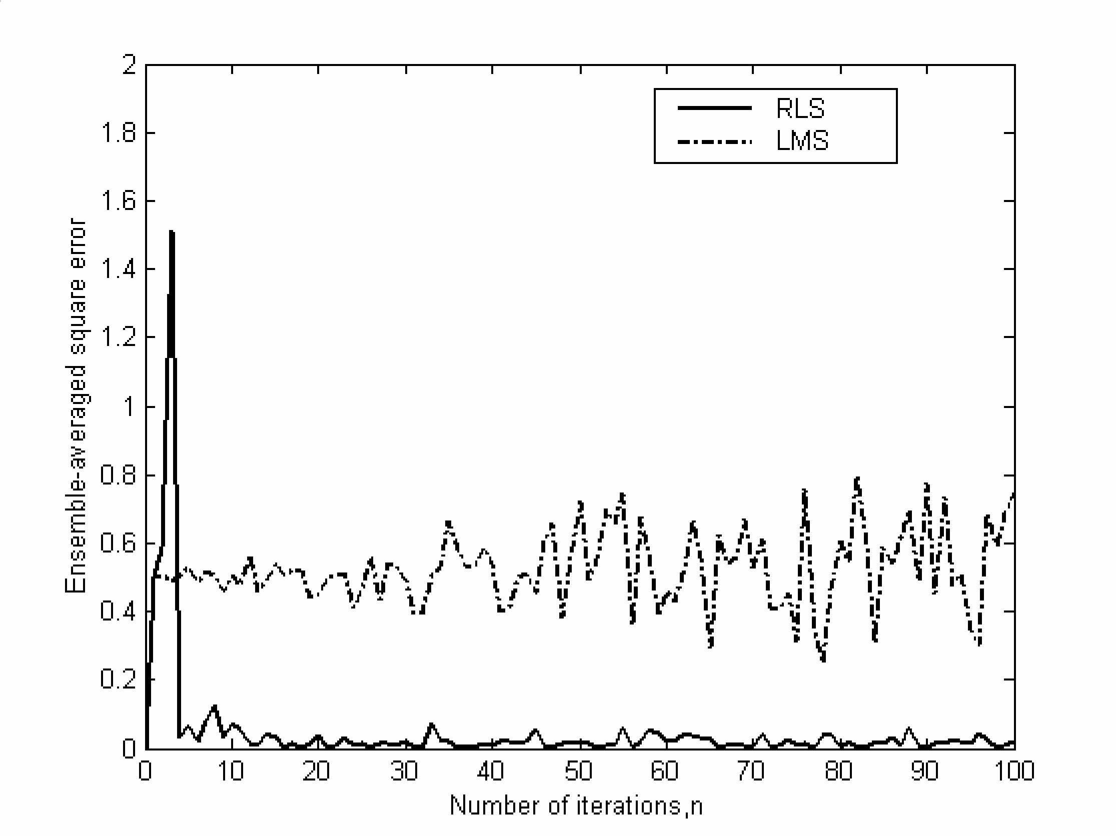

Fig. 4.4: Comparison of the learning curves of RLS and LMS algorithms

The BER performance of the adaptive receiver for two users with different values of SIR is shown in Fig. 4.5. 8-PSK-modulated signals with a frame of 126 symbols including 16 training symbols are considered in the simulations. At low SNR, BER performance of the proposed receiver, when both user has equal power, is 1-2 dB worse than both STBC with 2 transmit and 1 receive antenna system described in [14] and MMSE interference suppression scheme described in [44] at SIR=0 dB. The performance degradation is due to the weight estimation errors of RLS algorithm where as a perfect channel state information is assumed in [44]. However, the BER performance of the proposed receiver and the STBC system with 2 transmit and 1 receive antenna is similar at high SNR. This is due to the less weight estimation error of RLS algorithm at high SNR. Furthermore, the proposed receiver performs better than the differential detection scheme proposed in [21], where the channel state information is not required. The BER performance for SIR=10 and 20 dB is also shown in Fig. 4.5, where performance of the desired user increases with SIR. One can see from the Fig. 4.5 that the BER performance of the desired user approaches the performance of STBC with two transmit and two receive antennas when the interference decreases (SIR increases).

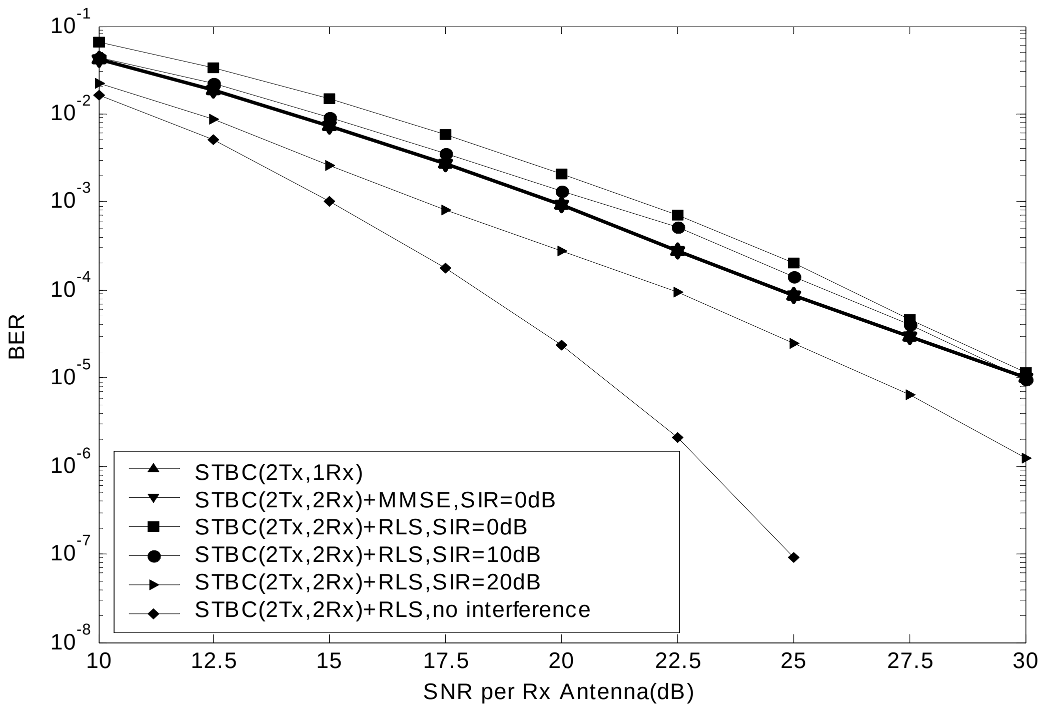

Fig. 4.5: The BER performance of MMSE and adaptive receiver of two co-channel user system

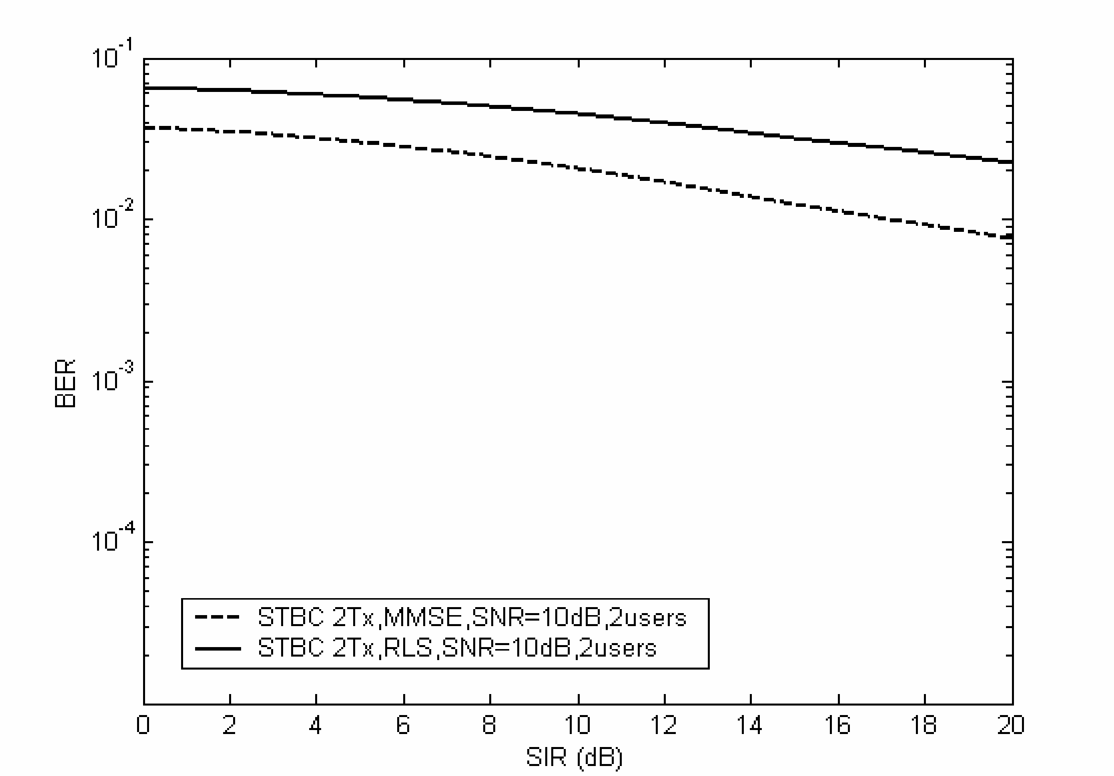

Fig. 4.6: The BER performance of adaptive receiver and MMSE scheme for two co-channel user system at fixed SNR

In Fig. 4.6, the BER performance of the proposed adaptive receiver is shown together with the MMSE scheme proposed in [44] for two users. The SNR is fixed at 10 dB and SIR is varied from 0 dB to 20 dB. As stated earlier, due to weight estimation errors, the performance of the adaptive receiver is little bit worse than that of the MMSE method. The figure also shows the change of BER with SIR of the adaptive receiver is similar that of the MMSE scheme. However, the complexity of the proposed receiver is much less than that of the MMSE scheme.

Fig. 4.7 shows the performance of the proposed receiver together with MMSE scheme for different number of users. It can be observed that the BER of both the proposed receiver and the MMSE scheme remain almost constant. This is due to the fact that the interference is well suppressed when the number of receive antennas is equal to the number of users.

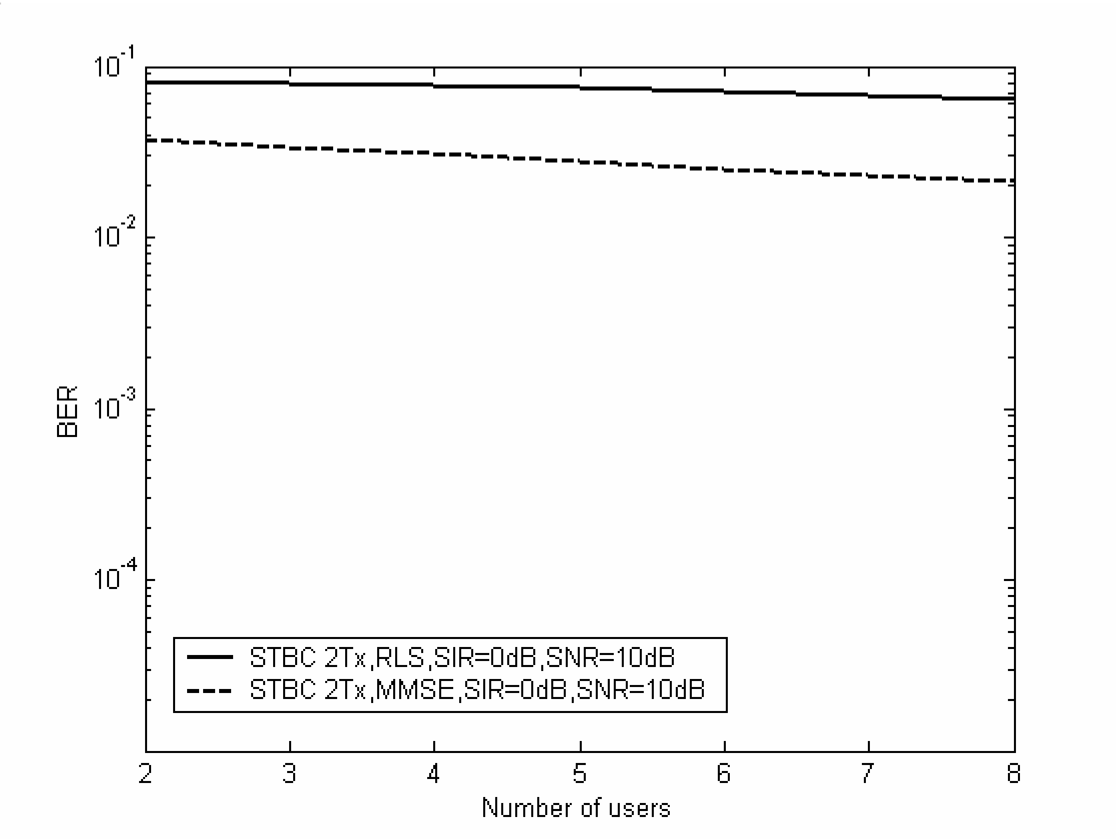

Fig. 4.7. The BER performance of adaptive receiver and MMSE scheme for multiple co-channel user systems at fixed SIR and SNR

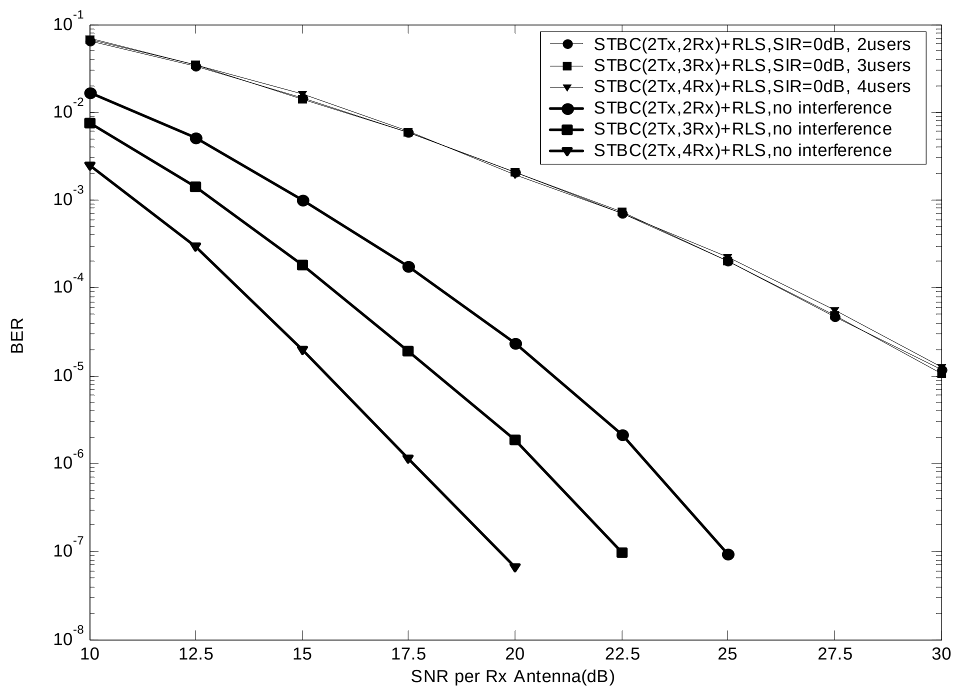

Fig. 4.8: The BER performance of adaptive receiver of two, three and four co-channel user systems

Furthermore, it can be observed from Fig. 4.8 that the interference is well suppressed and diversity order is maintained when the number of receive antennas is equal to number of users. When interferes disappear, the BER performance of the desired user approaches the performance of a STBC system that has a diversity order of $2M$.

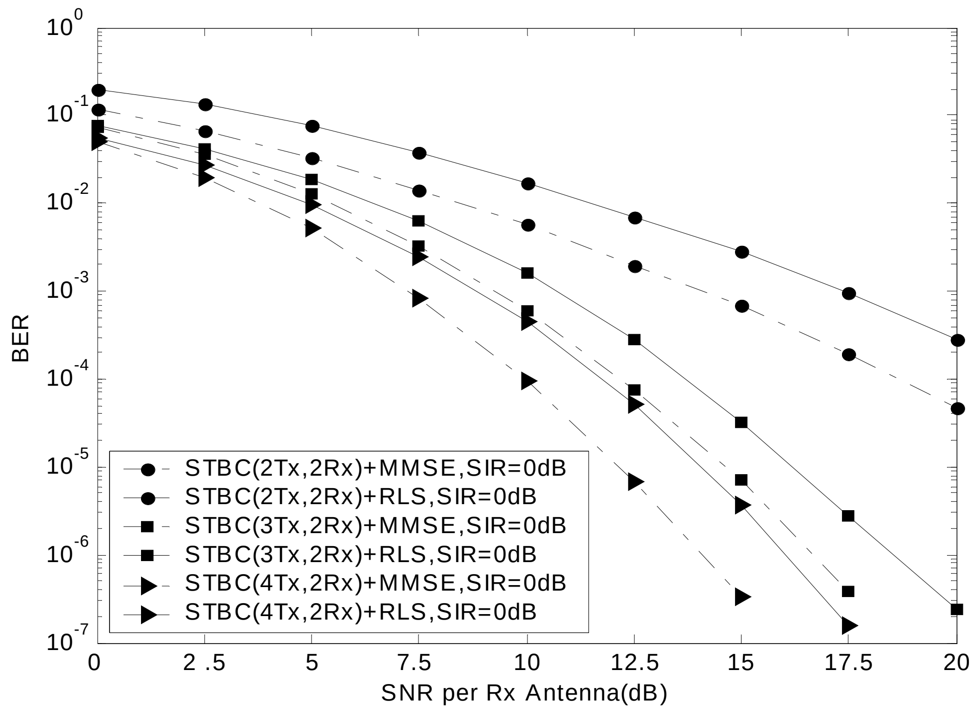

Fig. 4.9: The BER performance of MMSE and adaptive receiver of two co-channel user systems; each user equipped with two, three and four transmit antennas, 2 bits/s/Hz

Fig. 4.9 shows the BER performance of the proposed receiver and the MMSE scheme for the transmission of 2 bits/s/Hz using two, three, and four transmit antennas for two synchronous users. The system with two transmits antennas employs QPSK constellation and full rate orthogonal space-time block code described in [14]. The systems with three and four antennas employ 16-QAM constellations and the half rate orthogonal space-time block codes proposed in [16]. The performance of the receiver increases with SNR and number of transmits antennas. Due to the errors in weight estimation, the BER performance of the proposed receiver is 1 to 2 dB worse than the MMSE schemes described in [44,45].

4.5 Application of Adaptive Receiver in Multirate Multiuser Systems

In this section, the adaptive receiver scheme is applied for STBC multirate multiuser systems. By employing different modulation techniques such as BPSK, QPSK, 8-PSK and 16 QAM, the variable bit rate can be achieved where symbol duration is the same for all users. The simplicity and receiver structure is still maintained and BER performance is compared with MMSE interference suppression scheme.

4.5.1 Simulation Results of Multirate Systems

In the simulations, multirate systems having two co-channel users and each equipped with two transmit antennas uses the space-time block code, are considered. The receiver is equipped with two receive antennas. The channel is assumed to be quasi-static flat fading. Each user is employed with different modulation techniques such as BPSK, QPSK and 8PSK. It is also assumed that each user has the same symbol rate & equal power and the transmission is synchronized.

The Fig. 4.10 and Fig. 4.11 show the BER performance of the multirate multiuser systems for both MMSE and adaptive receiver (LSE-RLS) schemes. In Fig. 4.10, first and second users equipped with 8-PSK and QPSK modulation schemes, respectively. As a result, the first user can support 3 bits/Hz/sec data rate and second user can support 2 bits/Hz/Sec data rate. In Fig. 4.11, first user and second user utilize the BPSK and QPSK modulation schemes to achieve the 1 bit/Hz/Sec and 2 bits/Hz/Sec data rates, respectively.

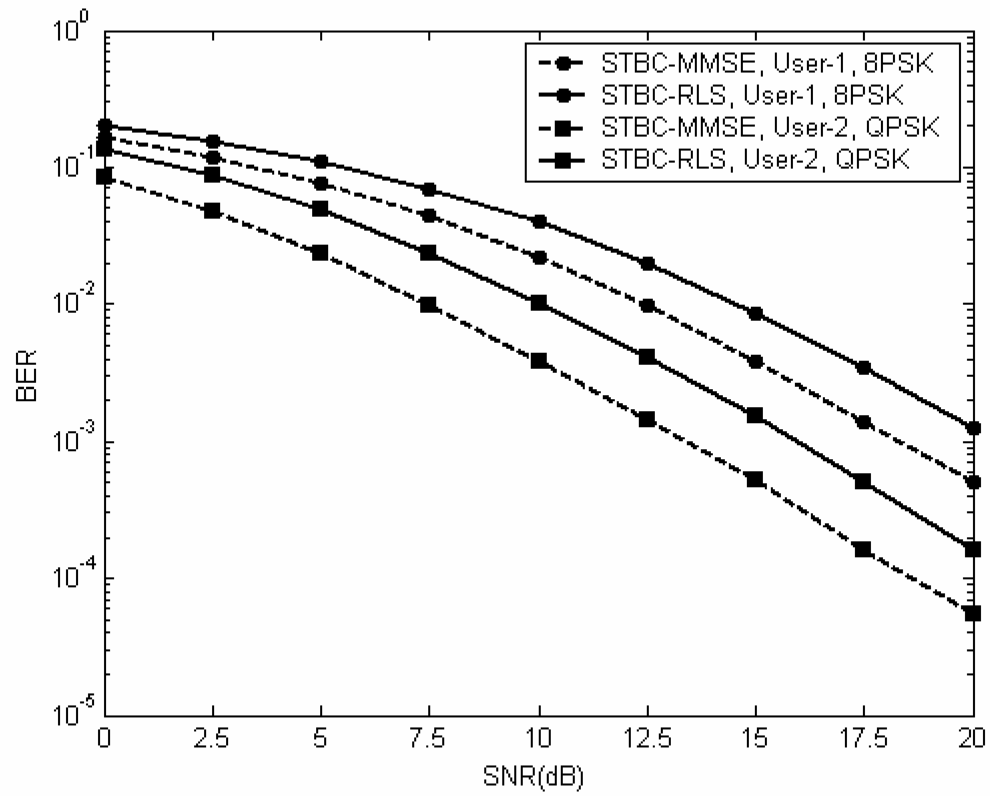

Fig. 4.10 The BER performance of STBC MMSE and LSE-RLS schemes for multirate multiuser systems using 8PSK and QPSK Modulations

The BER performance of user 1 deviates from the BER performance of user 2. This is due to the fact that the minimum Euclidian distance between the adjacent symbols in the signal constellation diagram differs from one modulation scheme to other modulation scheme where the average energy of the signal constellations is equal. As expected, the performance of multirate scheme is similar to the performance of equal rate system for both MMSE and LSE-RLS interference suppression schemes that is given in chapter 4.

Due to the channel estimation errors the BER performance of the Adaptive receiver scheme is little bit worse than that of the MMSE scheme. On the other hand, as mentioned earlier, the channel state information is not important for our adaptive receiver scheme which also has less computational complexity than the MMSE scheme.

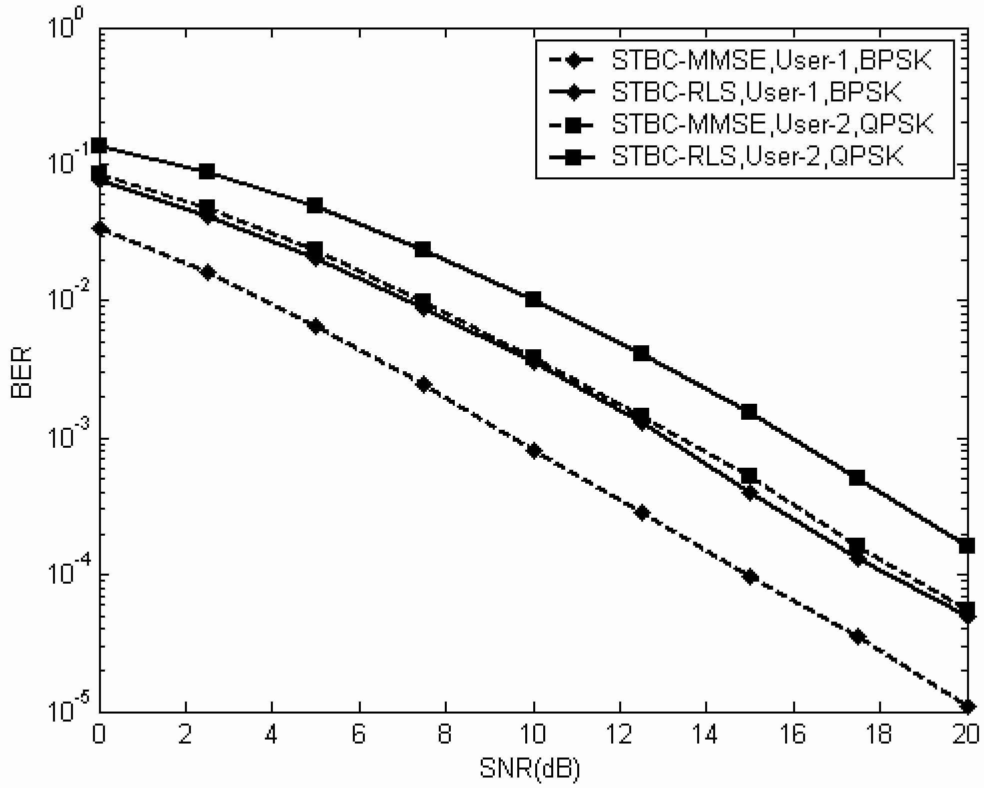

Fig. 4.11 The BER performance of STBC MMSE and LSE-RLS schemes for multirate multiuser systems using BPSK and QPSK Modulations

4.6 Summary

In this chapter, an adaptive receiver for STBC multiuser systems with two, three and four transmit antennas is proposed. The proposed adaptive receiver suppresses the interference without any explicit knowledge of the channel and interference. In particular, the weights are estimated adaptively by RLS algorithm using training symbols. However, simulation results show that BER performance of the proposed receiver is 1 to 2 dB worse than that of the MMSE scheme described in [44] and [45] where the channel information is assumed to be known perfectly at the receiver. In addition, the complexity of this scheme is less than that of [44] and [45], where the computation of matrix inverse is needed. Furthermore, the proposed receiver performs better than the differential detection scheme proposed in [21,22], where the channel knowledge is not required. The proposed adaptive receiver scheme applied for multirate multiuser systems. Multirate is achieved by employing different modulation schemes for each user. The performance of the multi rate system is shown as the performance of equal rate user system.

This receiver can be used in a multiuser system as well as in a system that needs high throughput for individual user. Particularly, the proposed receiver could be used to increase the capacity and data rates in future wireless communication systems.

CHAPTER 5

AN ITERATIVE INTERFERENCE CANCELLATION RECEIVER FOR STBC MULTIRATE MULTIUSER SYSTEMS

5.1 Introduction

Next generation wireless communication systems are expected to provide a variety of services integrating voice, data and video. Because of inherent difference in source rates of different services, designing a multirate communication system is an imperative [55]. In CDMA systems, multicode transmission and variable processing gain are the two popular schemes to adopt the multirate services. In Wideband TDMA systems, variable burst/symbol rate is proposed for one transmit one receive antenna systems [55]. In this chapter, iterative interference cancellation schemes are proposed for both equal and variable rate multiuser STBC systems.

5.2 Equal Rate System

The proposed iterative scheme first decodes the first user by applying MMSE interference cancellation technique given in section 3.3 and then uses the first user data to decode the data of the second user. Subsequently, the second user data is used, in an iterative manner, to decode the data of the first user and the first user data is then used to decode the data of the second user.

5.2.1 System Model

Consider two synchronous users employing equal symbol rates and each equipped with 2 transmit antennas. The receiver is equipped with 2 receive antennas. Two receive antennas are used to suppress the interference using minimum mean square estimation (MMSE) method and then used to increase the diversity using an iterative interference cancellation (IIC) method.

The flat fading path gain from i^{th} transmit antenna of first user to j^{th} receive antenna of the receiver is defined by $h_{i,j}$. Similarly, $g_{i,j}$ is the path gain between i^{th} transmit antenna of second user and j^{th} receive antenna of the receiver. The symbols $\{c_1, c_2\}$ and $\{s_1, s_2\}$ are transmitted from the first and second users respectively. Let the received signals from the first receive antenna during the first and second symbol intervals be $r_{1,1}$ and $r_{2,1}$ respectively, where

$$r_{1,1} = h_{1,1}c_1 + h_{2,1}c_2 + g_{1,1}s_1 + g_{2,1}s_2 + \eta_{1,1} \tag{5.1}$$

$$r_{1,2} = -h_{1,1}c_2^* + h_{2,1}c_1^* - g_{1,1}s_2^* + g_{2,1}s_1^* + \eta_{1,2}\,. \tag{5.2}$$

The noise $\eta_{i,j}$ is modeled as a zero mean complex gaussian process with variance $(1/2\gamma)$ per complex dimension. Similarly, let the received signal from the second antenna over two consecutive symbol periods be $r_{2,1}$ and $r_{2,2}$ respectively, where

$$r_{2,1} = h_{1,2}c_1 + h_{2,2}c_2 + g_{1,2}s_1 + g_{2,2}s_2 + \eta_{2,1} \tag{5.3}$$

$$r_{22} = -h_{12}c_2^* + h_{22}c_1^* - g_{12}s_2^* + g_{22}s_1^* + \eta_{22}\ . \tag{5.4}$$

Now, the overall signal vector can be written as,

$$\boldsymbol{r} = \begin{bmatrix} r_{1,1} \\ r_{1,2}^* \\ r_{2,1} \\ r_{2,2}^* \end{bmatrix} = \begin{bmatrix} h_{1,1} & h_{2,1} & g_{1,1} & g_{2,1} \\ h_{2,1}^* & -h_{1,1}^* & g_{2,1}^* & -g_{1,1}^* \\ h_{1,2} & h_{2,2} & g_{1,2} & g_{2,2} \\ h_{2,2}^* & -h_{1,2}^* & g_{2,2}^* & -g_{2,1}^* \end{bmatrix} \begin{bmatrix} c_1 \\ c_2 \\ s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} \eta_{1,1} \\ \eta_{1,2}^* \\ \eta_{2,1} \\ \eta_{2,2}^* \end{bmatrix}. \tag{5.5}$$

Above equation can be further simplified to,

$$\boldsymbol{r = H.\tilde{C} + \eta} \tag{5.6}$$

where $\boldsymbol{H} = \begin{bmatrix} h_{1,1} & h_{2,1} & g_{1,1} & g_{2,1} \\ h_{2,1}^* & -h_{1,1}^* & g_{2,1}^* & -g_{1,1}^* \\ h_{1,2} & h_{2,2} & g_{1,2} & g_{2,2} \\ h_{2,2}^* & -h_{1,2}^* & g_{2,2}^* & -g_{2,1}^* \end{bmatrix}$, $\boldsymbol{\tilde{C}} = \begin{bmatrix} c_1 \\ c_2 \\ s_1 \\ s_2 \end{bmatrix}$ and $\boldsymbol{\eta} = \begin{bmatrix} \eta_{1,1} & \eta_{1,2}^* & \eta_{1,2} & \eta_{2,2}^* \end{bmatrix}^T$.

5.2.2 Iterative Interference Cancellation

5.2.2.1 Decoding the First User using MMSE

In order to decode the signal c_1 and c_2 using the MMSE interference cancellation technique as described in section 3.3, the fact that the first and second column of $\boldsymbol{H}$ should be orthogonal is used.

Therefore, $\boldsymbol{w}_{c1} = \boldsymbol{M}_1^{-1}\boldsymbol{h}_1$ and $\boldsymbol{w}_{c2} = \boldsymbol{M}_1^{-1}\boldsymbol{h}_2$ where $\boldsymbol{M}_1 = \boldsymbol{HH}^* + \frac{1}{\gamma}\mathbf{I}_4$

$$\boldsymbol{\hat{c}} = \arg\min_{\hat{c}_1 \varepsilon \boldsymbol{C}} \left\{ \left\| \boldsymbol{w}_{c1}\boldsymbol{r} - \hat{c}_1 \right\|^2 + \left\| \boldsymbol{w}_{c2}\boldsymbol{r} - \hat{c}_2 \right\|^2 \right\} \tag{5.7}$$

5.2.2.2. Iterative Decoding of the Second User

The data of the second user can be decoded by using the two consecutive decoded symbols $\hat{c}_1$ and $\hat{c}_2$ of the first user. After subtracting the first user's contributions to the received signal, new receive signal can be written as,

$$\hat{r}_1 = r_{1,1} - h_{1,1}\hat{c}_1 - h_{2,1}\hat{c}_2 \tag{5.8}$$

$$\hat{r}_2 = r_{1,2}^* - h_{2,1}^*\hat{c}_1 + h_{1,1}^*\hat{c}_2 \tag{5.9}$$

$$\hat{r}_3 = r_{2,1} - h_{1,2}\hat{c}_1 - h_{2,2}\hat{c}_2 \tag{5.10}$$

$$\hat{r}_4 = r_{2,2}^* - h_{2,2}^*\hat{c}_1 + h_{1,2}^*\hat{c}_2 \tag{5.11}$$

For real signal constellations, it can be written as

$$\hat{\boldsymbol{r}}_1 = \begin{bmatrix}\hat{r}_1 & \hat{r}_2 & \hat{r}_3 & \hat{r}_4\end{bmatrix}^T = \tilde{\boldsymbol{G}}\tilde{\boldsymbol{s}} + \tilde{\boldsymbol{\eta}}, \tag{5.12}$$

where $\tilde{\boldsymbol{G}} = \begin{bmatrix} g_{1,1} & g_{2,1} \\ g_{2,1}^* & -g_{1,1}^* \\ g_{1,2} & g_{2,2} \\ g_{2,2}^* & -g_{1,2}^* \end{bmatrix}$, $\tilde{\boldsymbol{s}} = \begin{bmatrix} s_1 \\ s_2 \end{bmatrix}$ and $\tilde{\boldsymbol{\eta}} = \begin{bmatrix} \eta_{1,1} \\ \eta_{1,2}^* \\ \eta_{2,1} \\ \eta_{2,2}^* \end{bmatrix}$.

Using the ML decoding rule, the second user symbols can be obtained from the following equations:

$$\hat{\boldsymbol{s}} = \begin{bmatrix}\hat{s}_1 & \hat{s}_2\end{bmatrix}^T = \arg\min_{\hat{s}\varepsilon \boldsymbol{S}} \left\{\hat{\boldsymbol{r}}_1 - \tilde{\boldsymbol{G}}\hat{\boldsymbol{s}}\right\} \tag{5.13}$$

5.2.2.3. Iterative Decoding of First User

Similarly, the user's decoded symbols can be used to iteratively obtain the better estimates for the first user. Thus the subtracted receive signal can be written as,

$$\hat{r}_5 = r_{1,1} - g_{1,1}\hat{s}_1 - g_{2,1}\hat{s}_2 \tag{5.14}$$

$$\hat{r}_6 = r_{1,2}^* - g_{2,1}^*\hat{s}_1 + g_{1,1}^*\hat{s}_2 \tag{5.15}$$

$$\hat{r}_7 = r_{2,1} - g_{1,2}\hat{s}_1 - g_{2,2}\hat{s}_2 \quad (5.16)$$

$$\hat{r}_8 = r_{2,2}^* - g_{2,2}^*\hat{s}_1 + g_{1,2}^*\hat{s}_2 \quad (5.17)$$

For real signal constellations, it can be written as

$$\hat{\boldsymbol{r}}_2 = \begin{bmatrix} \hat{r}_5 & \hat{r}_6 & \hat{r}_7 & \hat{r}_8 \end{bmatrix}^T = \tilde{\boldsymbol{H}}\tilde{\boldsymbol{c}} + \tilde{\boldsymbol{\eta}}, \quad (5.18)$$

where $\tilde{\boldsymbol{H}} = \begin{bmatrix} h_{1,1} & h_{2,1} \\ h_{2,1}^* & -h_{1,1}^* \\ h_{1,2} & h_{2,2} \\ h_{2,2}^* & -h_{1,2}^* \end{bmatrix}$, $\tilde{\boldsymbol{c}} = \begin{bmatrix} c_1 \\ c_2 \end{bmatrix}$ and $\tilde{\boldsymbol{\eta}} = \begin{bmatrix} \eta_{1,1} \\ \eta_{1,2}^* \\ \eta_{2,1} \\ \eta_{2,2}^* \end{bmatrix}$

Now the ML decoding rule is reduced to,

$$\hat{\boldsymbol{c}} = \begin{bmatrix} \hat{c}_1 & \hat{c}_2 \end{bmatrix}^T = \arg\min_{\hat{c}\varepsilon \boldsymbol{C}} \left\{ \hat{\boldsymbol{r}}_2 - \tilde{\boldsymbol{H}}\hat{\boldsymbol{c}} \right\} \quad (5.19)$$

The steps given in section 5.2.2.2 and 5.2.2.3 can be used iteratively to increase the BER performance of the system.

5.3 Multirate System

In this section, we extend the above mentioned iterative interference cancellation scheme to a multirate system. First, the scheme decodes the high data rate user by applying the MMSE interference cancellation technique and then uses decoded symbols of the high data rate user to decode the low data rate user. Subsequently, in an iterative manner the decoded symbol of the low data rate user is used to decode the high data rate user. These iterations can be repeated to improve the BER performance of the systems.

5.3.1 System Model

According to the STBC scheme, two signals are transmitted simultaneously from both transmit antennas at a given symbol period. The signals transmitted from first and second antenna of the first user denoted by c_1 and c_2 respectively. During the next symbol period, $-c_2^*$ and c_1^* are transmitted from first and second antenna respectively. In the mean time, second user also uses the same STBC as above but symbol rate is different from the first user. At a given symbol period, s_1 and s_2 are transmitted from first and second transmit antennas of the second user. Next symbol period of second user, $-s_2^*$ and s_1^* are transmitted from first and second antenna respectively. As shown in Table 5.1, the symbol duration of first user is T_1 and the symbol duration of second user is $T_2(=2T_1)$.

Table 5.1: Transmitting Symbols

Antenna	Time			
	t	$t+T_1$	$t+2T_1$	$t+3T_1$
1st User Tx 1	c_1	$-c_2^*$	c_3	$-c_4^*$
1st User Tx 2	c_2	c_1^*	c_4	c_3^*
2nd User Tx 1	s_1	s_1	$-s_2^*$	$-s_2^*$
2nd User Tx 2	s_2	s_2	s_1^*	s_1^*

Let $r_{j,1}, r_{j,2}, r_{j,3}$ and $r_{j,4}$ be the received signal at the j^{th} receive antenna over four consecutive symbol periods (T_1). Received signal can be written as,

$$r_{1,1} = h_{1,1}c_1 + h_{2,1}c_2 + g_{1,1}s_1 + g_{2,1}s_2 + \eta_{1,1} \tag{5.20}$$

$$r_{1,2} = h_{2,1}c_1^* - h_{1,1}c_2^* + g_{1,1}s_1 + g_{2,1}s_2 + \eta_{1,2} \tag{5.21}$$

$$r_{2,1} = h_{1,2}c_1 + h_{2,2}c_2 + g_{1,2}s_1 + g_{2,2}s_2 + \eta_{2,1} \tag{5.22}$$

$$r_{2,2} = h_{2,2}c_1^* - h_{1,2}c_2^* + g_{1,2}s_1 + g_{2,2}s_2 + \eta_{2,2} \tag{5.23}$$

$$r_{1,3} = h_{1,1}c_3 + h_{2,1}c_4 + g_{2,1}s_1^* - g_{1,1}s_2^* + \eta_{1,3} \tag{5.24}$$

$$r_{1,4} = h_{2,1}c_3^* - h_{1,1}c_4^* + g_{2,1}s_1^* - g_{1,1}s_2^* + \eta_{1,4} \tag{5.25}$$

$$r_{2,3} = h_{1,2}c_3 + h_{2,2}c_4 + g_{2,2}s_1^* - g_{1,2}s_2^* + \eta_{2,3} \tag{5.26}$$

$$r_{2,4} = h_{2,2}c_3^* - h_{1,2}c_4^* + g_{2,2}s_1^* - g_{1,2}s_2^* + \eta_{2,4} \tag{5.27}$$

where $h_{1,j}$ and $h_{2,j}$ are the flat fading path gains for the first user from first and second transmit antennas to the j^{th} receive antenna of the receiver respectively. Similarly, $g_{1,j}$ and $g_{2,j}$ are the flat fading path gains for the second user. The noise $\eta_{1,j}$ and $\eta_{2,j}$ are modeled as a zero mean complex gaussian process with variance $(1/2\gamma)$ per complex dimension where γ is SNR at the j^{th} receive antenna receiver.

The overall signal can be written in vector form as,

$$\tilde{\boldsymbol{r}}_1 = \begin{bmatrix} r_{1,1} \\ r_{1,2}^* \\ r_{2,1} \\ r_{2,2}^* \end{bmatrix} = \tilde{\boldsymbol{H}}_1 \tilde{\boldsymbol{C}}_1 + \tilde{\boldsymbol{\eta}}_1 \tag{5.28}$$

where, $\tilde{\boldsymbol{H}}_1 = \begin{bmatrix} h_{1,1} & h_{2,1} & g_{1,1} & g_{2,1} \\ h_{2,1}^* & -h_{1,1}^* & g_{1,1}^* & g_{2,1}^* \\ h_{1,2} & h_{2,2} & g_{1,2} & g_{2,2} \\ h_{2,2}^* & h_{1,2}^* & g_{1,2}^* & g_{2,2}^* \end{bmatrix}$, $\tilde{\boldsymbol{C}}_1 = \begin{bmatrix} c_1 \\ c_2 \\ s_1 \\ s_2 \end{bmatrix}$ and $\tilde{\boldsymbol{\eta}}_1 = \begin{bmatrix} \eta_{1,1} \\ \eta_{1,2}^* \\ \eta_{2,1} \\ \eta_{2,2}^* \end{bmatrix}$.

5.3.2 Iterative Interference Cancellation

5.3.2.1 Decoding the High Data Rate User using MMSE

To decode the signal c_1 and c_2 using the MMSE interference cancellation technique, first and second column of $\tilde{\boldsymbol{H}}_1$ should be orthogonal.

Therefore, $\boldsymbol{w}_{c1} = \boldsymbol{M}_1^{-1}\boldsymbol{h}_1$ and $\boldsymbol{w}_{c2} = \boldsymbol{M}_1^{-1}\boldsymbol{h}_2$ where $\boldsymbol{M}_1 = \tilde{\boldsymbol{H}}_1\tilde{\boldsymbol{H}}_1^* + \frac{1}{\gamma}\mathbf{I}_4$

$$\hat{\boldsymbol{c}}_1 = \arg\min_{\hat{c}_1 \varepsilon \boldsymbol{C}} \left\{ \left\| \boldsymbol{w}_{c1}\tilde{\boldsymbol{r}}_1 - \hat{c}_1 \right\|^2 + \left\| \boldsymbol{w}_{c2}\tilde{\boldsymbol{r}}_1 - \hat{c}_2 \right\|^2 \right\} \tag{5.29}$$

Similarly, the signal c_3 and c_4 can be decoded as described below,

$$\tilde{\boldsymbol{r}}_2 = \begin{bmatrix} \boldsymbol{r}_3 \\ \boldsymbol{r}_4 \end{bmatrix} = \tilde{\boldsymbol{H}}_2\tilde{\boldsymbol{C}}_2 + \tilde{\boldsymbol{\eta}}_2 \tag{5.30}$$

$$\tilde{\boldsymbol{H}}_2 = \begin{bmatrix} h_{1,1} & h_{2,1} & g_{2,1} & -g_{1,1} \\ h_{2,1}^* & -h_{1,1}^* & g_{2,1}^* & -g_{1,1}^* \\ h_{1,2} & h_{2,2} & g_{2,2} & -g_{1,2} \\ h_{2,2}^* & h_{1,2}^* & g_{2,2}^* & -g_{1,2}^* \end{bmatrix}, \ \tilde{\boldsymbol{C}}_2 = \begin{bmatrix} c_1 \\ c_2 \\ s_1^* \\ s_2^* \end{bmatrix} \text{ and } \tilde{\boldsymbol{\eta}}_2 = \begin{bmatrix} \eta_{1,3} \\ \eta_{1,4}^* \\ \eta_{2,3} \\ \eta_{2,4}^* \end{bmatrix}$$

Then, $\boldsymbol{w}_{c3} = \boldsymbol{M}_2^{-1}\boldsymbol{h}_1$ and $\boldsymbol{w}_{c4} = \boldsymbol{M}_2^{-1}\boldsymbol{h}_2$ where $\boldsymbol{M}_2 = \tilde{\boldsymbol{H}}_2\tilde{\boldsymbol{H}}_2^* + \frac{1}{\gamma}\mathbf{I}_4$

$$\hat{\boldsymbol{c}}_2 = \arg\min_{\hat{c}_2 \varepsilon \boldsymbol{C}} \left\{ \left\| \boldsymbol{w}_{c3}\tilde{\boldsymbol{r}}_2 - \hat{c}_3 \right\|^2 + \left\| \boldsymbol{w}_{c4}\tilde{\boldsymbol{r}}_2 - \hat{c}_4 \right\|^2 \right\} \tag{5.31}$$

5.3.2.2. Iterative Decoding of the Low Data Rate User

The low data rate user can be decoded by using the four consecutive decoded symbols of the high data rate user.

To decode the signal s_1 and s_2, the decoded signals $\hat{c}_1, \hat{c}_2, \hat{c}_3$ and $\hat{c}_4$ have to be used.

$$\hat{r}_1 = r_{1,1} - h_{1,1}\hat{c}_1 - h_{2,1}\hat{c}_2 \quad (5.32)$$

$$\hat{r}_2 = r_{1,4}^* - h_{2,1}^*\hat{c}_3 + h_{1,1}^*\hat{c}_4 \quad (5.33)$$

$$\hat{r}_3 = r_{2,1} - h_{1,2}\hat{c}_1 - h_{2,2}\hat{c}_2 \quad (5.35)$$

$$\hat{r}_4 = r_{2,4}^* - h_{2,2}^*\hat{c}_3 - h_{1,2}^*\hat{c}_4 \quad (5.36)$$

For real signal constellations,

$$\tilde{\boldsymbol{r}}_3 = \begin{bmatrix} \hat{r}_1 & \hat{r}_2 & \hat{r}_3 & \hat{r}_4 \end{bmatrix}^T = \tilde{\boldsymbol{G}}\tilde{\boldsymbol{s}} + \tilde{\boldsymbol{\eta}}_3, \quad (5.37)$$

where $\tilde{\boldsymbol{G}} = \begin{bmatrix} g_{1,1} & g_{2,1} \\ g_{2,1}^* & -g_{1,1}^* \\ g_{1,2} & g_{2,2} \\ g_{2,2}^* & -g_{1,2}^* \end{bmatrix}$, $\tilde{\boldsymbol{s}} = \begin{bmatrix} s_1 \\ s_2 \end{bmatrix}$ and $\tilde{\boldsymbol{\eta}}_3 = \begin{bmatrix} \eta_{1,1} \\ \eta_{1,4}^* \\ \eta_{2,1} \\ \eta_{2,4}^* \end{bmatrix}$,

and the ML decoding rule is given by,

$$\hat{\boldsymbol{s}} = \begin{bmatrix} \hat{s}_1 & \hat{s}_2 \end{bmatrix}^T = \arg\min_{\hat{s}\varepsilon \boldsymbol{S}} \left\{ \tilde{\boldsymbol{r}}_3 - \tilde{\boldsymbol{G}}\hat{\boldsymbol{s}} \right\} \quad (5.38)$$

5.3.2.3. Iterative Decoding of the High Data Rate User

As given in section 5.3.2.2, we can use iterative scheme to decode the high data rate user.

To decode the signal c_1 and c_2, we have to use the decoded signals $\hat{s}_1$ and $\hat{s}_2$.

$$\hat{r}_5 = r_{1,1} - g_{1,1}\hat{s}_1 - g_{2,1}\hat{s}_2 \quad (5.39)$$

$$\hat{r}_6 = r_{1,2}^* - g_{1,2}^*\hat{s}_1^* - g_{2,1}^*\hat{s}_2^* \quad (5.40)$$

$$\hat{r}_7 = r_{2,1} - g_{1,2}\hat{s}_1 - g_{2,2}\hat{s}_2 \quad (5.41)$$

$$\hat{r}_8 = r_{2,2}^* - g_{1,2}^*\hat{s}_1^* - g_{2,2}^*\hat{s}_2^* \quad (5.42)$$

For real signal constellations,

$$\tilde{\boldsymbol{r}}_4 = \begin{bmatrix} \hat{r}_5 & \hat{r}_6 & \hat{r}_7 & \hat{r}_8 \end{bmatrix}^T = \tilde{\boldsymbol{H}}\tilde{\boldsymbol{c}} + \tilde{\boldsymbol{\eta}}_4 , \quad (5.43)$$

where $\tilde{\boldsymbol{H}} = \begin{bmatrix} h_{1,1} & h_{2,1} \\ h_{2,1}^* & -h_{1,1}^* \\ h_{1,2} & h_{2,2} \\ h_{2,2}^* & -h_{1,2}^* \end{bmatrix}$, $\tilde{\boldsymbol{c}} = \begin{bmatrix} c_1 \\ c_2 \end{bmatrix}$ and $\tilde{\boldsymbol{\eta}}_4 = \begin{bmatrix} \eta_{1,1} \\ \eta_{1,2}^* \\ \eta_{2,1} \\ \eta_{2,2}^* \end{bmatrix}$,

and the ML decoding rule is given by,

$$\hat{\boldsymbol{c}} = \begin{bmatrix} \hat{c}_1 & \hat{c}_2 \end{bmatrix}^T = \arg\min_{\hat{c}\varepsilon \boldsymbol{C}} \left\{ \tilde{\boldsymbol{r}}_4 - \tilde{\boldsymbol{H}}\hat{\boldsymbol{c}} \right\} \quad (5.44)$$

Similarly, the signal c_3 and c_4 can be decoded using iterative decoding.

5.4 Simulation Results

In this section, the simulation results for the proposed schemes are presented. In these simulations, both equal rate and multirate STBC systems are considered. Both systems have each user equipped with two transmit antennas uses the space-time block code. The receiver is equipped with two receive antennas. The channel is assumed to be quasi-static flat fading. Real signal constellation such as BPSK is used. In the equal rate system, both users have same symbol duration T_1. On the other hand, the first user has T_1 symbol duration and the second user has $T_2 (= 2T_1)$ symbol duration in the multirate system.

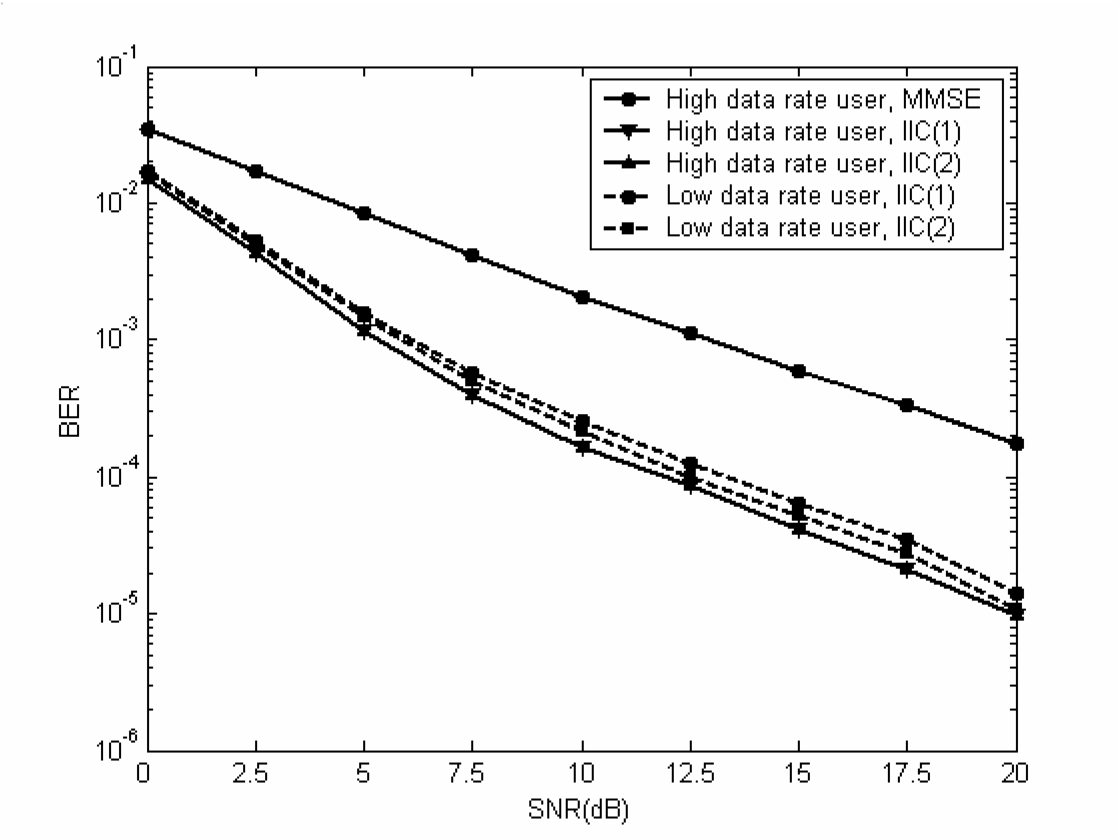

Fig. 5.1 The BER performances of STBC-IIC multirate multiuser systems

Fig. 5.1 shows the BER performance of the proposed IIC scheme for STBC multirate multiuser systems. Proposed IIC scheme performs better than the MMSE scheme described in [44] and it approaches the diversity order of two transmit and two receive antenna system. As expected, the performance margin diminishes with number of iterations. Further, the Fig. 5.1 illustrates that the performance improvement by second iteration is not very significant. Thus, one has to consider the tradeoff between the iterations for better performance, the processing delay and receiver complexity when designing a system.

In Fig. 5.2, BER performance of both equal rate and multirate systems are compared. The BER performance of multirate systems is little bit worse than that of equal rate when we use the MMSE scheme. This is because of the loss of orthogonality in multirate systems. On other hand, when we use the proposed IIC the BER performance of the multirate system is as same as an equal rate system.

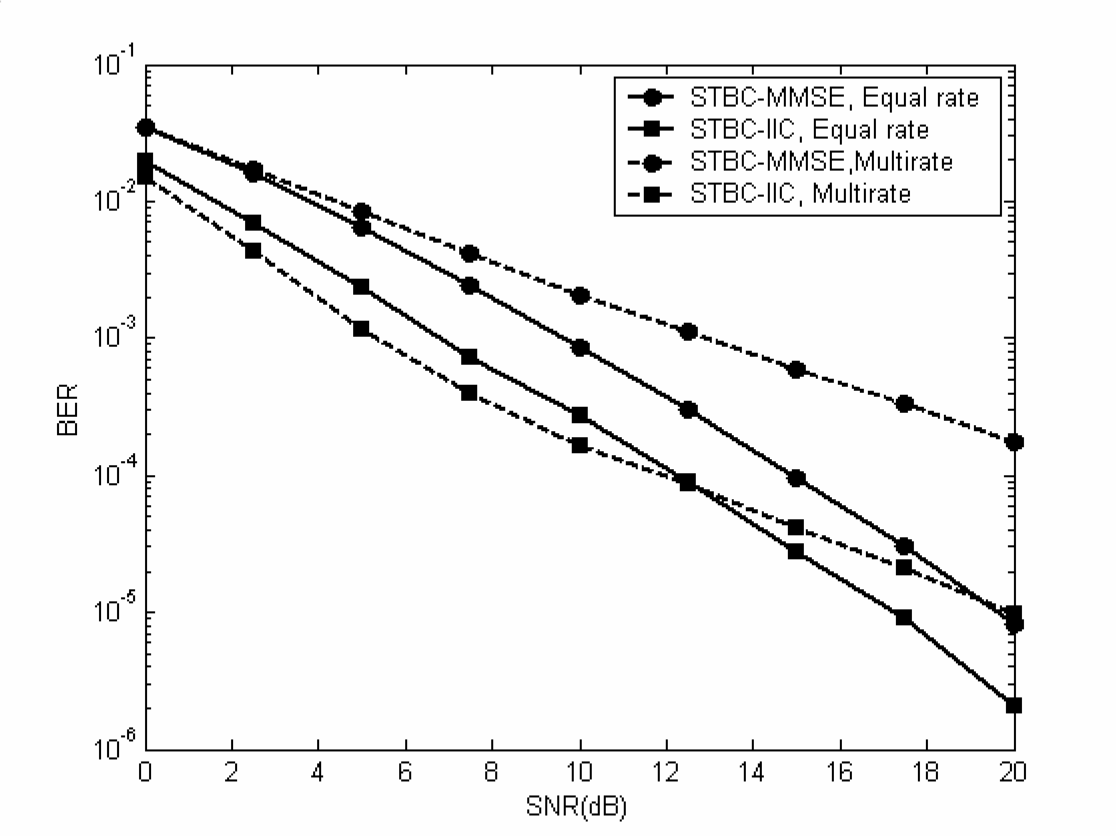

Fig. 5.2 The BER performances of STBC-IIC for both multirate and equal rate multiuser systems

5.5 Summary

In this chapter, an iterative interference cancellation STBC scheme for both multirate and equal rate multiuser is developed. The proposed IIC scheme is based on minimum mean square estimation (MMSE) interference cancellation and maximum likelihood detection. In particular, the first user is decoded using MMSE IC scheme and then the decoded data of the first user is used to decode the second user data using an IIC scheme. Similarly, the IIC scheme can be used to decode the first user. In the simulation results, it has been shown that the bit error rate performance of the proposed IIC is better than that of the conventional non-iterative MMSE scheme. In addition, the proposed scheme performs well not only in multirate multiuser systems but also in equal rate multiuser systems. As expected, the performance margin diminishes with number of iterations.

CHAPTER 6

CONCLUSIONS AND FUTURE WORK

6.1 Conclusions

In this thesis, a detailed study of interference cancellation schemes for STBC multiuser systems was presented. Various practical limitation and complexity issues, which are associated with existing interference cancellation schemes, were discussed. Proposed adaptive receiver in STBC multiuser systems offers the LSE interference cancellation and RLS weight estimation. In addition, an adaptive interference cancellation scheme for STBC multirate multiuser systems was proposed and that of the proposed scheme was analyzed via simulations.

Firstly, an adaptive receiver for STBC multiuser systems with two, three and four transmit antennas is proposed. In this scheme, interference is suppressed without any explicit knowledge of the channel and interference. In particular, the weights are estimated adaptively by RLS algorithm using training symbols. Simulation results show that the proposed receiver yields a similar performance as the MMSE scheme described in [44] and [45] where the channel information is assumed to be known perfectly at the receiver. In addition, the complexity of this scheme is less than that of [44,45], where the computation of matrix inverse is needed. Furthermore, the proposed receiver performs better than the differential detection scheme proposed in [21,22], where the channel knowledge is not required. This receiver can be used in a multiuser system as well as in a system that needs high throughput for individual user. Particularly, the proposed adaptive receiver could be used to increase the capacity and data rate in future communication systems.

Secondly, an iterative interference cancellation receiver for STBC multirate multiuser is developed based on MMSE interference cancellation. In particular, the high rate user is decoded first using MMSE IC scheme and then the decoded data of high rate user is used to decode the low rate user data using IIC scheme. Furthermore, IIC scheme can be used iteratively to decode both users. The BER performance of the proposed IIC scheme receiver is better than that of the conventional non-iterative MMSE scheme. In addition, the proposed scheme performs well not only in multirate multiuser systems but also in equal rate multiuser systems. As expected, the performance margin diminishes with number of iterations. The proposed IIC receiver can be used in next generation mobile communication systems where multirate services are provided.

6.2 Future Work

Some topics of interest that may be considered for further research are listed below.

Firstly, Reducing the BER performance margin between adaptive receiver and MMSE IC scheme is imperative. It can be achieved by implementing an adaptive algorithm to vary the initialization constant δ of the RLS algorithm. Generally, δ is a small positive constant for high SNR and large positive constant for low SNR.

Secondly, more considerations may be given to reduce the weight estimation errors and computational complexity of adaptive receiver. To make the estimation errors small, number of training symbols may be increased. Increasing training symbols is not a feasible solution because it increases the overhead of the transmission. On the other hand, using the algorithms that have better numerical stability and better convergence speed than RLS based algorithms can reduce estimation errors. The RLS algorithm has numerical instability, which arises because of the way in which the Riccati difference equation is formulated [49]. The numerical instability of the RLS algorithm is eliminated in QR-RLS algorithm, which is accomplished by working with the incoming data matrix via QR decomposition, rather than working with time-average correlation matrix of input data as in the standard RLS algorithm. Furthermore, the QR-RLS algorithms are very easy to implement in the digital hardware by using systolic array. Thus, QR-RLS algorithm can be used instead of RLS algorithm in the proposed adaptive receiver, which will reduce the computational complexity and better BER performance.

Another topic of interest is related to the application of adaptive receiver in frequency selective channels. Due to multipath delays of transmitted signals in frequency selective channel, the proposed adaptive receiver cannot be used directly in frequency selective channel environments. In the literature, the techniques that are used to transmit and receive signals in frequency selective channels are different from flat fading channels. As described in chapter 1, STBC-OFDM [30-33] and STBC-FDE [34-40] techniques can be used with proposed adaptive receiver to decode the signals without any explicit knowledge of the channel state information.

Finally, both adaptive receiver and IIC receiver could be utilized with some modifications in CDMA based communication systems. Future study on this subject has to include the analytical studies and simulation results on the performance and complexity features of the proposed schemes.

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