LAYERED SPACE TIME ARCHITECTURES FOR MIMO WIRELESS CHANNELS

by

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THESIS DISSERTATION SUBMITTED IN PARTIAL FULFILLMENT
OF
THE REQUIREMENTS FOR THE DEGREE OF

DOCTOR OF PHILOSOPHY

Queen Mary
University of London

Summer 2006

DECLARATION

I hereby declare that the work presented in this thesis is my own work unless otherwise stated.

Signed: ………………………………………….

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ABSTRACT

The demand for mobile communication systems with high data rates and improved link quality for a variety of applications has dramatically increased in recent years. New concepts and methods are necessary in order to cover this huge demand, which counteract (or take advantage of) the impairments of the mobile communication channel, attempting to optimize the use of limited resources such as bandwidth and power. Multiple antenna systems are an efficient means for increasing the performance, but in order to utilize the huge potential of multiple antenna concepts it is necessary to resort to new transmit strategies, referred to as Space-Time Codes, which, in addition to the time and spectral domain, also use the spatial domain.

Layered Space-Time (LST) codes were proposed as a scheme to improve spatial multiplexing gain provided by MIMO systems. However, an exhaustive amount of research work has been put into to try to improve their poor order of diversity. Space-time block codes (STBC) on the other hand are able to provide maximal diversity gain with little or no spatial multiplexing gain. In this thesis, the issues relating to deficiencies in both these schemes are addressed. Methods are proposed to improve the performance of LST using a novel approach called known interference layer BLAST (KIL-BLAST). The advantages of this new scheme are investigated in the presence of imperfect channel state information at the receiver. The rate-diversity trade-off of KIL-BLAST is also introduced. Further, in an effort to improve rate-diversity trade-off, Layered Space-time block codes (LSTBC) are introduced as a novel approach which allows the co-existence of LST and STBC. Through the generation of simulated numerical results LSTBC is shown to be a good hybrid capable of extracting the advantages of both schemes to improve the rate-diversity trade-offs.

DEDICATION

To my mum and dad for their love and support.

ACKNOWLEDGEMENTS

Amongst the challenges of independent research are motivation and dedication. The contributions of some key people in both my academic and personal life has lead to the completion of this work.

I would like to thank my parents who have been my main pillar of encouragement. To my brothers Bilal and Hassan, who made some immeasurable sacrifices to ensure that I have all the support that is required, a debt I can never repay. To Maheen, who was the main driving force behind this significant achievement and who's encouragement made this work pleasurable. Finally to Emaan my 6 month old niece for typing the last few words of my thesis without slobbering all over my computer and Bhabi for giving us such a wonderful gift.

The support and commitment by Dr. Schormans from the start of my PhD. through to the end has been most significant. His advice on issues related both to my day to day work and my long-term goals have always helped me stay in check and confident with my work.

I would also like to thank Dr. Chen for his support and encouragement in the early days of my PhD. His encouragement got me started on my research work and his guidance assisted me in choosing the most challenging areas of research.

Mention has to also be made of Ahtisham and Basit who kept encouraging me during social times and never letting me forget what they expect of me.

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LIST OF SYMBOLS

$\mathbf{G}_n$	n-antenna STBC generator matrix
$\mathbf{x}^n$	n^{th} column of transmit matrix
$\mathbf{X}$	Transmission Matrix
$\mathbf{X}^H$	Hermitian Transpose of $\mathbf{X}$
$\mathbf{I}_n$	$n\times n$ identity matrix
H	Channel Matrix
$\mathbf{H}_d^{i-1}$	Deflated Channel Matrix
$h_{i,j}$	Channel fading coefficient for the path from antenna i to receive antenna j
r_t^j	Received signal at antenna j at time t
d	Distance
min	Minimum of argument
max	Maximum of argument
P	Probability

List of Symbols

n_t^j	AWGN noise signal at antenna j at time t
n_R	Number of receive antennas
n_T	Number of transmit antennas
L	Frame length
s^2	Variance
q	Number of bits per symbol
A^T	Transpose of Matrix A
P	Total transmit power
e	Channel estimation matrix
I_k	Interference on layer k
r	Rank of matrix
l	Eigenvalue
E	Codeword difference matrix
N_o	Noise Power Spectral Density
w	Weighting filter

GLOSSARY

3G	THIRD GENERATION
AWGN	ADDITIVE WHITE GAUSSIAN NOISE
b/s/HZ	BITS PER SECOND PER HERTZ
BER	BIT ERROR RATE
BS	BASE STATION
CDMA	CODE DIVISION MULTIP LE ACCESS
CSI	CHANNEL STATE INFORMATION
dB	DECIBEL
EP	ERROR PROPAGATION
FEC	FORWARD ERROR CORRECTION
FER	FRAME ERROR RATE
GSM	GLOBAL SYSTEM FOR MOBILE
HLST	HORIZONTAL LAYERED SPACE-TIME
LST	LAYERED SPACE-TIME
LSTBC	LAYERED SPACE TIME BLOCK CODING
KIL-BLAST	KNOWN-INTERFERENCE-LAYER BLAST
MIMO	MULTIPLE-INPUT-MULTIPLE-OUTPUT
ML	MAXIMUM LIKELIHOOD
MUD	MULTI-USER DETECTION

MMSE	MINIMUM MEAN SQUARED ERROR
PEP	PAIRWISE ERROR PROBABILITY
PSK	PHASE SHIFT KEYING
QAM	QUADRATURE AMPLITUDE MODULATION
QPSK	QUADRATURE PHASE SHIFT KEYING
Rx	RECEIVER
SER	SYMBOL ERROR RATE
SISO	SINGLE-INPUT-SINGLE-OUTPUT
SNR	SIGNAL TO NOISE RATIO
STC	SPACE-TIME CODING
STBC	SPACE-TIME BLOCK CODING
STTCM	SPACE-TIME TRELLIS CODED MODULATION
STTUTC	SPACE-TIME TURBO TRELLIS CODING
Tx	TRANSMITTER
UMTS	UNIVERSAL MOBILE TELECOMMUNICATIONS SYSTEM
V-BLAST	VERTICAL BELL LABS LAYERED SPACE-TIME
ZF	ZERO FORCING

CHAPTER 1 INTRODUCTION

1.1 Overview

The rapid growth of wireless voice subscribers, the growth of the Internet, and the increasing use of portable computing devices suggest that the demand for wireless Internet access will rise rapidly over the next few years. Rapid progress in digital and RF technology is making possible highly compact and integrated terminal devices, and the introduction of sophisticated wireless data software is making wireless Internet access more user-friendly and providing more value. Meeting the quality of service (QoS) currently provided by Internet technology for high data rate applications on wireless terminals suggests that data throughput in the region of 2-5Mbps over the air interface will be required in macrocellular environments and up to 10Mbps in microcellular and indoor environments.

The famous work of Shannon in [1] provided in a relatively abstract form the foundations for design of efficient wireless communication systems. Teletar [8] and Foschini [13] demonstrated the capability of wireless channels to achieve significantly high data rates which is known as the Shannon capacity of wireless channels. Due to channel impairments such as multipath propagation, the capacity of wireless channels cannot be fully exploited. Temporal coding such as convolution codes for forward error correction (FEC) at the transmitter and Viterbi decoding at the receiver in GSM does provide some form of protection against such impairments. The redundancy introduced by this form of temporal coding significantly reduces the spectral efficiency in bits/sec/Hz (b/s/Hz) of the available bandwidth. Let us consider a system which provides a nominal spectral efficiency of 2 b/s/Hz with a half rate convolution code. This would mean that achieving a link rate of 2 Mbps would require a bandwidth of 1MHz. Without the temporal coding, a spectral efficiency of 4 b/s/Hz would halve the required bandwidth. With the scarcity of the wireless spectrum, employing highly redundant techniques such as temporal coding will not suffice for spectral efficient communications. Hence alternative ways would have to be developed which are capable of achieving high spectral efficiencies.

1.2 Utilizing Diversity to Improve Wireless Link Performance

The basic definition of diversity is the presence of a wide range of variation in the qualities or attributes under discussion. For wireless links, this translates to having a number of copies of the desired signal at the receiver which vary in terms of signal to noise ratio (SNR). Examples of the methods that can be used to provide this form of diversity are:

- Temporal diversity: Replicas of the same information signal are transmitted in different time slots, where the separation between the time slots is greater than the coherence time of the channel [2].

- Frequency diversity: Replicas of the same information signal are transmitted in different frequency bands, where the separation between the frequency bands is greater than the coherence bandwidth of the channel [3].
- Space diversity: Replicas of the same information signal are transmitted over uncorrelated spatial channels. In order to form such channels, the antennas are required to be positioned sufficiently apart [4].
- Polarization Diversity: This is an example of space diversity where horizontal and vertical polarization signals are transmitted by two different polarized antennas and received by two different polarized antennas [5].

In practice however, all forms of diversity schemes cannot be implemented because of the nature of the channel available and of the signalling employed. For instance, a slow fading channel that has a long coherence time cannot support temporal diversity with practical interleaving depths. Similarly, frequency diversity is not feasible when the coherence bandwidth of the channel is comparable to the signal bandwidth. However, irrespective of the channel characteristics, space diversity can always be efficiently implemented as long as the antenna elements are sufficiently placed apart, so as to have uncorrelated fading channels at the transmitter and/or receiver. Another attractive feature is the fact that unlike time and frequency diversity, space-diversity does not introduce any loss in bandwidth efficiency.

Traditionally, space diversity techniques have been applied to the receiver end, in practical communication systems. For example, the use of multiple base station antennas for receive diversity in the uplink, which helps to achieve power-efficient data transmission by reducing interference [6]. Conversely, when applying this to present day small receiver systems such as mobile handsets, several parameters must be considered, such as size, complexity and signal correlation, which result in the limited use of multiple receive antennas [7]. This motivates the need to move complexity to the base station, and is done by introducing several transmit antennas. When these multiple antennas are applied at both the transmitter and receiver end, a multiple-input-multiple-output (MIMO) system is created. Information theoretic contributions by Telatar [8], and Foschini and Gans [9] demonstrate that the capacity of a MIMO system exceeds the capacity of a SISO system. It is established that, if the receiver has perfect knowledge of the channel and if the fade coefficients between any transmit-receive antenna pair are independent, the average channel capacity of a system with n_T transmit antennas and n_R receive antennas is approximately $min(n_T, n_R)$ times higher than that of a single antenna system, for the same total transmitting power and bandwidth. This exciting finding led to the proposal of space-time coding (STC) as a method for exploiting MIMO channels [10].

Space-time codes introduce temporal and spatial correlation into signals transmitted from different antennas, in order to provide diversity at the receiver, and coding gain over an uncoded system without impairing bandwidth efficiency. Today, space-time coded schemes can be found in several forms, such as, space-time trellis coded modulation (STTCM) [10], space-time block coding (STBC) [11], space-time turbo trellis coding (STTuTC) [12] and layered space-time architectures (LST) [13, 14]; all of which provide high data rates with a given transceiver complexity. This is stimulating considerable research interest and extensive investigation at present. In particular, the construction of space-time coding schemes where the aim is to find the best trade-off between three conflicting goals of minimising the decoder complexity, maximising the error performance, and maximising the information rate.

1.3 Space-Time Coding for MIMO Channels

Current transmission schemes over MIMO channels typically fall into two categories: data rate maximization or diversity maximization schemes, although there has been some effort toward unification recently. The first kind focuses on improving the average capacity behaviour by transmitting independent signals over each transmit antenna. However, in order to protect transmission against errors caused by channel fading and noise plus interference, the individual streams should be encoded jointly. This leads to a second kind of approach. As the level of redundancy is increased between the transmit antennas through joint encoding, the amount of independence between the signals decreases. In fact it is possible to code the signals so that the effective data rate is back to that of a single antenna system. Effectively, each transmit antenna then sees a differently encoded, fully redundant version of the same signal. In this case, the multiple antennas are only used as a source of spatial diversity and not to increase data rate.

STCs are aimed at realizing joint encoding across multiple transmit antennas. In these schemes, a number of code symbols equal to the number of transmit antennas are generated using a *space–time encoder* and transmitted simultaneously, one symbol from each antenna. Encoding is done such that by using the appropriate signal processing and decoding procedure at the receiver, the diversity gain and/or the coding gain is maximized. STC were first presented in [15] and was inspired by the delay diversity scheme of Wittneben [16,17]. However, the key development of the STC concept was originally revealed in [10] in the form of trellis codes, which required a multidimensional (vector) Viterbi algorithm at the receiver for decoding. These codes were shown to provide a diversity benefit equal to the number of transmit antennas in addition to a coding gain that depends on the complexity of the code (i.e., number of states in the trellis) without any loss in bandwidth efficiency.

Space–time block codes (STBCs) due to their simple construction and the fact that they can be decoded using simple linear processing at the receiver eventually lead to the popularity of STCs. Although STBC codes give the same diversity gain as the STTC for the same number of transmit antennas, they provide zero or minimal coding gain. Similar to STTC, space-time turbo trellis coding (STTuTC) combines the diversity advantage of space-time coding, the coding benefits of turbo codes and the bandwidth efficiency of coded modulation. Layered space-time codes come under the category of maximizing data rate due to the fact that there is no joint encoding across the transmit antennas.

1.4 STC – Rate, Diversity and Complexity Trade-offs

In this section the motivation behind the research work presented in this thesis is described. As briefly mentioned in Section 1.2, current research work in STC is geared at obtaining systems which provide the best rate/diversity/complexity trade-offs. These trade-offs are described in this section in the order of complexity, followed by rate and finally diversity.

STTC employ complex trellis encoding across multiple transmit antennas. When the number of antennas is fixed, the decoding complexity of STTC (measured by the number of trellis states at the decoder) increases exponentially as a function of the diversity level and transmission rate [10]. If STTC is considered complex due to its encoding/decoding methodology, STTuTC schemes make use of two STTC encoders; consequently, the receiver end employs two STTC decoders. Further, the decoding performance is dependent on the number of iterations which makes the entire scheme even more complex. In contrast the STBC scheme supports maximum likelihood (ML) detection based only on linear processing at the receiver. For LST, the complexity of ML decoding is high and considered prohibitive when many antennas or high-order modulations are used. However, due to the nature of LST codes, sub-optimal receivers such as zero forcing (ZF) and minimum mean squared error (MMSE) applied to multi-user detection (MUD) for code division multiple access (CDMA) systems (e.g. as in [18]) can be used. ZF and MMSE require simple linear processing at the receiver which are considered to be of low complexity. Hence in terms of complexity, STBC and LST stand out as the best suited techniques.

The rate r_s of a wireless link is defined as the number of independent information symbols transmitted per symbol interval. In the case of a SISO link, r_s =1. STTC, STTuTC and STBC are known to provide rates of r_s = 1. The first realization of STBC presented by Alamouti in [19] are known to be the only codes which provide a rate of r_s =1 for two transmit antennas. Codes for more than two transmit antennas are presented in [11] with a maximum rate of r_s = ¾. In contrast LST codes provide a rate equal to the number of transmit antenna $r_s = n_T$. Hence, in terms of rate, STBC and LST stand out again as the best suited techniques.

STTC and STTuTC provide both diversity and coding gains as compared to STBC which is capable of offering a diversity gain only. However, optimal outer multilevel codes (MLC) presented in [20] can be concatenated to STBC to provide a coding gain as shown in [21]. Similarly turbo trellis coded modulation is applied to STBC in [22]. Both STTC and STBC are capable of providing a full diversity order of $n_R \times n_T$. LST in comparison provides a diversity order of $n_R - n_T + 1$ with the use of a ZF receiver and n_R at best using a MMSE receiver [23]. Thus it is fair to say that LST architectures are outperformed in terms of diversity by other STC schemes but can provide a coding gain by employing an outer code as for STBC.

Based on the arguments presented above, LST and STBC can be picked out as the two best STC schemes for the maximization of data rate and the maximization of diversity gain, respectively. Also based on the fact that both schemes require simple linear processing at the receiver which is preferred in order to keep the transmitter and receiver complexity low [17]. However, both complement each other with the former providing a high multiplexing gain with little or no diversity gain and the latter providing a high diversity gain with little or no multiplexing gain. Combining both STBC and LST can yield both gains simultaneously. In order to achieve this tradeoff between the two gains, [7] presented a combined array processing and space-time coding architecture, in which the transmit stream is partitioned into different groups and in each group STC is applied. In [24] this architecture is further developed with the Alamouti's scheme adopted as the component STC code used for each group and the transmit power was optimized to minimize the average frame error rate (FER).

1.5 Aims and Outline of the Thesis

In this thesis the principles of STBC and LST are studied in order to provide a system architecture which achieves an optimum trade-off between rate and diversity.

1.5.1 Contributions Presented in the Thesis

The contributions made in this thesis are listed as follows:

- A novel approach called KIL-BLAST is presented which suppresses the effects of error propagation (EP) in LST architectures in an attempt to improve the diversity performance of ZF receivers in the presence of channel estimation errors. The scheme developed is adapted to square MIMO systems with $n_T = n_R$ and results are presented for 3×3 and 4×4 antenna configurations. The approach was published in a journal paper [40].

- A novel approach is proposed and developed which combines both STBC and LST in order to provide a trade-off between rate and diversity. It is called a Layered Space Time Block Coding Scheme (LSTBC). A significant improvement in diversity for LST is demonstrated with a rate r_s of $n_T - 1$ compared to the schemes in [7] and [24] which provide a rate of $n_T/2$. Results are produced for $n_T = n_R = 4$, 5 and 6 antenna configurations with a single STBC layer in each. This approach was published in two conference papers [26] and [27].

- The performance of LSTBC is studied in terms of multi-user detection. Improvements are made to the receiver algorithm to improve the performance of LSTBC with increasing number of transmit and receive antennas, replicating a cellular CDMA environment. It was shown that with an increasing number of users transmitting at the same power, the overall error rate performance matched that of LST.

1.5.2 Organization of the Thesis

The main concepts of STBC codes are presented in Chapter 2 with emphasis on the Alamouti scheme which is used as component codes for LSTBC. An analysis of the performance of STBC codes in the presence of an equal power interferer is also presented. This is key to the development of the LSTBC scheme.

Chapter 3 presents the concepts of LST codes. The MUD methods in CDMA developed for unmixing the channels using ZF and MMSE are presented. A detailed analysis of the effects of error propagation (EP) are presented. It is shown that the first layer to be decoded dominates the performance of the whole architecture.

In Chapter 4, the first contribution of this thesis is presented. A novel scheme which guarantees the decision statistics of the first layer to be perfect, thereby significantly reducing the effects of EP. This scheme is known as known-interference-layer BLAST (KIL-BLAST). Further study on the trade-off provided by the adoption of this scheme is presented Numerical results are used to justify the value of this scheme.

Building on the success of KIL-BLAST, a more robust transmit-receive architecture is presented in Chapter 5 which integrates STBC and LST. A detailed study of how interference effects on STBC demonstrated in Chapter 2 are mitigated is presented followed by an iterative method of decoding which makes use of the highest diversity order decision statistics in the first iteration to yield higher performance. Further study on the trade-off provided by the adoption of this scheme is presented. Comparison results are produced to justify the use of this scheme. Finally, the performance of LSTBC is evaluated in a CDMA multi-user environment and improvements are made to the algorithm. Results are presented to justify these improvements.

Finally Chapter 6 concludes the thesis and some recommendations for possible future work are presented.

CHAPTER 2 SPACE-TIME BLOCK CODING

Space-time block codes address the issue of high complexity decoding solutions for MIMO wireless channels. The Alamouti code [19] was the first space-time block code to provide full transmit diversity. In this chapter, the Alamouti space-time coding scheme for two transmit antennas is presented. This is followed by a generalization of the space-time block coding scheme for more than two transmit and receive antennas. The effect of interference on space-time block codes is key to the work presented later in this thesis and is studied in this chapter.

2.1 Alamouti Scheme

The Alamouti scheme is historically the first space-time block code to provide full transmit diversity for systems with two transmit antennas. It is worthwhile to mention that delay diversity schemes [24] can also achieve full diversity, but they introduce interference between symbols and complex detectors are required at the receiver.

2.1.1 Encoding of Alamouti Space-Time Codes

A block diagram of the encoder for two transmit antennas is shown in figure 2.1. We assume that M-ary modulation scheme is used. In the Alamouti space-time encoder, each group of m information bits is first modulated, where $m = \log_2 M$. Then, the encoder takes a block of two modulated symbols x_1 and x_2 in each encoding operation and maps them to the transmit antennas according the code matrix given by

$$\mathbf{G}_2 = \begin{bmatrix} x_1 & -x_2^* \\ x_2 & x_1^* \end{bmatrix} \qquad \textbf{Eq 2-1}$$

The encoder outputs are transmitted over two consecutive symbol periods from the two transmit antennas. In the first symbol period, signals x_1 and x_2 are transmitted from antenna one and antenna two, respectively. In the second symbol period, signal $-x_2^*$ is transmitted from transmit antenna one and signal x_1^* from transmit antenna two, where x_1^* is the complex conjugate of x_1.

From here it can be seen that encoding is done in both the time and the space domains. The transmit sequences from antennas one and two by $\mathbf{x}^1$ and $\mathbf{x}^2$ are denoted as

$$\mathbf{x}^1 = [x_1, -x_2^*]$$

$$\mathbf{x}^2 = [x_2, x_1^*] \qquad \textbf{Eq 1-2}$$

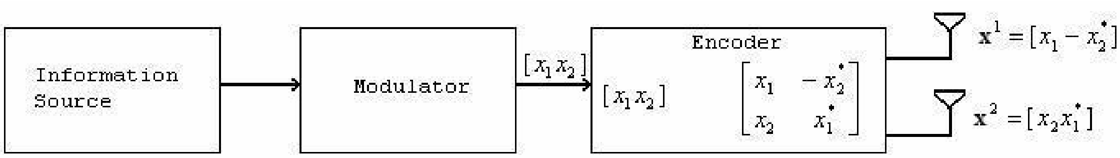

Figure 1-1 Alamouti Space-Time Encoder

The key feature of the Alamouti scheme is that the transmit sequences from the two transmit antennas are orthogonal over a frame interval, since the inner product of the sequences $\mathbf{x}^1$ and $\mathbf{x}^2$ is zero, i.e.

$$\mathbf{x}^1 \cdot \mathbf{x}^2 = x_1 x_2^* - x_2^* x_1 = 0 \qquad \textbf{Eq 1-3}$$

The code matrix has the following property

$$\mathbf{X} \cdot \mathbf{X}^H = \begin{bmatrix} |x_1|^2 + |x_2|^2 & 0 \\ 0 & |x_1|^2 + |x_2|^2 \end{bmatrix}$$

$$= \left(|x_1|^2 + |x_2|^2\right)\mathbf{I}_2 \qquad \textbf{Eq 1-4}$$

where $\mathbf{I}_2$ is a 2×2 identity matrix.

Let us assume that one receive antenna is used at the receiver. The block diagram of the receiver for the Alamouti scheme is shown in figure 2.2. The fading channel coefficients from the first and second transmit antennas to the receive antenna at time t are denoted by $h_1(t)$ and $h_2(t)$, respectively. Assuming that the fading coefficients are constant across two consecutive symbol transmission periods, they can be expressed as follows

$$h_1(t) = h_1(t+T) = h_1 = |h_1| e^{jq_1} \qquad \textbf{Eq 1-5}$$

and

$$h_2(t) = h_2(t+T) = h_2 = |h_2| e^{jq_2} \quad \textbf{Eq 1-6}$$

where $|h_i|$ and q_i, i = 0, 1, are the amplitude gain and phase shift for the path from transmit antenna I to the receive antenna, and T is the symbol duration. At the receive antenna, the received signals over two consecutive symbol periods, denoted by r_1 and r_2 for time t and $t+T$, respectively, can be expressed as

$$r_1 = h_1 x_1 + h_2 x_2 + n_1 \quad \textbf{Eq 1-7}$$

$$r_2 = -h_1 x_2^* + h_2 x_1^* + n_2 \quad \textbf{Eq 1-8}$$

where n_1 and n_2 are independent complex variables with zero mean and power specral density $No/2$ per dimension, representing additive white Gaussian noise samples at time t and $t+T$, respectively.

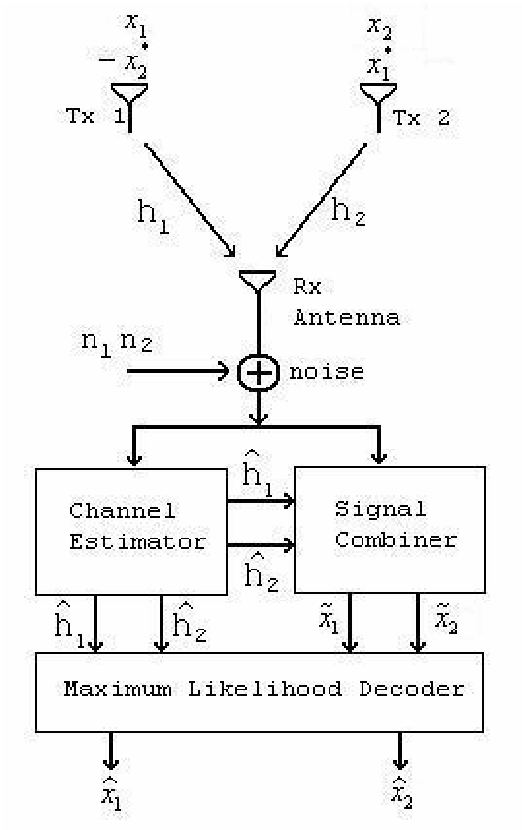

Figure 1-2 Alamouti Receiver

2.1.2 Decoding of Alamouti Space-Time Codes

The channel coefficients h_1 and h_2 are assumed to be perfectly estimated at the receiver. It is also assumed that all the signals in the modulation constellation are equiprobable and a maximum likelihood decoder (MLD) is employed at the receiver. The MLD chooses a pair of signals $(\hat{x}_1, \hat{x}_2)$ from the signal modulation constellation to minimize the distance metric

$$d^2\left(r_{1_1} h_1\hat{x}_1 + h_2\hat{x}_2\right) + d^2\left(r_{2_1} - h_1\hat{x}_2^* + h_2\hat{x}_1^*\right)$$

$$= \left| r_1 - h_1 \hat{x}_1 - h_2 \hat{x}_2 \right|^2 + \left| r_1 + h_1 \hat{x}_2^* - h_2 \hat{x}_1^* \right|^2 \qquad \textbf{Eq 1-9}$$

over all possible values of $\hat{x}_1$ and $\hat{x}_2$. Substituting Eq 2.7 and Eq 2.8 into Eq 2.9, the maximum likelihood decoding can be represented as

$$(\hat{x}_1, \hat{x}_2) = \arg \min_{(\hat{x}_1, \hat{x}_2) \in C} \left(|h_1|^2 + |h_2|^2 - 1 \right)\left(|\hat{x}_1|^2 + |\hat{x}_2|^2 \right) + d^2(\tilde{x}_1, \hat{x}_1) + d^2(\tilde{x}_2, \hat{x}_2) \qquad \textbf{Eq 1-10}$$

where C is the set of all possible modulated symbol pairs $(\hat{x}_1, \hat{x}_2)$, $\tilde{x}_1$ and $\tilde{x}_2$ are two decision statistics constructed by combining the received signals with channel state information (CSI). The decision statistics are given by

$$\tilde{x}_1 = h_1^* r_1 + h_2 r_2^*$$

$$\tilde{x}_2 = h_2^* r_1 - h_1 r_2^* \qquad \textbf{Eq 1-11}$$

Substituting r_1 and r_2 from Eq 2.7 and Eq 2.8 respectively, into Eq 2.11, the decision statistics can be written as,

$$\tilde{x}_1 = \left(|h_1|^2 + |h_2|^2 \right) x_1 + h_1^* n_1 + h_2 n_2^*$$

$$\tilde{x}_2 = \left(|h_1|^2 + |h_2|^2 \right) x_2 - h_1 n_2^* + h_2^* n_1 \qquad \textbf{Eq 1-12}$$

For a given channel realization h_1 and h_2, the decision statistics $\tilde{x}_1, i = 1,2,$ is only a function of $x_i, i = 1,2$. Thus, the maximum likelihood decoding rule (Eq 2.10) can be separated into two independent decoding rules for x_1 and x_2, given by

$$\hat{x}_1 = \arg \min_{\hat{x}_1 \in S} \left(|h_1|^2 + |h_2|^2 - 1 \right) |x_1|^2 + d^2(\tilde{x}_1, \tilde{x}_1) \qquad \textbf{Eq 1-13}$$

and

$$\hat{x}_2 = \arg\min_{\hat{x}_2 \in S} \left(|h_1|^2 + |h_2|^2 - 1\right)|x_2|^2 + d^2(\tilde{x}_2, \tilde{x}_2) \qquad \textbf{Eq 1-14}$$

respectively. For *M*-PSK signal constellations, $\left(|h_1|^2 + |h_2|^2 - 1\right)|x_i|^2$, i = 1, 2, are constant for all signal points, given the channel fading coefficients. Therefore, the decision rules in Eq 2.13 and Eq 2.14 can be further simplified to

$$\hat{x}_1 = \arg\min_{\hat{x}_1 \in S} d^2(\tilde{x}_1, \tilde{x}_1)$$

$$\hat{x}_2 = \arg\min_{\hat{x}_2 \in S} d^2(\tilde{x}_2, \tilde{x}_2) \qquad \textbf{Eq 1-15}$$

2.2 Performance Results of STBC

In this section, the improvement in symbol error rate (SER) performance of a wireless link from one transmit antenna to more than one transmit antenna with the use of space-time block coding is demonstrated. Simulation results for an uncoded SISO link and MISO links using two and three transmit antennas are shown in Figure 2-3. The two transmit and one receive antenna configuration uses the $\mathbf{G_2}$ transmit matrix given in Eq 2-1. For the three transmit and one receive antenna configuration, the transmit matrix $\mathbf{G_3}$ is given as

$$\mathbf{G}_3 = \begin{bmatrix} x_1 & -x_2^* & \dfrac{x_3^*}{\sqrt{2}} & \dfrac{x_3^*}{\sqrt{2}} \\ x_2 & x_1^* & \dfrac{x_3^*}{\sqrt{2}} & \dfrac{-x_3^*}{\sqrt{2}} \\ \dfrac{x_3}{\sqrt{2}} & \dfrac{x_3}{\sqrt{2}} & \dfrac{\left(-x_1 - x_1^* + x_2 - x_2^*\right)}{2} & \dfrac{\left(x_2 + x_2^* + x_1 - x_1^*\right)}{2} \end{bmatrix} \qquad \textbf{Eq 1-16}$$

From Eq2-16, it is clear that three independent symbols x_1, x_2 and x_3 are transmitted over four time periods. This implies that **G3** is a rate $3/4$ code. Both the uncoded and two transmit antenna configuration use a 8-PSK (3 bits per symbol) modulation scheme. The three transmit antenna configuration uses 16QAM (4 bits per symbol). Hence, the transmission rate for all three schemes is the fixed at 3b/s/Hz.

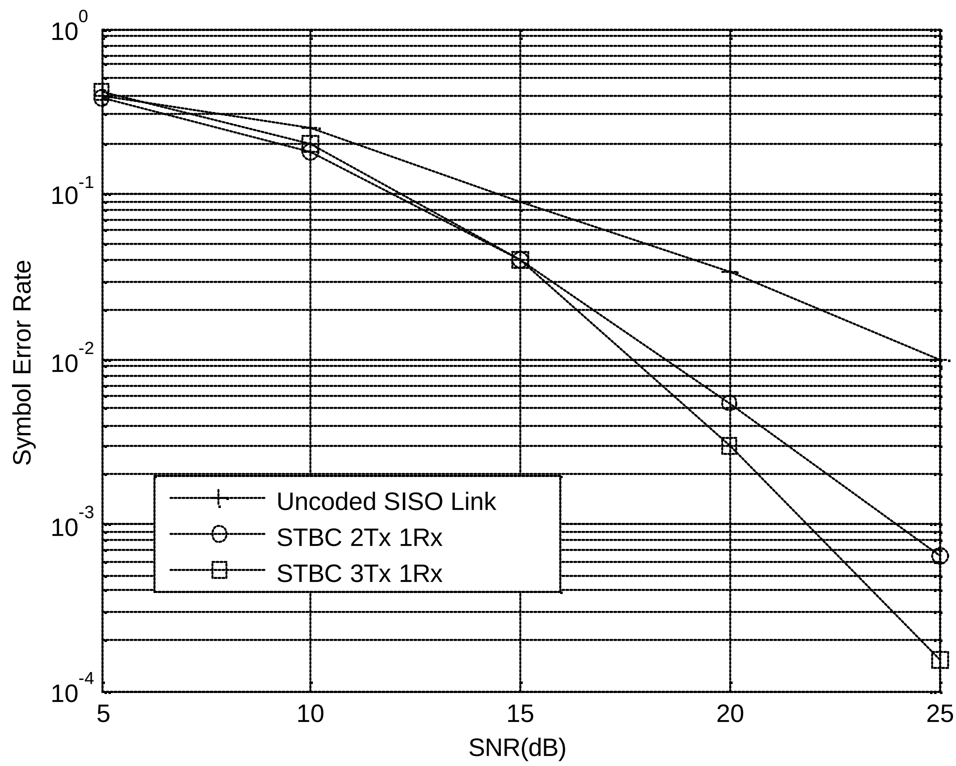

Figure 1-3 SER Performance of STBC with 1 Receive Antenna

From Figure 2-3, it can be observed that at a symbol error rate of 10^{-2} the STBC configuration using 2 transmit antennas is better than the uncoded configuration by about

7dB. The improvement in performance of the 3 transmit antenna configuration is approximately 1dB. However, the improvement in performance is seen to increase more rapidly with increasing SNR.

2.3 Interference Analysis of STBC

The orthogonality of STBC codes is susceptible to interference effects caused by independent signals. From Eq 2-4, a property of Alamouti space time codes is that the inner product of the transmit sequences is zero. To simply demonstrate the effects of interference, a signal x_3 can be considered as an equal power interferer transmitted from a third antenna over two consecutive time periods as shown in Figure 2-5.

The transmit sequences $\mathbf{x}^1$ and $\mathbf{x}^2$ are now given as

$$\mathbf{x}^1 = [x_1, -x_2^*, x_3]$$

and

$$\mathbf{x}^2 = [x_2, x_1^*, x_3] \qquad \textbf{Eq 1-17}$$

The dot product of the two sequences is given as

$$\mathbf{x}^1 \cdot \mathbf{x}^2 \neq 0 \qquad \textbf{Eq 1-18}$$

Hence, due to the lack of orthogonality of the transmit vectors, simple linear decoding as discussed in Section 2.1.2 cannot be used to separate the signals x_1 and x_2. From Figure 2-4, the performance of the STBC code in the presence of the interferer can be observed to be severely degraded. In order to mitigate this, some form of interference suppression technique would need to be adopted to reduce the impact of the x_3. One proposed technique is to increase the number of receive antennas over which the desired signals are received and decoded. The receiver signal processing for multiple receive antennas is considered next in Section 2.3.1. The significance of STBC performance in the presence of interference will be presented later in Chapter 5 of this thesis.

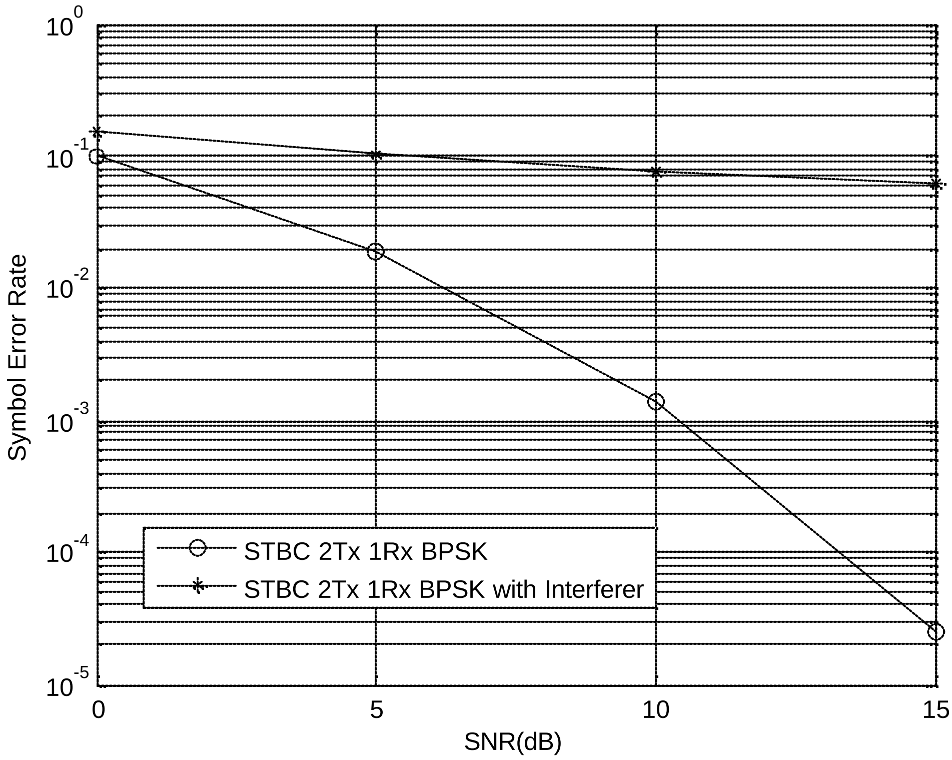

Figure 1-4 Loss in Orthogonality caused by interfering signal

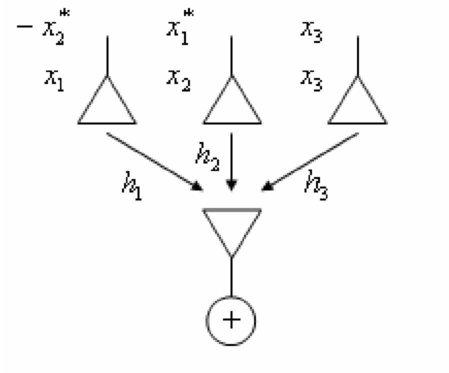

Figure 1-5 STBC with Interferer

2.3.1 Improved STBC Performance with Multiple Receive Antennas

The Alamouti scheme can be applied for a system with two transmit and n_R receive antennas. Encoding and transmission for this configuration is identical to the case of a single receive antenna as given in Section 2.1.2. The received signals at the j^{th} receive antennas at time t and $t + T$ are denoted as r_1^j and r_2^j respectively.

$$r_1^j = h_{j,1}x_1 + h_{j,2}x_2 + n_1^j$$

$$r_2^j = -h_{j,1}x_2^* + h_{j,2}x_1^* + n_2^j \qquad \textbf{Eq 1-19}$$

where $h_{j,i}$, $i =$ 1, 2, j= 1, 2,…, n_R, is the fading coefficient for the path from antenna i to receive antenna j, and n_1^j and n_2^j are the noise signals for the receive antenna j at time t and $t+T$, respectively. The receiver constructs two decision statistics based on the linear combination of the received signals. The decision statistics, denoted by $\tilde{x}_1$ and $\tilde{x}_2$, are given by

$$\tilde{x}_1 = \sum_{j=1}^{n_R} h_{j,1}^* r_1^j + h_{j,2}\left(r_2^j\right)^*$$

$$= \sum_{i=1}^{2}\sum_{j=1}^{n_R} \left|h_{j,i}\right|^2 x_1 + \sum_{j=1}^{n_R} h_{j,1}^* n_1^j + h_{j,2}\left(n_2^j\right)^*$$

$$\tilde{x}_2 = \sum_{j=1}^{n_R} h_{j,2}^* r_1^j - h_{j,1}\left(r_2^j\right)^*$$

$$= \sum_{i=1}^{2}\sum_{j=1}^{n_R} \left|h_{j,i}\right|^2 x_2 + \sum_{j=1}^{n_R} h_{j,2}^* n_1^j - h_{j,1}\left(n_2^j\right)^* \qquad \textbf{Eq 1-20}$$

The MLD rules for the two independent signals x_1 and x_2 are given by

$$\hat{x}_1 = \arg\min_{\hat{x}_1 \in S}\left[\left(\sum_{j=1}^{n_R}\left(\left|h_{j,1}\right|^2 + \left|h_{j,2}\right|^2\right) - 1\right)\left|\hat{x}_1\right|^2 + d^2\left(\tilde{x}_1, \hat{x}_1\right)\right] \qquad \textbf{Eq 1-21}$$

$$\hat{x}_2 = \arg\min_{\hat{x}_2 \in S}\left[\left(\sum_{j=1}^{n_R}\left(\left|h_{j,1}\right|^2 + \left|h_{j,2}\right|^2\right) - 1\right)\left|\hat{x}_2\right|^2 + d^2\left(\tilde{x}_2, \hat{x}_2\right)\right] \qquad \textbf{Eq 1-22}$$

For M-PSK modulation, all the signals in the constellation have equal energy. The MLD rules are equivalent to the case of a single receive antenna given in Eq 2-15.

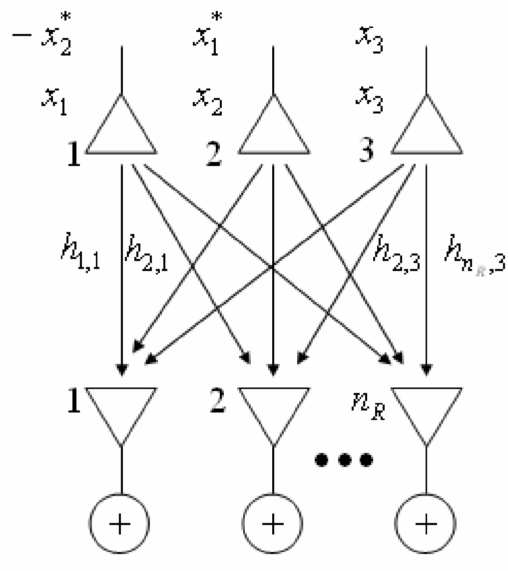

Figure 1-6 STBC with Multiple Receive Antennas

Figure 2-7 shows the improvement in performance of a 3b/s/Hz system when moving from a two transmit and one receive antenna configuration to a two transmit and two receive antenna configuration. Here the diversity order achieved by the former is two whilst the latter configuration achieves full diversity order of four. This can be observed from the difference in gradient of the two curves

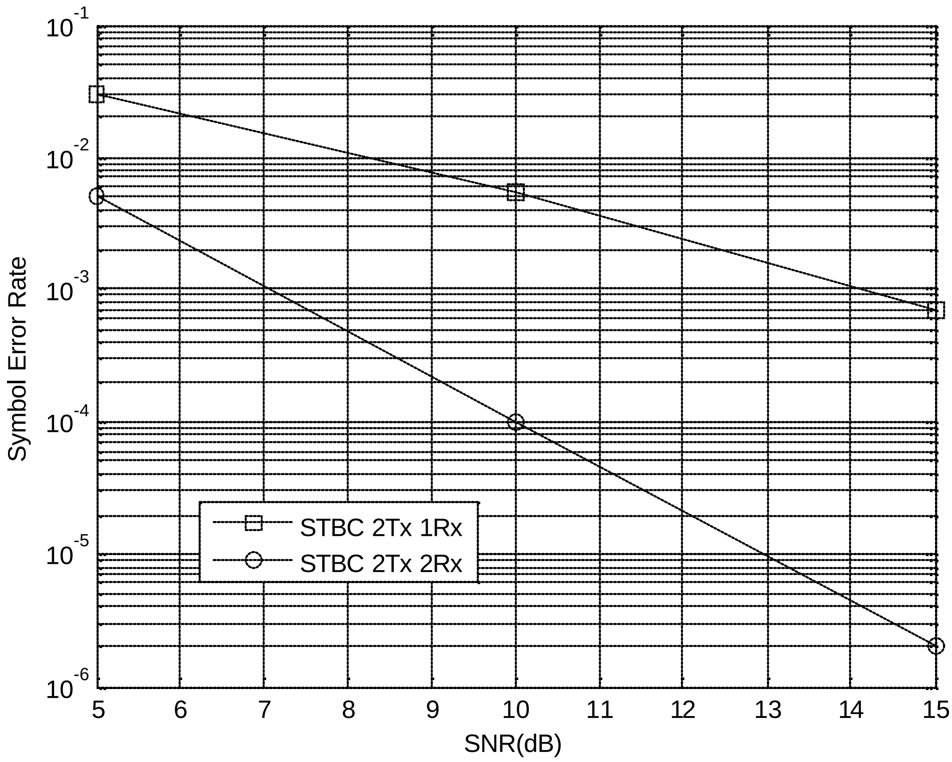

Figure 1-7. Improvement in diversity order with 2 receive antennas

2.3.2 Mitigating Effects of Interferer using Multiple Receive Antennas

The concept of diversity is based on multiple copies of a signal available at the receiver for decoding. This can be achieved using more than one transmit antenna with the same signal or one transmit antenna with multiple receive antennas receiving a copy of the same signal. Although it can be observed from Figure 2-7 that the use of additional receive antennas increases the performance, the same cannot be said in the presence of an interferer. In fact it is shown in Figure 2-8 that the performance tends to further degrade as the number of receive antennas is increased. This is as a consequence of the increasing signature of the interfering signal at the receiver.

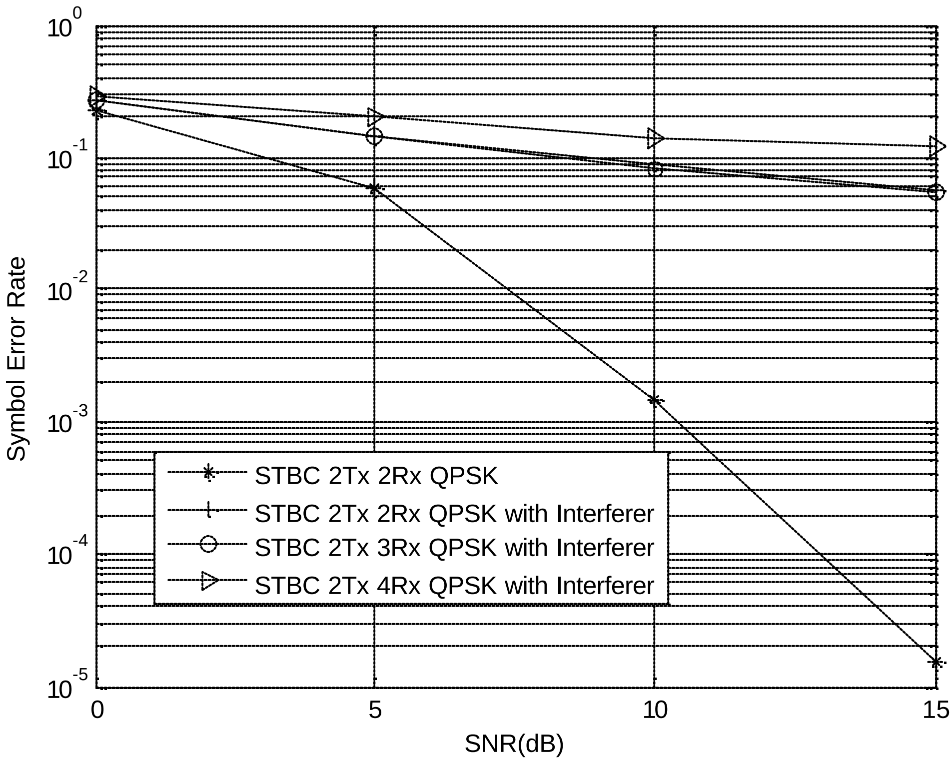

Figure 1-8 Further loss in performance caused by interfering signal

2.4 Application of STBC codes

Space-time block codes although able to provide high diversity orders also have high rate penalties. The Alamouti space-time code for two transmit and two receive antennas is capable of providing a rate of one. This implies that for every time period, one independent symbol is transmitted. For complex constellations, the maximum achievable rate for more than two transmit antennas is ¾. Hence, space-time block codes do not provide any multiplexing gain with respect to SISO channels, but they do possess a diversity order of ideally $n_T n_R$.

In the past, schemes have been proposed [7 and 26] , which propose a form of stacking (where two or more antennas are grouped vertically in a schematic representation) of antenna pairs which employ the Alamouti space-time block codes independently as shown in Figure 2-9. However, from the discussions in the previous sections, unless the signals from each pair of antennas are orthogonal, the effects of interference would render them ineffective.

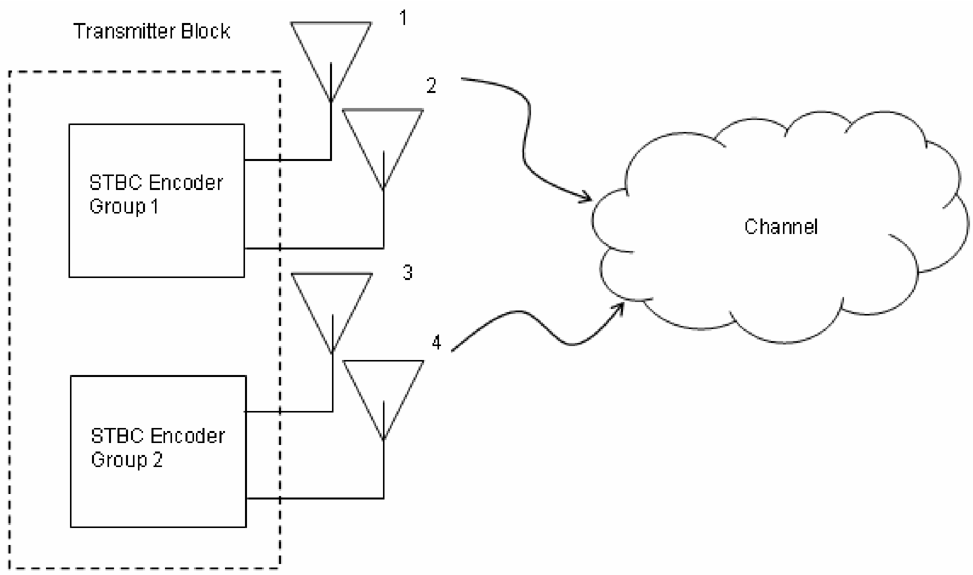

Figure 1-9 Combination of independent Alamouti Space-Time Codes

In Figure 2-9, both STBC groups are considered as synchronous co-channel terminals transmitting with the same power. At the receiver end, it was proposed in [7] that the minimum number of receive antennas is $n_R = 2$. MMSE interference cancellation with ML techniques that exploit the structure of space-time block codes to perfectly suppress the interference from the co-channel terminal while decoding the desired signal are employed. A two step MMSE interference cancellation technique which provides improved performance is also presented. However, the maximum rate provided by this scheme is two (one per group) which is low in contrast to a rate of four which can be provided by a layered space time (LST) coding system transmitting independent symbols from each antenna. Hence it is desirable to develop a technique which will employ the simple concept of stacking but provide a higher rate.

CHAPTER 3 LAYERED SPACE-TIME CODING

Space-time block codes make use of both the temporal and spatial dimensions in order to exploit the diversity in MIMO channels. However, they possess one key limitation in the context of multiplexing gain, which is that the spatial dimension is not exploited to the fullest. The spatial dimension is the key driver for MIMO channels and holds great promise for wireless system performance. LST proposed by Foschini in [13] was designed to improve the multiplexing gain by transmitting n_T independent data streams. Here, n_T information streams are transmitted simultaneously, in the same frequency band using n_T transmit antennas. The receiver uses $n_R = n_T$ antennas to separate and detect the n_T transmitted signals. The separation process involves a combination of interference suppression and interference cancellation. The separated signals are then decoded by using conventional decoding algorithms developed for 1-D-component codes (1-D refers to one dimension in space) such as MMSE and ZF, leading to much lower complexity compared to ML decoding used for STBC codes.

In this chapter, the principles of LST codes and transmitter architectures are presented. This is followed by the study of signal processing techniques used to decouple and detect the LST signals. ZF and MMSE interference suppression methods are considered. Further, as a contribution to this thesis, some key results are presented to demonstrate the effects of EP in LST architectures.

3.1 LST Transmitters

Design of LST architectures, depends on whether error control coding is used or not, and on the way the modulated symbols are assigned to transmit antennas. An uncoded LST structure, known as Vertical Bell Laboratories lAyered space-time (VBLAST) scheme [9], is illustrated in Figure 3-1.

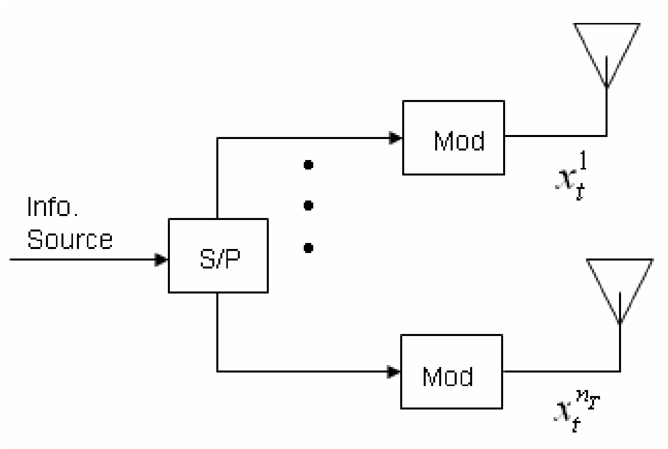

Figure 3-1 VBLAST Architecture

The input information sequence, denoted by **c**, is first demultiplexed into n_T sub-streams and each of them is subsequently modulated by an *M*-level modulation scheme and transmitted from a transmit antenna. The signal processing chain related to an individual sub-stream is referred to as a *layer*. The modulated symbols are arranged into a transmission matrix, denoted by **X,** which consists of n_T rows and *L* columns, where *L* is the transmission block length. The t^{th} column of the transmission matrix, denoted by $\mathbf{x}_t$, consists of the modulated symbols $x_t^1, x_t^2, ..., x_t^{n_T}$, where t = 1, 2, …, *L*. At a given time *t*, the transmitter sends the t^{th} column from the transmission matrix, one symbol from each antenna. That is, a transmission matrix entry x_t^i is transmitted from antenna *i* at time *t*.

For example, in a system with three transmit antennas, the transmission matrix **X** is given by

$$\mathbf{X} = \begin{bmatrix} x_1^1 & x_2^1 & x_3^1 & \ldots \\ x_1^2 & x_2^2 & x_3^2 & \ldots \\ x_1^3 & x_2^3 & x_3^3 & \ldots \end{bmatrix} \qquad \textbf{Eq 1-1}$$

The sequence $x_1^1, x_2^1, x_3^1,...$ is transmitted from antenna 1, the sequence $x_1^1, x_2^i, x_3^i,...$ is transmitted from antenna 2 and the sequence $x_1^3, x_2^3, x_3^3,...$ is transmitted from antenna 3.

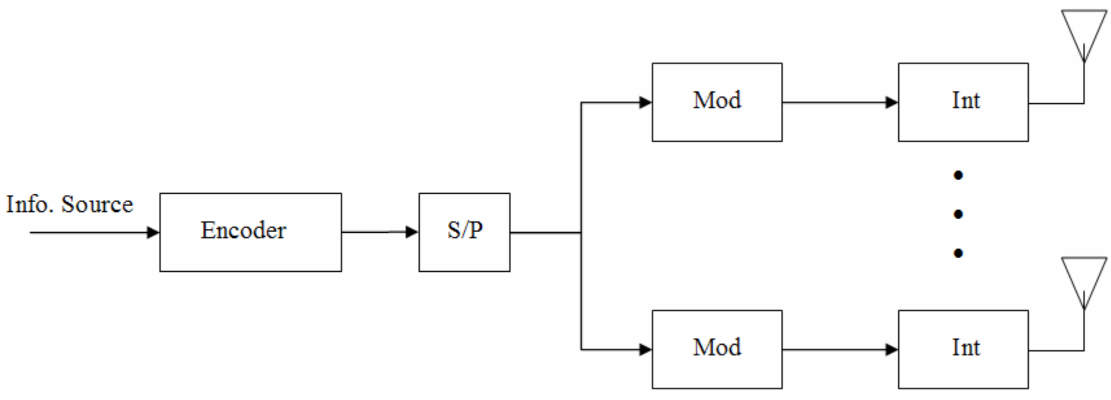

Figure 1-2 HLST Architecture

3.2 LST Receivers

The signals transmitted are assumed to propagate through a rich-scattering environment which causes the signals on different paths to interfere with each other upon reception at the receiver [51]. This interference is represented by the following matrix operation

$$\mathbf{r}_t = \mathbf{H}\mathbf{x}_t + \mathbf{n}_t \qquad \textbf{Eq 1-2}$$

where $\mathbf{r}_t$ is an n_R-component column matrix of the received signals across n_R receive antennas, $\mathbf{x}_t$ is the t^{th} column in the transmission matrix $\mathbf{X}$ and $\mathbf{n}_t$ is an n_R-component column matrix of the AWGN noise signals at the receive antennas, where the noise variance per receive dimension is denoted by s^2. It is also normally assumed that perfect channel state information (CSI) is available at the receiver.

3.2.1 Sub-Optimal Receivers

LST structures can be viewed as a synchronous code division multiple access (CDMA) system in which the number of transmit antennas is equal to the number of users. The interference between the antennas is equivalent to multiple access interference (MAI) in CDMA, while the complex fading coefficients correspond to the spreading sequences. Applying the same analogy at the receiver, multiuser receiver structures derived for CDMA can be directly applied to LST systems. Maximum likelihood (ML) detectors provide optimum receiver for an uncoded LST. It computes ML statistics as in the Viterbi algorithm. The complexity of this detection algorithm is exponential in the number of the transmit antennas.

For coded LST schemes, the optimum receiver performs joint detection and decoding on an overall trellis obtained by combining the trellises of the layered space-time coded and the channel code. The complexity of the receiver is an exponential function of the product of the number of the transmit antennas and the code memory order. For many systems, the exponential increase in implementation complexity may make the optimal receiver impractical even for a small number of transmit antennas. Thus, in this chapter, less complex receiver architectures are examined which are known to have good performance/complexity trade-offs.

The original VLST receiver [9] is based on a combination of interference suppression and cancellation. Conceptually, each transmitted sub-stream is considered in turn to be the desired symbol and the remaining substreams are treated as interferers. which are suppressed by a ZF approach [9]. This detection algorithm produces ZF based decision statistics for a desired sub-stream from the received signal vector **r**, which contains a *residual* interference from other transmitted sub-streams. Subsequently, a decision on the desired sub-stream is made from the decision statistics and its interference contribution is regenerated and subtracted out from the received vector **r**. Thus **r** now contains a lower level of interference and this will increase the probability of correct detection of other sub-streams.

To illustrate this operation, a three transmit three receive antenna configuration is shown in Figure 3-3. The first detected sub-stream is that received on antenna three. It is a composite signal made up of the sum of symbols 1, 2 and 3 scaled by their corresponding channel coefficients plus noise. This can be represented mathematically as

$$r_3 = h_{3,1}x_1 + h_{2,1}x_2 + h_{3,3}x_3 + n_3 \qquad \textbf{Eq 1-3}$$

where r_3 is the third component of the received vector $\mathbf{r}$, n_3 is the third component of the noise vector $\mathbf{n}$, x_i is the symbol transmitted from antenna i and $h_{j,i}$ is the channel coefficient between receive antenna j and transmit antenna i.

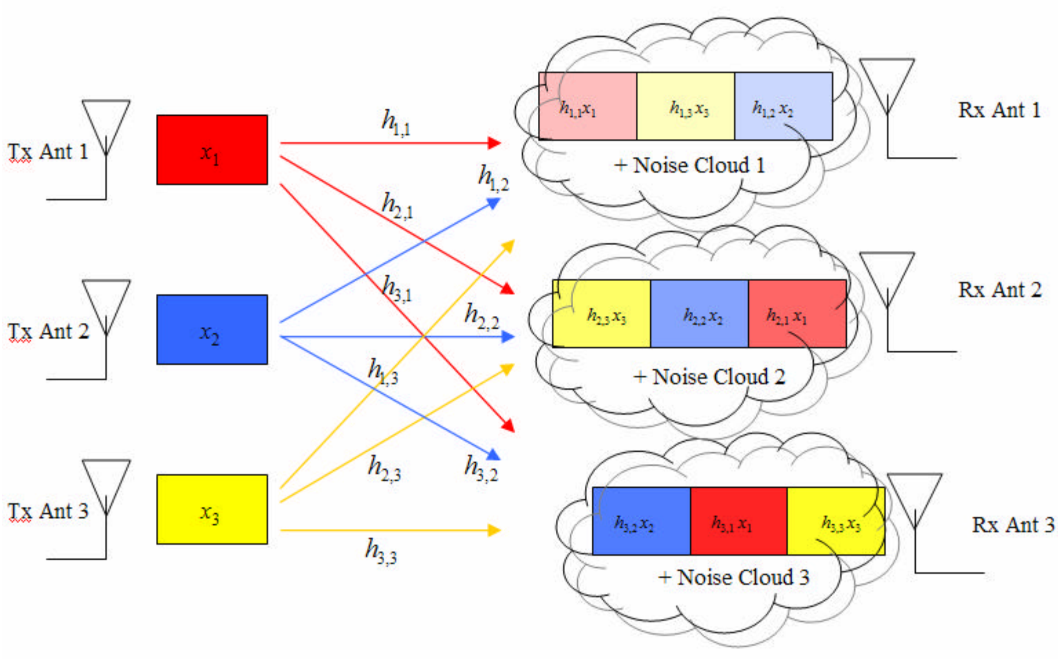

Figure 1-3 Example of 3 Transmit, 3 Receive V-BLAST Architecture

The algorithm used suppresses the interference from the signals transmitted by antennas one and two. Assuming that the interference is suppressed perfectly, the new signal at antenna three, denoted by y_3 can be represented as

$$y_3 = h_{3,3}x_3 + n_3 \qquad \textbf{Eq 1-4}$$

Assuming perfect CSI, we can compensate for the channel coefficient $h_{3,3}$ using the following transformation

$$z_3 = y_3 (h_{3,3})^H = h_{3,3} x_3 (h_{3,3})^H + n_3 (h_{3,3})^H$$

Eq 1-5

$$= x_3 + h_3$$

where $(\bullet)^H$ is represents the conjugate transpose operation, $h_3 = n_3 (h_{3,3})^H$ is the transformed noise component, maintaining the same i.i.d characteristics as n_3. The signal z_3 is forwarded on to a detector which provides an estimate of the signal transmitted by antenna three.

For the second sub-stream on receive antenna two, the interference from transmit antenna one is suppressed, while the interference from transmit antenna three ($h_{3,3} x_3$) is subtracted from the received vector **r**. Finally, this process is repeated for the first sub-stream, which, assuming that the previous signals were detected correctly, will be free from interference.

The ZF strategy is only possible if the number of receive antennas is at least as large as the number of transmit antennas. Another drawback of this approach is that achievable diversity depends on a particular layer. If the ZF strategy is used to remove interference and if n_R receive antennas are available, it is possible to remove

$$n_i = n_R - d_o$$

Eq 1-6

interferers with diversity order of d_o [9]. The diversity order can be expressed as

$$d_o = n_R - n_i$$

Eq 1-7

If the interference suppression starts at layer n_T, then at this layer ($n_T - 1$) interferers need to be suppressed. Assuming that $n_R = n_T$, the diversity order in this layer, according to Eq

3-7 is 1. In the 1st layer, there are no interferers to be suppressed, so the diversity order is $n_R = n_T$.

3.2.2 ZF Receiver Using QR Decomposition with Combined Suppression and Interference Cancellation

Any $n_R \times n_T$ matrix $\mathbf{H}$, where $n_R \geq n_T$, can be decomposed as

$$\mathbf{H} = \mathbf{U_R R} \qquad \text{Eq 1-8}$$

where $\mathbf{U_R}$ is a $n_R \times n_T$ unitary matrix and $\mathbf{R}$ is an $n_T \times n_T$ upper right triangular matrix, with entries $\left(R_{i,j}\right)_t = 0$, for $i > j, i, j = 1,2,...n_T$, represented as

$$\mathbf{R} = \begin{bmatrix} \left(R_{1,1}\right)_t & \left(R_{1,2}\right)_t & \cdots & \left(R_{1,n_T}\right)_t \\ 0 & \left(R_{2,2}\right)_t & \cdots & \left(R_{2,n_T}\right)_t \\ 0 & 0 & \cdots & \left(R_{3,n_T}\right)_t \\ \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \cdots & \left(R_{n_T,n_T}\right)_t \end{bmatrix} \qquad \text{Eq 1-9}$$

The decomposition of the matrix $\mathbf{H}$, as in Eq 3-9, is called *QR Factorization*. An n_T-component column matrix $\mathbf{y}$ is introduced, obtained by multiplying from the left of the receive vector $\mathbf{r}$, given by Eq 3-2 by $\mathbf{U}_\mathbf{R}^T$

$$\mathbf{y} = \mathbf{U}_\mathbf{R}^T \mathbf{r} \qquad \text{Eq 1-10}$$

or

$$\mathbf{y} = \mathbf{U}_\mathbf{R}^T \mathbf{H x} + \mathbf{U}_\mathbf{R}^T \mathbf{n} \qquad \text{Eq 1-11}$$

Substituting the QR decomposition of **H** from Eq 3-8 into Eq 3-11, we get for **y**

$$\mathbf{y} = \mathbf{R}\mathbf{x} + \mathbf{n}' \qquad \textbf{Eq 1-12}$$

where $\mathbf{n}' = \mathbf{U}_{\mathbf{R}}^{T}\mathbf{n}$ is an n_T-component column matrix of i.i.d AWGN noise signals. As **R** is upper-triangular, the i^{th} component in **y** depends only on the i^{th} and higher layer transmitted symbols at time *t*, as follows.

$$y_t^i = \left(R_{i,i}\right)_t x_t^i + n_t'^i + \sum_{j=i+1}^{n_T} \left(R_{i,j}\right)_t x_t^j \qquad \textbf{Eq 1-13}$$

Consider x_t^i is the current desired detected signal. Eq 3-13 shows that y_t^i contains a lower level of interference than in the received signal $\mathbf{r}_t$, as the interference from x_t^l, for $l < i$, is suppressed. The third term in Eq 3-13 represents contributions from other interferers, $x_t^{i+1}, x_t^{i+2}, \ldots, x_t^{n_T}$, which can be cancelled by using the available decisions $\hat{x}_t^{i+1}, \hat{x}_t^{i+2}, \ldots, \hat{x}_t^{n_T}$, assuming that they have been detected. The decision statistics on x_t^i, denoted by y_t^i, can be rewritten as

$$y_t^i = \sum_{j=1}^{n_T} \left(R_{i,j}\right)_t x_t^j + n_t'^i \qquad i = 1,2,\ldots,n_T \qquad \textbf{Eq 1-14}$$

The estimated value of the transmitted symbol x_t^i is given by

$$\hat{x}_t^i = q\left(\frac{y_t^i - \sum_{j=i+1}^{n_T} \left(R_{i,j}\right)_t \hat{x}_t^j}{\left(R_{i,i}\right)_t} \right) \qquad i = 1,2,\ldots,n_T \qquad \textbf{Eq 1-15}$$

where $q(x)$ denotes the hard decision on x.

For the example given in Figure 3-3, the decision statistics for various layers (or sub-streams) can be expressed as

$$y_t^1 = \left(R_{1,1}\right)_t x_t^1 + \left(R_{1,2}\right)_t x_t^2 + \left(R_{1,3}\right)_t x_t^3 + n_t'^1 \qquad \textbf{Eq 1-16}$$

$$y_t^2 = \left(R_{2,2}\right)_t x_t^2 + \left(R_{2,3}\right)_t x_t^3 + n_t'^2 \qquad \textbf{Eq 1-17}$$

$$y_t^3 = \left(R_{3,3}\right)_t x_t^3 + n_t'^3 \qquad \textbf{Eq 1-18}$$

The estimate on the transmitted symbol x_t^3, denoted by $\hat{x}_t^3$, can be obtained from Eq 3-18 as

$$\hat{x}_t^3 = q\left(\frac{y_t^3}{\left(R_{3,3}\right)_t} \right) \qquad \textbf{Eq 1-19}$$

The contribution of $\hat{x}_t^3$ is cancelled from Eq 3-17 and the estimate on x_t^2 is obtained as

$$\hat{x}_t^2 = q\left(\frac{y_t^2 - \left(R_{2,3}\right)_t \hat{x}_t^3}{\left(R_{2,2}\right)_t} \right) \qquad \textbf{Eq 1-20}$$

Finally, after cancelling out $\hat{x}_t^3$ and $\hat{x}_t^2$, we obtain $\hat{x}_t^1$

$$\hat{x}_t^1 = q\left(\frac{y_t^1 - (R_{1,3})_t \hat{x}_t^3 - (R_{1,2})_t \hat{x}_t^2}{(R_{1,1})_t}\right) \quad \textbf{Eq 1-21}$$

The described algorithm applies to VBLAST. In coded LST schemes, the soft decision statistics on x_t^i, given by the arguments in the $q(\bullet)$ expressions on the right-hand side in Eqs 3-19, 3-20 and 3-21, are passed to the channel decoder, which estimates $\hat{x}_t^i$.

In the above example the decision statistics $y_t^{n_T}$ is computed first, then $y_t^{n_T-1}$, and so on. The performance can be improved if the layer with the maximum SNR is detected first, followed by the one with the next largest SNR and so on.

3.2.3 Minimum Mean Square Error (MMSE) Receiver with Combined Suppression and Interference Cancellation

In the MMSE detection algorithm, the expected value of the mean square error between the transmitted vector $\mathbf{x}$ and a linear combination of the received vector $\mathbf{w}^H\mathbf{r}$ is minimized

$$\min\ E\{(\mathbf{x} - \mathbf{w}^H\mathbf{r})^2\} \quad \textbf{(6.23)} \quad \textbf{Eq 1-22}$$

where $\mathbf{w}$ is an $n_R \times n_T$ matrix of linear combination coefficients given by [8]

$$\mathbf{w}^H = [\mathbf{H}^H\mathbf{H} + s^2\mathbf{I}_{n_T}]^{-1}\mathbf{H}^H \quad \textbf{Eq 1-23}$$

s^2 is the noise variance and $\mathbf{I}_{n_T}$ is an $n_T \times n_T$ identity matrix. The decision statistics for the symbol sent from antenna i at time t is obtained as

$$y_t^i = \mathbf{w}_i^H \mathbf{r} \qquad \textbf{Eq 1-24}$$

where $\mathbf{w}_i^H$ is the i^{th} row of $\mathbf{w}^H$ consisting of n_R components. The estimate of the symbol sent by antenna i, denoted by $\hat{x}_t^i$, is obtained by making a hard decision on y_t^i

$$\hat{x}_t^i = q\left(y_t^i\right) \qquad \textbf{Eq 1-25}$$

In an algorithm with interference suppression only, the detector calculates the hard decision estimates by using Eq 3-24 and Eq 3-25 for all transmit antennas.

In a combined interference suppression and interference cancellation process, the receiver starts from antenna n_T and computes its signal estimate by using Eq 3-24 and Eq 3-25. The received signal $\mathbf{r}$ at this level is denoted by $\mathbf{r}^{n_T}$. For calculation of the next antenna signal (n_T -1), the interference contribution of the hard estimate $\hat{x}_t^{n_T}$ is subtracted from the received signal $\mathbf{r}^{n_T}$ and this modified received signal is denoted by $\mathbf{r}^{n_T-1}$ is used in computing the decision statistics for antenna (n_T – 1) in Eq 3-24 and its hard estimate from Eq 3-25. In the next level, corresponding to antenna (n_T – 2), the interference from n_T – 1 is subtracted from the received signal $\mathbf{r}^{n_T-1}$ and this signal is used to calculate the decision statistics in Eq 3-24 for antenna (n_T – 2). This process continues for all other levels up to the first antenna.

After detection of level i, the hard estimate $\hat{x}_t^i$ is subtracted from the received signal to remove interference contribution, giving the received signal for level $i-1$

$$\mathbf{r}^{i-1} = \mathbf{r}^{i} - \hat{x}_t^i \mathbf{h}_i \qquad \text{Eq 1-26}$$

where $\mathbf{h}_i$ is the i^{th} column in the channel matrix $\mathbf{H}$, corresponding to the path attenuations from antenna i. The operation $\hat{x}_t^i \mathbf{h}_i$ in Eq 3-26 replicates the interference contribution caused by $\hat{x}_t^i$ in the received vector. $\mathbf{r}^{i-1}$ is the received vector free from interference coming from $\hat{x}_t^{n_T}, \hat{x}_t^{n_T-1}, \ldots, \hat{x}_t^i$. For estimation of the next antenna signal $\hat{x}_t^{i-1}$, is the signal $\mathbf{r}^{i-1}$ is used in Eq 3-24 instead of $\mathbf{r}$. Finally, a deflated version of the channel matrix is calculated, denoted by $\mathbf{H}_d^{i-1}$, by deleting column i from $\mathbf{H}_d^i$. The deflated matrix $\mathbf{H}_d^{i-1}$ at the $(n_T - i + 1)^{th}$ cancellation step is given by

$$\mathbf{H}_d^{i-1} = \begin{bmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,i-1} \\ h_{2,1} & h_{2,2} & \cdots & h_{1,i-1} \\ \vdots & \vdots & \vdots & \vdots \\ h_{n_R,1} & h_{n_R,2} & \cdots & h_{n_R,i-1} \end{bmatrix} \qquad \text{Eq 1-27}$$

This deflation is needed as the interference associated with the current symbol has been removed. This deflated matrix $\mathbf{H}_d^{i-1}$ is used in Eq 3-23 for computing the MMSE coefficients and the signal estimate from antenna i – 1. Once the symbols from each antenna have been estimated, the receiver repeats the process on the vector $\mathbf{r}_{t+1}$ received at time (t + 1). This algorithm is summarised in Table 3-1.

Set i = n_T
and $\mathbf{r}^{n_T}$ = $\mathbf{r}$
while i ≥ 1
{

$$\mathbf{w}^H = \left[\mathbf{H}^H\mathbf{H} + \sigma^2\mathbf{I}_{n_T}\right]^{-1}\mathbf{H}^H$$

$$y_t^i = \mathbf{w}_i^H\mathbf{r}$$

$$\hat{x}_t^i = q\left(y_t^i\right)$$

$$\mathbf{r}^{i-1} = \mathbf{r}^i - \hat{x}_t^i\mathbf{h}_i$$

Compute $\mathbf{H}_d^{i-1}$ by deleting column i from $\mathbf{H}_d^i$

$$\mathbf{H} = \mathbf{H}_d^{i-1}$$

i = i – 1
}

Table 1-1 MMSE algorithm for V-BLAST

The receiver can be implemented without the interference cancellation step Eq 3-26. This will reduce system performance but some computational cost can be saved. Using cancellation requires that MMSE coefficients be recalculated at each iteration, as **H** is deflated. With no cancellation, the MMSE coefficients are only computed once, as **H** remains unchanged. The most computationally intensive operation in the detection algorithm is the computation of the MMSE coefficients. A direct calculation of the MMSE coefficients based on Eq 3-23, has a complexity that is polynomial in the number of transmit antennas. However, on slow fading channels, it is possible to implement adaptive MMSE receivers with the complexity being linear in the number of transmit antennas.

The described algorithm is for uncoded LST systems. The same detector can be applied to coded systems. The receiver consists of the described MMSE interference suppressor/canceller followed by the decoder. The decision statistics, y_t^i, from Eq 3-24, are passed to the decoder which makes the decision on the symbol estimate $\hat{x}_t^i$.

3.3 Performance of LST Architectures

In this section, error rate performances of the receiver algorithms presented in Section 3.2.2 and 3.2.3 are studied through the use of simulation. The simulation results highlight key properties of LST coding which can in some cases be viewed as limitations. In the simulations, a total transmit block length of 100 symbols is used. In Figure 3-4, the performance of an uncoded $n_T = 4$, $n_R = 4$ MIMO system using the V-BLAST architecture in Figure 3-1 is presented, employing QPSK modulation on each layer. The results shown are for both the ZF and MMSE receiver approaches.

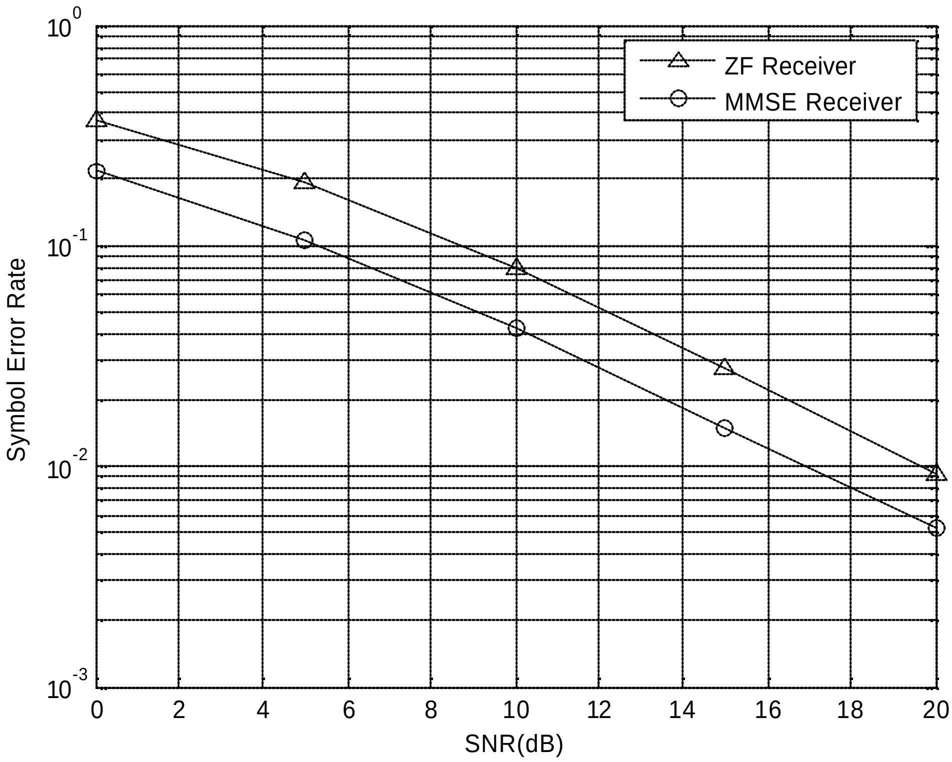

Figure 1-4 Comparison of LST ZF and MMSE receivers $n_T = n_R = 4$

As shown in Figure 3-4, both schemes achieve the same diversity order which is represented by a similar gradient for both symbol error rate (SER) curves with increasing SNR. However, at a SER of 1×10^{-2}, the MMSE scheme provides a 2dB improvement in performance over the ZF scheme. This improvement is due to the fact that in addition to nulling out the interferers, the MMSE scheme takes into consideration the noise on the channel represented by the s^2 term in Eq 3-23. A similar comparison in performance is observed when $n_T = n_R = 5$ as shown in Figure 3-5 below.

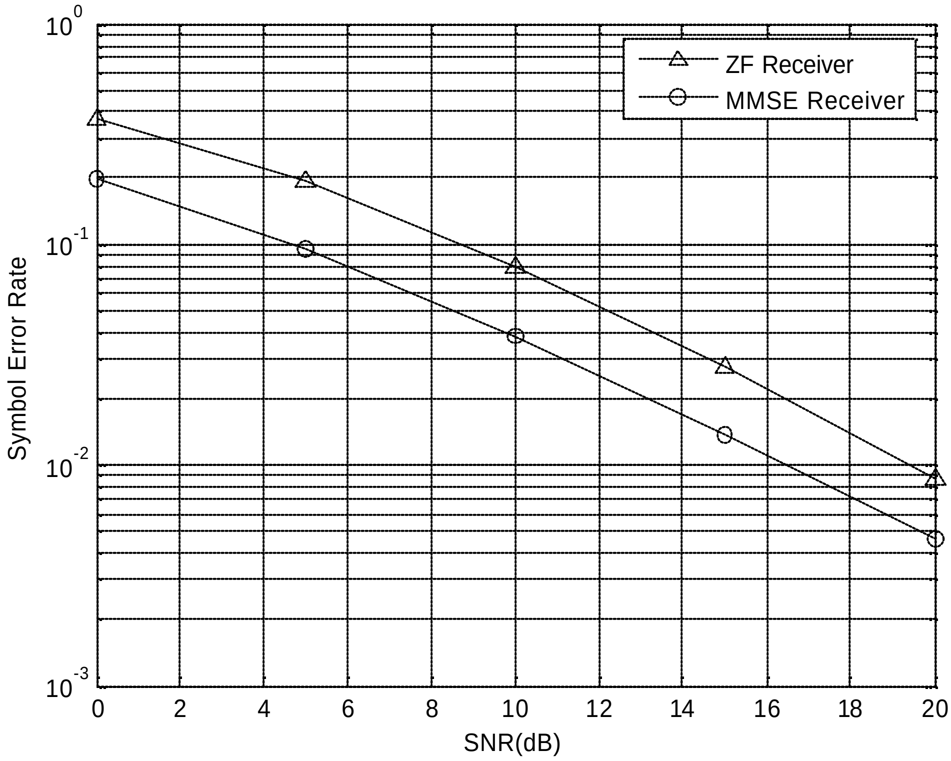

Figure 1-5 Comparison of LST ZF and MMSE receivers $n_T = n_R = 5$

The MMSE algorithm described in Section 3.2.3 can be regarded as a conventional algorithm which begins decoding from the n_T^{th} layer. A more robust MMSE algorithm [25] which uses ordered successive interference cancellation can be employed. Here, the layer with the highest SNR is decoded first followed by the layer with the next highest SNR and so on. From Figure 3-6, for a $n_T = n_R = 4$ architecture with QPSK modulation on each layer, an improvement of 1dB can be observed at a SER of 1×10^{-2} using ordered MMSE over the conventional MMSE receiver. The performance curves for all three schemes decay at the same rate with increasing SNR which indicates that the diversity order achieved is the same in all cases.

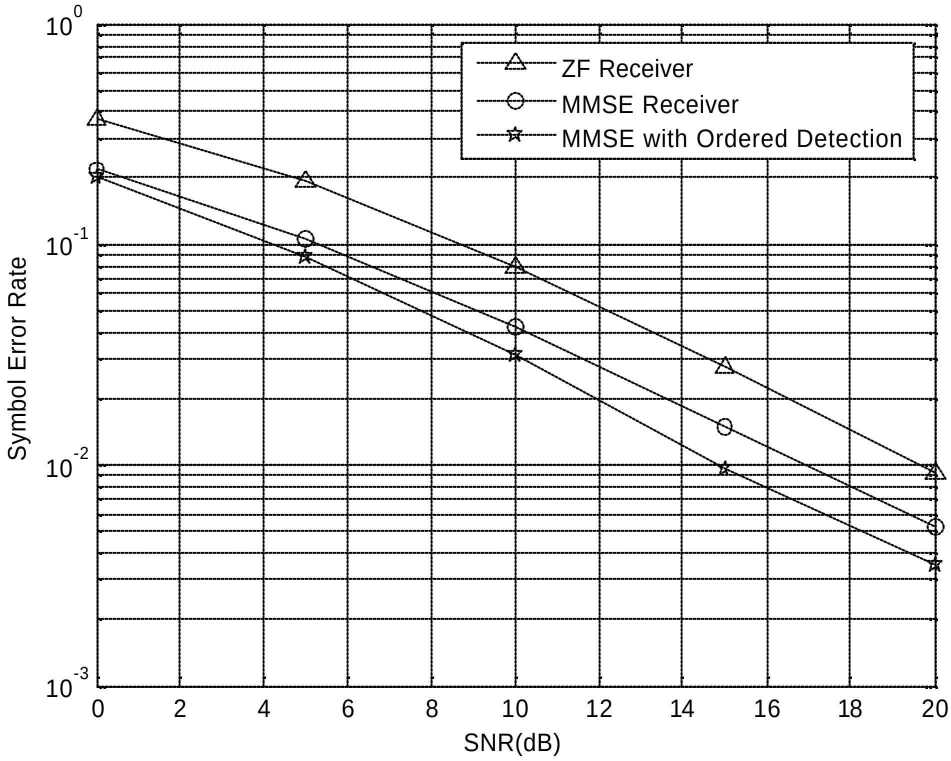

Figure 1-6 Comparison of ZF, MMSE and Ordered MMSE Receivers n_T=n_R=4

In Figure 3-7, the performance of the ZF and MMSE receivers is compared for higher modulation levels. Although the previous results indicate an approximate improvement of 2dB for increasing SNR using QPSK, with 8PSK, this improvement is not the same. Hence, as the constellation size of the transmitted symbols increases, the performance of the MMSE scheme approaches that of the ZF scheme.

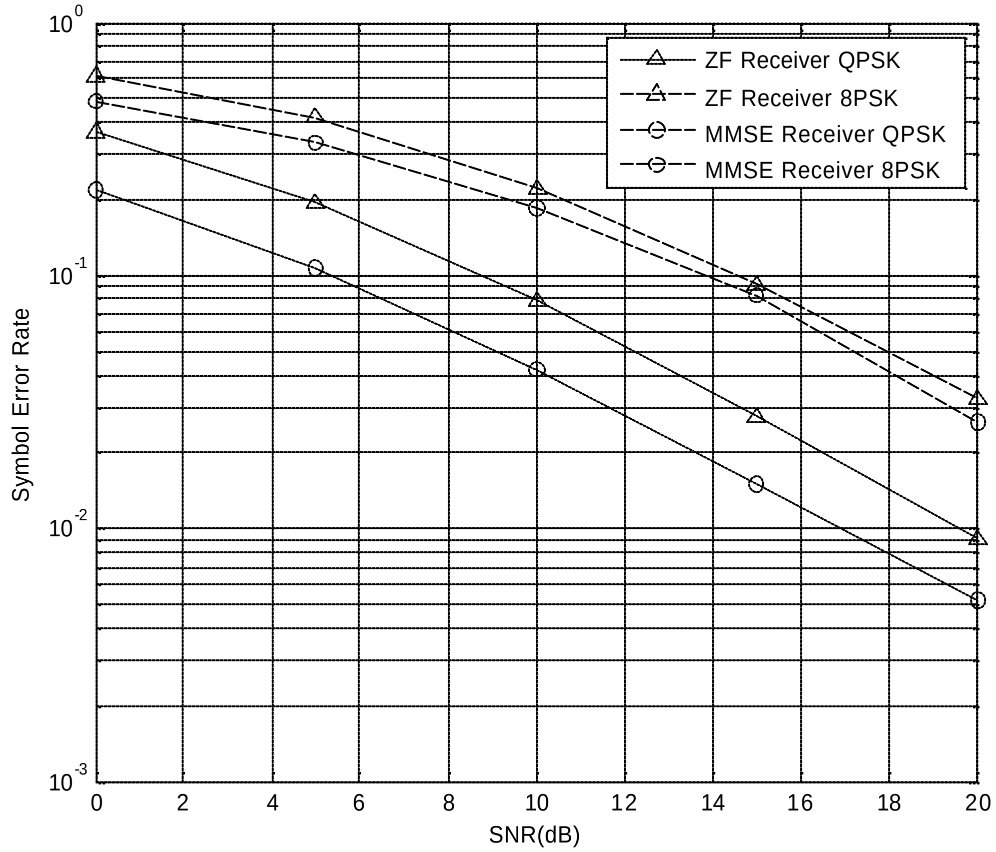

Figure 1-7 Performance of MMSE and ZF with higher constellations

Finally the performance of both the ZF and MMSE schemes is evaluated with an increasing number of transmit and receive antennas. Figure 3-8 shows the performance of the ZF scheme with $n_T = n_R = 4$, 5 and 6 with QPSK modulation applied to each layer. It can be observed that increasing the number of transmit and receive antennas does not lead to an improvement in performance, as the number of interferers at the receiver increases proportionately.

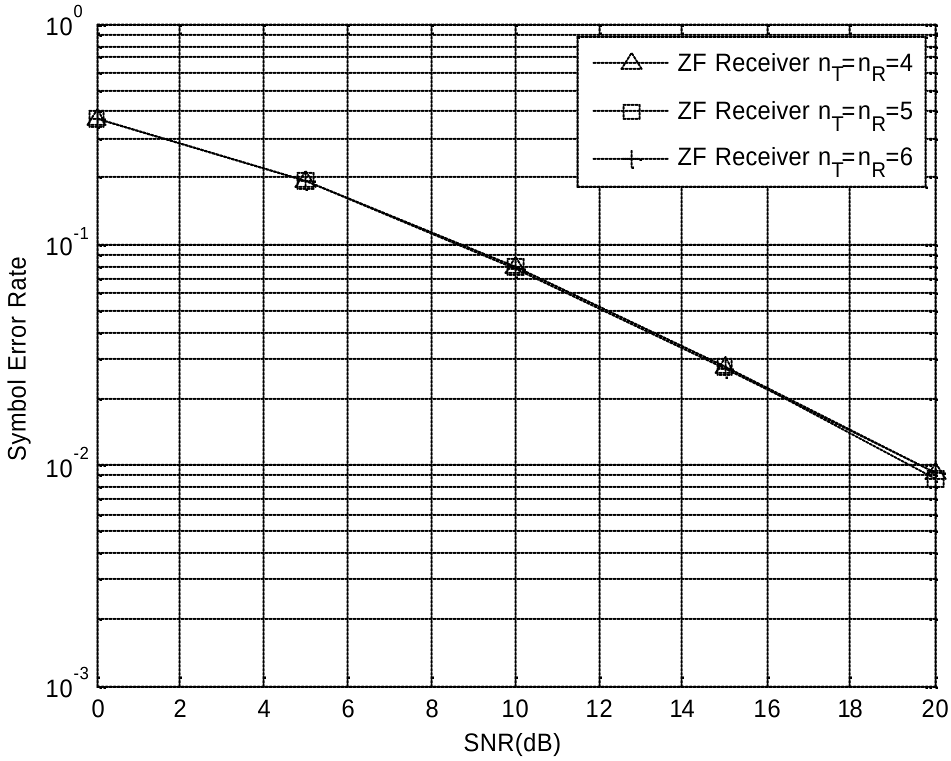

Figure 1-8 LST performance with increasing n_T and n_R (ZF Receiver)

From Figure 3-9, it can be observed that the MMSE receiver provides a slight enhancement to the SER performance as the number of transmit and receive antennas increase proportionately. This is owed mainly to the fact that MMSE does not enhance the noise at the receiver compared to ZF as shown in Eq 3-8.

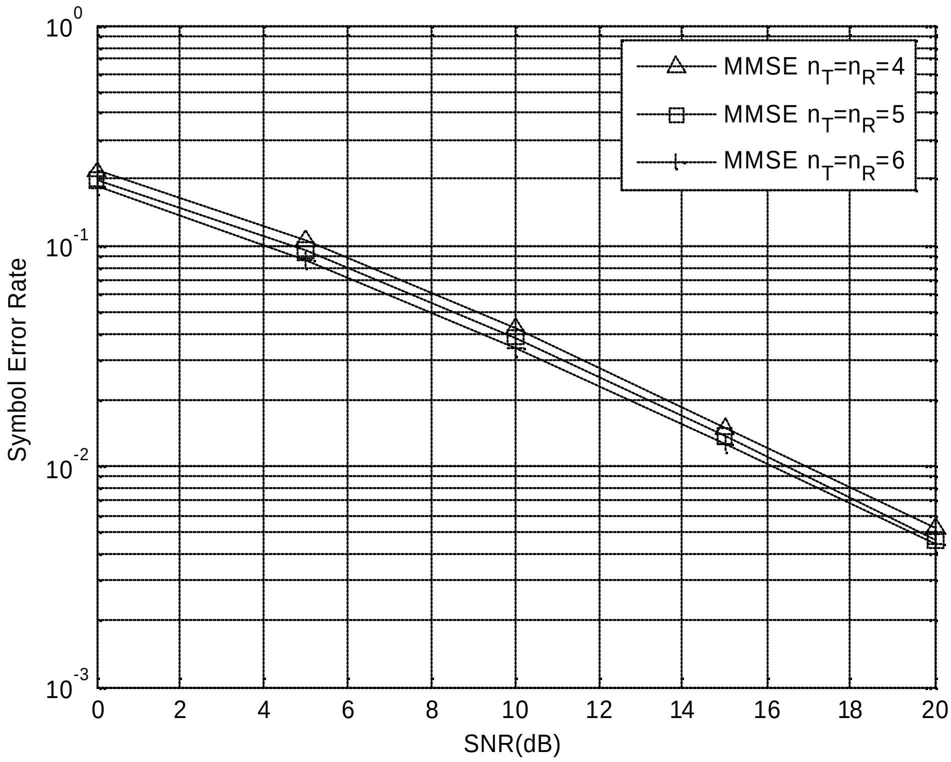

Figure 1-9 LST performance with increasing n_T and n_R (MMSE Receiver)

CHAPTER 4 KNOWN INTERFERENCE LAYER BLAST (KIL-BLAST)

Conventional V-BLAST architectures have been shown in [27] to be limited to achieving an overall diversity order of n_R - n_T + 1. The same level of diversity order is achieved by each individual layer. Hence for a square MIMO configuration with $n_T = n_R$, the maximum diversity order is 1. As shown in Chapter 3 for LST, the receiver processes the signals using successive interference cancellation, in which past decisions on decoded layers are used to cancel interference caused on the remaining layers.

The diversity order of 1 for square LST systems is dependent on the decisions for previous layers being made correctly and their interference contribution removed without any residual interference remaining. However, in practice, due the nature of the channel and noise at the receiver, achieving perfect cancellation without the residue is almost impossible. The highest contributing factor to this residual interference is the effect of imperfect channel state information (CSI) at the receiver. Although the broad assumption presented in most research work today is that the channel can be estimated perfectly at the receiver, a significant bandwidth overhead [28] is required to achieve this. Hence it is useful to develop signaling schemes which are resilient against the effects of imperfect CSI.

For layered space-time codes, the effect of imperfect CSI is additional to the effect of EP, which arises as a result of poor decoding of previous layers. For example, if the first layer to be detected is incorrectly decoded, its reconstruction is inaccurate and hence its interference is not fully suppressed. In this chapter the effects of EP are studied. In addition, imperfect CSI is modeled and performance degradation caused by it is demonstrated. A novel approach is presented which assumes that the signal on the first decoded layer is known at the receiver is presented. This scheme is called 'Known Interference Layer' BLAST (or KIL-BLAST) and was published in [40].

4.1 Error Propagation and the Genie-BLAST Concept

In order to provide a clearer understanding of the effects of EP, the concept of Genie-BLAST is introduced. For the remainder of this thesis, Genie BLAST is simply referred to as 'Genie'. Genie means that real interference suppression is performed for each layer, but for subsequent layers ideal detection of the signals of preceding layers is assumed. Hence the first decoded layer attains the lowest order of diversity. The next decoded layer attains a diversity order of one greater than the previous layer as the interference from the previous layer is assumed to be non existent.

In other words, for a $n_T = n_R = 4$ MIMO configuration, the second layer to be decoded (with perfect detection of the first layer) is analogous to the first layer of a $n_T = 3$ and $n_R = 4$ MIMO configuration. Therefore the diversity order of this layer is given as ($n_R - n_T + 1$) which is equal to two. Similarly, the third layer to be decoded (with perfect detection of the first and second layers) is analogous to the first layer of a $n_T = 2$ and $n_R = 4$ MIMO configuration with a diversity order of 3. Consequentially, the fourth layer to be decoded attains a diversity order of 4. In Figure 4-1 we present simulation results for a $n_T = n_R = 4$ MIMO configuration with QPSK modulation on each layer in the presence of Genie.

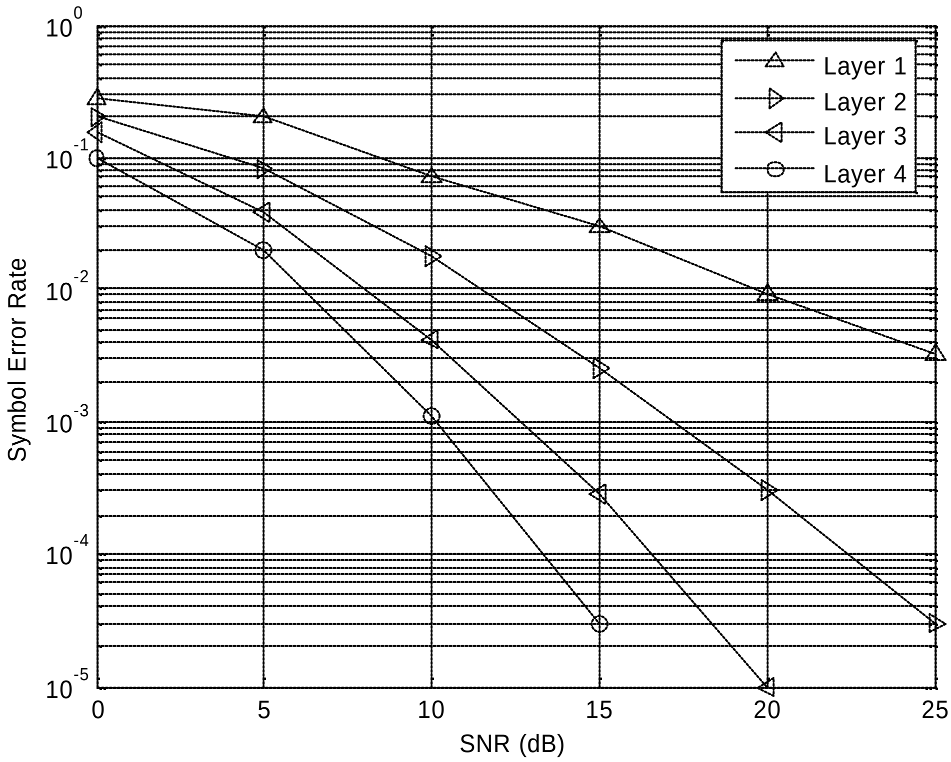

Figure 4-1 Genie BLAST performance for $n_T = n_R = 4$ with QPSK Modulation

Clearly, in the first decoding step the error curve decays in inverse proportion to the signal-to-noise ratio (SNR), therefore showing a diversity level of one [30]. The subsequent curves for antennas 2 through 4 have a much steeper decay rate. They take profit from the subtraction of previously detected symbols, thereby increasing the diversity level up to 4, which happens in the case when there is only one signal left to be detected by the four receive antennas.

In Figure 4-2 we demonstrates the effects caused by EP in the original V-BLAST scheme. Here it is observed that the poor decoding of the first layer dominates the over all system performance and all successive decoded layers achieve more or less the same performance and diversity order of 1.

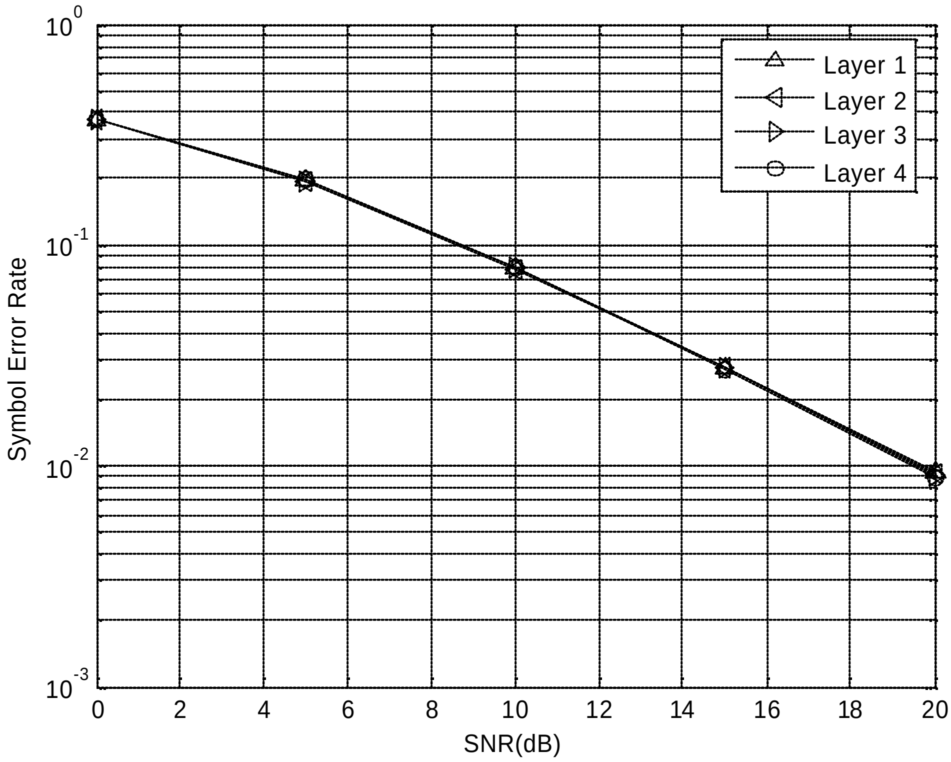

Figure 1-2 Error Propagation in original V-BLAST architecture

It is to be noted however, that each successive layer does achieve a slightly better performance than the preceding layer. To demonstrate this, Figure 4-3 provides a magnified depiction of the performance curves shown in Figure 4-2 at 20dB.

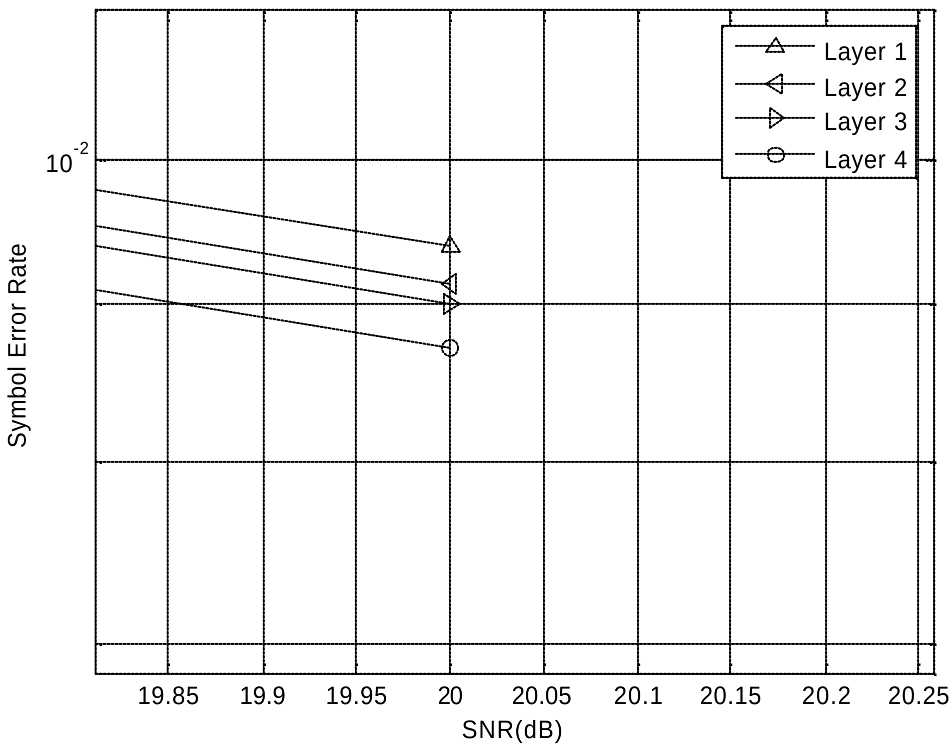

Figure 1-3 Marginal improvement in performance due to successive interference cancellation

4.2 KIL-BLAST and the Genie Concept

The requirement for high SNR on the first decoded layer in V-BLAST has been demonstrated in the previous sections. KIL-BLAST simply extends the Genie concept by assuming that the information transmitted on the layer to be decoded first is known at the receiver. Hence this scheme requires a pre-determined decoding sequence. This methodology effectively turns a $n_T = n_R = 4$ MIMO configuration into a $n_T = 3$ and $n_R = 4$ configuration. However, KIL-BLAST goes one stage further by changing the layer which transmits the known information based on the CSI provided by the receiver, i.e., a closed loop system. This concept has already been practically realized in CDMA to control the power of individual users. For KIL-BLAST, the receiver estimates the channel and sends a feedback signal to the transmitter to change the 'known information layer' for the remainder of the block length, assuming a quasi-static slow fading channel. In a fast fading environment, the complexity of this scheme is considerably lower than the water-filling algorithm presented in [8] as it simply requires switching the antenna on which the information sequence is sent.

In a conventional LST scheme with $n_T = n_R = 4$, all n_T antennas send out independent signals. For a BPSK modulation scheme with a frame size of 100 symbol intervals, the total number of symbols transmitted in each frame is 400. In KIL-BLAST however, the effective number of symbols transmitted in the whole frame equal to 300, which clearly shows a ¼ loss in the data throughput of the system. Figure 4-4 depicts the performance curves for a KIL-BLAST and original V-BLAST architecture with $n_T = n_R = 4$ configuration and QPSK modulation. The spectral efficiency of the V-BLAST architecture is 8b/s/Hz while that of the KIL-BLAST architecture is 6b/s/Hz. However, for this loss in throughput, it can be observed from the difference in the slope of the curves that a significant improvement in overall diversity order is achieved. At a SER of 1×10^{-2}, an improvement in SNR of approximately 9dB is achieved by KIL-BLAST over V-BLAST.

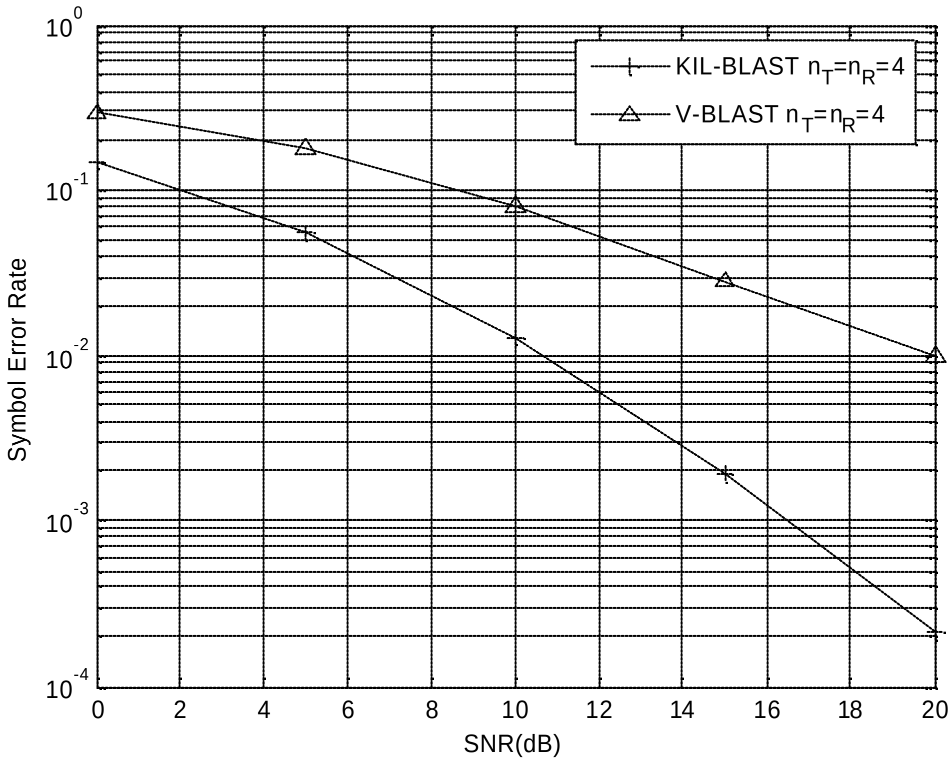

Figure 1-4 Increased diversity order provided by KIL-BLAST

Using a 2^q order modulation on each transmit antenna and with n_R receive antennas ($n_R = n_T$), the throughput for KIL-BLAST is given as $(qn_T - q)$ b/s/Hz as compared to V-BLAST which has a throughput of $(q \times n_T)$ b/s/Hz. Although there is a loss of q b/s/Hz throughput over V-BLAST, it is shown that with imperfect channel estimates at the receiver, better performance is obtained. This is due to the known interference induced into the channel, which helps obtain better estimate of the first detected substream. This also allows for the modulation order to be stepped up to 2^{q+1} or more and still obtain an improved error rate performance.

As perfect CSI cannot always be guaranteed at the receiver, its effect with respect to nulling and cancellation schemes can be regarded as unwanted interference. An error in channel estimation can be modeled as a noise/interferer term which is a Gaussian random variable (r.v) with zero mean and fixed variance s_e. In terms of RF impairments, error vector magnitude (EVM) can be considered as the variance s_e to be fixed for all values of SNR [32].

4.3 System Model

The KIL-BLAST architecture consists of n_T transmit and n_R receive antennas with $n_T = n_R$. The symbol transmitted from the k^{th} parallel stream is denoted as x_k and is drawn from a 2^q order constellation. In each transmission period the n_T-dimensional vector $\mathbf{x} = [x_1, \ldots, x_{n_T}]^T$ is transmitted with the symbol x_{n_T} known at the receiver. For a frame length of L=100 symbols, $x_{n_T,1\cdots L}$ is a random sequence whose components are randomly drawn from the 2^q order constellation so as to keep the variance of **x,** given as E[$\mathbf{xx}^H$] = (P/n_T) $\boldsymbol{I}_M$, constant. **P** is the total transmitted power for all n_T antennas, $(?)^H$ stands for the Hermitian transpose and $\boldsymbol{I}_M$ is a $n_T \times n_T$ identity matrix. **H** is the $n_R \times n_T$ channel impulse response matrix in which the elements $h_{i,j}$ are uncorrelated Gaussian r.v's with zero mean and unit variance, representing the gain between receive antenna i and transmit antenna j in a rich scattering environment. The received vector **r** is an $n_R \times 1$ component vector given as

$$\mathbf{r} = \mathbf{Hx} + \mathbf{n} \quad \textbf{Eq 1-1}$$

where **n** is a n_R×1 additive white Gaussian noise (AWGN) matrix with covariance (N_0/2) $\boldsymbol{I}_M$. The receiver processing is based on a QR-decomposition of the estimated channel matrix $\hat{\mathbf{H}} = \mathbf{H} + \hbar$ which gives the matrix $\hat{\mathbf{R}}$ which is upper right triangular and an orthogonal matrix $\hat{\mathbf{Q}}$. The n_T×1 decision statistics column vector **y** as given in [33] is

$$\mathbf{y} = \hat{\mathbf{Q}}\mathbf{r} = \hat{\mathbf{R}}\mathbf{x} - \hat{\mathbf{Q}}^H e\mathbf{x} + \hat{\mathbf{Q}}^H \mathbf{n} \qquad \text{Eq 1-2}$$

where e is a $n_R \times n_T$ channel estimation matrix with elements e_{ij} and are assumed to be complex Gaussian r.v's with zero mean and variance s_e. Higher values of s_e represent a poorer channel estimate at the receiver. From the decomposition of Eq 4-2, the decision statistic for the k^{th} substream can be given as

$$y_k = r_{k,k}x_k + \sum_{j=k+1,\forall k<n_T}^{n_T} r_{k,j}x_j + I_k + n_k \qquad \text{Eq 1-3}$$

where the second term is the interference from the preceding sub streams due to imperfect symbol detection and I_k is the interference term due to the channel estimation error.

4.4 Performance Results of KIL-BLAST

The detection of the first substream is critical to the performance of V-BLAST. As shown in Eq 4-3, the detection with respect to V-BLAST is now dependant on an extra term I_k which can be considered as a low power interferer. KIL-BLAST guarantees perfect detection of the first substream, as it already has knowledge of the symbols transmitted in this substream. This allows for the effect of this interference term to be suppressed and assures perfect decision feedback for the subsequent sub streams to be detected. Consequently, this improves the overall error rate performance in comparison to an equivalent V-BLAST architecture, and allows for a higher modulation level to be employed. In KIL-BLAST the diversity of each substream is increased from (n_R-n_T+1) to (n_R-n_T+2) which is equivalent to that of a V-BLAST n_T-1 transmit and n_R antenna configuration.

Figures 4-5 and 4-6 show the symbol error rate (SER) performance of four different schemes for increasing signal-to-noise ratio (SNR) with an estimation error matrix having variances of $s_e = 0.01$ and $s_e = 0.1$ respectively. For a $n_T = 3$ and $n_R = 3$ V-BLAST scheme with BPSK modulation and $s_e = 0.01$ (Figure 4-5), we obtain a SER performance of 10^{-2} at a SNR of 23dB and an error floor at SER of 7.5×10^{-3} similar to the results in [9]. With the same parameters, KIL-BLAST achieves an improvement of 11dB at the same SER owing to the trade-off in throughput on one antenna and with a lower error floor. There is also an improvement of 4dB over the equivalent V-BLAST n_T - 1 transmit and n_R receive architecture. It is important to note that the KIL-BLAST BPSK SER values presented here are for the two layers transmitting the useful data only for the purpose of fair comparison. We also consider a $n_T = 3$ and $n_R = 3$ KIL-BLAST scheme with QPSK modulation and $s_e = 0.01$, and observe an improvement of approximately 5.5dB at a SER of 10^{-2} with an error floor at approximately 5×10^{-4}. It can be observed from here that in the medium to high SNR region, a $n_T = 3$ and $n_R = 3$ V-BLAST BPSK scheme which provides a 3b/s/Hz can be outperformed by an equivalent $n_T = 3$ and $n_R = 3$ KIL-BLAST QPSK scheme which provides a 4b/s/Hz. Similar gains are achieved as shown in Figure 4-6 in the presence of a poorer channel estimate ($s_e = 0.1$), which demonstrates the robustness of KIL-BLAST with an increasing channel estimation error.

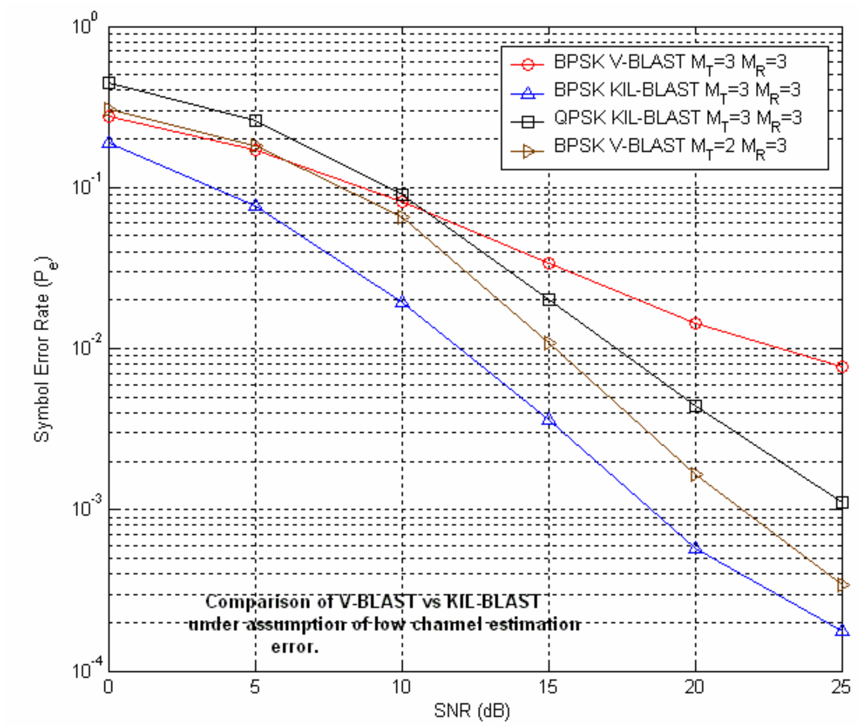

Figure 1-5 Comparison of V-BLAST versus KIL-BLAST under assumption of low channel estimation error

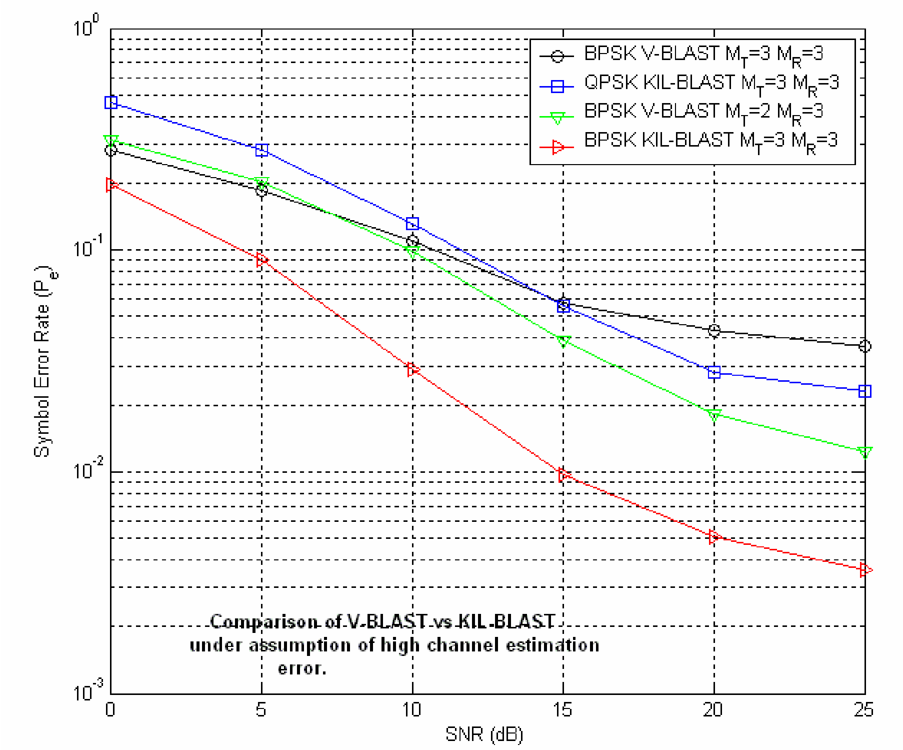

Figure 1-6 Comparison of V-BLAST vs. KIL-BLAST under assumption of high channel estimation error

4.5 Rate – Diversity Trade-offs offered by KIL-BLAST

KIL-BLAST has its own major drawback which is that there is a significant loss in the throughput of the system. This problem is easily resolved with the use of higher modulation levels on each of the other layers. Table 1 below shows a comparison of the throughput of different schemes with different modulation levels.

Modulation	*V-BLAST (bits per frame)*	*KIL-BLAST (bits per frame)*
BPSK	400	300
QPSK	800	600
8-PSK	1200	900
16-QAM	1600	1200

Table 1-1 Comparison of V-BLAST vs. KIL-BLAST data throughput, System with 100 symbol period frame length using n_T=4, n_R=4 configuration

It is a known fact that the higher the modulation level used for transmission, the lower the error rate performance of the system. This is because as the number of constellation points in a signal space increases; the minimum distance (d_{min}) between any two signal points is reduced making it more susceptible to channel effects [33]. Although we can achieve a higher data rate by using a higher modulation level as shown in the V-BLAST column of Table 1, the performance gets poorer as shown by the simulation result in Figure 4-7. At a BER of 10^{-2} the 8-PSK system would require approximately 9dB more power than a BPSK system and offers three times the throughput.

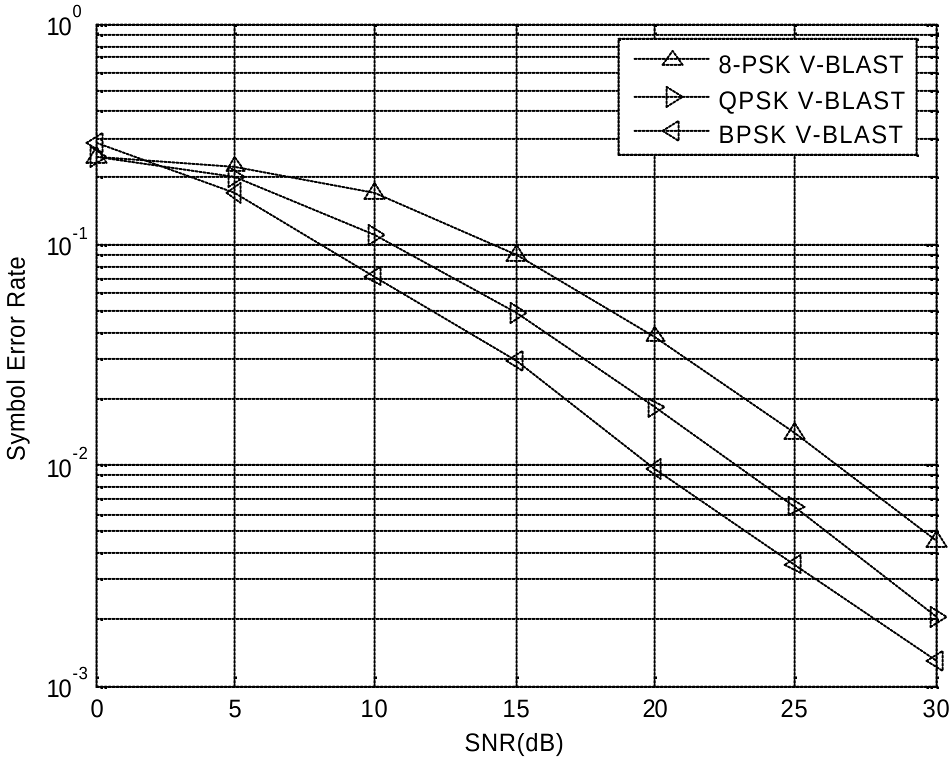

Figure 1-7 *M*-PSK Comparison for V-BLAST using $n_T = n_R = 4$

The 8-PSK KIL-BLAST system was compared to the V-BLAST BPSK and showed to have better performance Figure 4-8. The aim of system level design here was to develop a system which would provide the same performance as a BPSK system and a data-rate of a 8-PSK system. The 8-PSK KIL-BLAST has shown to provide slightly better performance than the BPSK system and more than double its data rate.

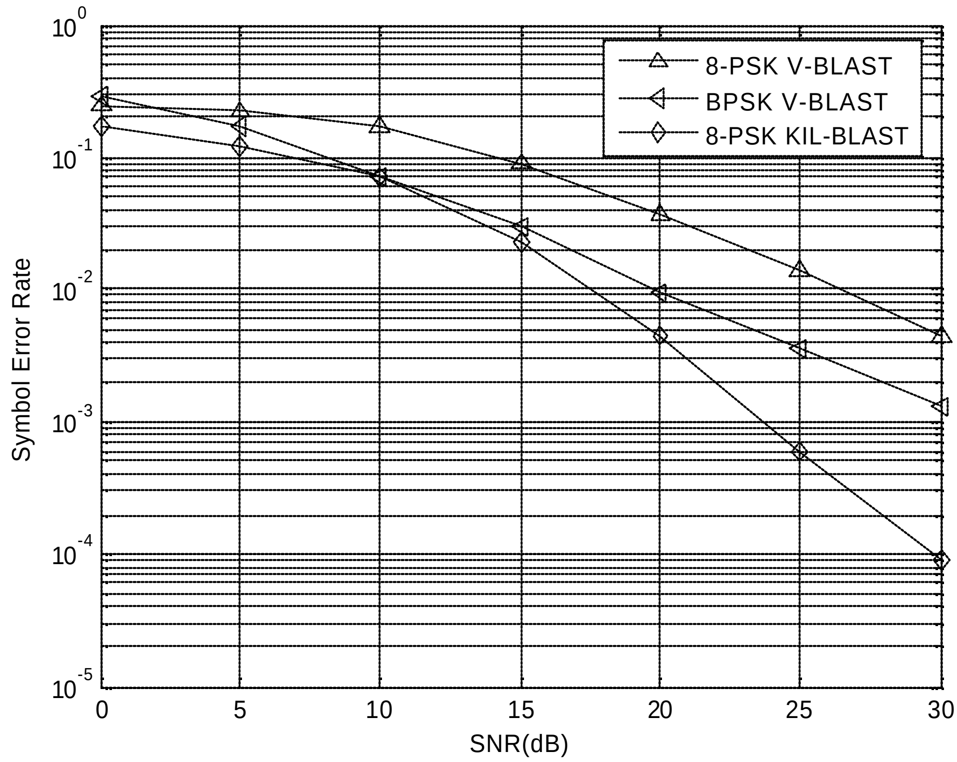

Figure 1-8 V-BLAST vs KIL-BLAST using $n_T = n_R = 4$

CHAPTER 5 LAYERED SPACE-TIME BLOCK CODES

As discussed in the introduction to this thesis, a MIMO system can provide two types of gains: diversity gain and spatial multiplexing gain. Most current research focuses on designing schemes to extract either maximal diversity gain *or* maximal spatial multiplexing gain. There are also schemes which switch between the two modes, depending on the instantaneous channel condition [35]. However, maximizing one type of gain may not necessarily maximize the other. For example, it was observed in [36] that the coding structure from the orthogonal designs for STBC [11], while achieving the full diversity gain, reduces the achievable spatial multiplexing gain. In fact, each of the two design goals addresses only one aspect of the problem. This makes it difficult to compare the performance between diversity-based and multiplexing-based schemes.

In [30] it is shown that for a given MIMO channel, both gains can in fact be achieved but there is a fundamental trade-off between how much of each type of gain can be extracted by any particular scheme. It is proved in [11] that a complex orthogonal design and the corresponding space–time block code which provides full diversity and full transmission rate is not possible for more than two antennas (see [11, Theorem 5.4.2]). For the specific cases of three and four transmit antennas [11] also proposed codes with 3/4 of the full transmission rate. In [37] a different strategy for designing space–time block codes was proposed. Here rate 1 codes that provide half of the maximum possible diversity for three and four transmit antennas were presented. However, this approach does not come close to yielding the rate $r_s = n_T$ provided by LST.

Conversely, LST codes have been shown to suffer from performance issues due to the low diversity order. As mentioned earlier in Chapter 4, the maximum diversity order achievable by V-BLAST is $n_R - n_T + 1$ with a constraint of $n_R \geq n_T$. KIL-BLAST presented in [40], targeted the diversity bottleneck caused by the first decoded layer in V-BLAST by attempting to limit the effect of EP. It was shown that the diversity order can be increased to $n_R - n_T + 2$ with a sacrifice in data rate which can be recovered by using higher modulation schemes. Similar methodologies were proposed in [29] and [30] which aimed at improving the diversity order of the first decoded layer by transmitting signals using a lower modulation level such as BPSK. The remaining layers would use higher modulation levels such as QPSK or 8-PSK.

Work presented in [29], [30], [37], [40] and [41] is geared towards sacrificing either diversity gain to improve spatial multiplexing gain, or sacrifice spatial multiplexing gain to improve diversity gain. In other words, the desire has been to obtain a suitable rate-diversity trade-off.

5.1 Literature Review of Rate-Diversity Trade-off Schemes

In [7], a combined array processing technique was developed which split the number of transmit antennas into small groups with individual space-time trellis codes, called component codes, and used to transmit information from each group of antennas. At the receiver a novel linear array processing technique called group interference suppression was adopted which treated signals from other groups as interference. In [7], different antenna groups are assigned different transmission power, hence the decoding order at the receiver is fixed (pre-ordered), namely, the group with the highest power allocation is decoded first. STBC codes were deployed over the grouped antennas in [26]. In [42] an optimum power allocation scheme was proposed.

The power difference allocated at the transmitter may not be maintained at the receiver after passing through a random channel. For a more reliable power allocation, a feedback channel could be used to obtain the CSI, but this reduces overall system throughput. Another possible solution is to increase the power difference, but this may cause more stringent requirements for RF design of amplifier and ADC. In [43] a post-ordering scheme with equal power allocation is proposed which generalises the ordered interference cancellation scheme mention in Chapter 3 for V-BLAST. Further, in [41], high rate optimal codes are designed where optimisation is performed over the choice of the number of layers, their size and different rate allocations. The decoder implemented in [41] is also based on the group interference suppression idea proposed in [7]. However, power allocation is also used in [41]. The approach used in all the above schemes is to sacrifice rate to improve diversity.

The linear dispersion frame work proposed in [44] spreads the symbols across space and time through matrix modulation and superposition with the objective of rate maximisation. Similar schemes designed for both diversity and rate maximisation were proposed in [35], [45] and [46], and these schemes can operate when $n_R < n_T$, however, for $n_R \geq n_T$, they provide no advantage over the V-BLAST scheme.

Following on from here, the second contribution of this thesis was published in [47]. This is a hybrid scheme called Layered Space-Time Block Codes (LSTBC), which combines both STBC and LST. However, compared to the schemes presented in [7], [26], [41], [42] and [43], LSTBC is unique in the following ways:

- It does not require complex power allocation computation as equal power transmission is used.
- At the transmitter, the split of the antennas does not group antennas specifically for STBC component codes. Instead, a single pair of antennas is grouped over which STBC component codes are deployed. The remaining $n_T - 2$ antennas transmit independent information similar to V-BLAST. This allows for a higher overall spectral efficiency than that provided by the schemes in [7], [26], [41], [42] and [43].
- The transmit architecture is simple and formation of the transmission matrix is done by simple matrix manipulation. The frame length of this scheme is 2 which is suited for both slow fading and fast fading environments.
- At the receiver, group interference suppression is not used. Instead a combination of MMSE and ML decoding is used in a closed loop form which maintains low complexity and yields high performance gains.

Subsequently in [48] and [49] a review of LSTBC receiver algorithms was presented which possessed some level of similarity to the algorithms presented earlier in [47]. Most recently in [50], the scheme developed in [47] was generalised for a combination of STBC and independent layers with more than two transmit antennas in the STBC layers.

5.2 LSTBC Concept

In ZF and MMSE for V-BLAST, the quality of decisions gradually increases as the receiver algorithm steps through the layers. The last decoded layer achieves a high SNR as interference from preceding layers is sequentially subtracted from the received signal as demonstrated in Chapters 3 and 4. However, the potential benefits of this high SNR signal are hardly exploited in the MMSE and ZF algorithms. This thesis now proposes an iterative scheme which re-decodes the signals in the lower layers by utilising the decision statistics of higher layers obtained in the first iteration is proposed.

5.3 System Model

A wireless communications system with $n_T > 2$ transmit antennas and $n_R = n_T$ receive antennas. Perfect CSI is assumed at the receiver and all antennas transmit with the same power. The channel is modeled by a $n_R \times n_T$ matrix **H** with elements h_{ij}, denoting the channel fading coefficient between transmit antenna j and receive antenna i. The coefficients h_{ij} are assumed constant over two symbol periods as in (1)and are modeled as independent samples of complex Gaussian random variables with mean zero and variance 0.5 per dimension.

5.3.1 Transmitter Model

Transmission frames have a length of two symbol periods and decoding is performed on a frame by frame basis. At each frame period, $2l + 2k(n_T - 2)$ bits arrive at the transmitter, where l and k are the number of bits per symbol chosen for the STBC layer and the independent layers respectively. The first $2l$ bits are forwarded to the 2^l-PSK modulators prior to STBC encoding as shown in Figure 5-1 (l=1). $\mathbf{G}_2$ represents the code which is utilized to encode the symbols on the STBC layer and is given as

$$\mathbf{G}_2 = \begin{bmatrix} x_1 & -x_2^* \\ x_2 & x_1^* \end{bmatrix} \qquad \textbf{Eq 5-1}$$

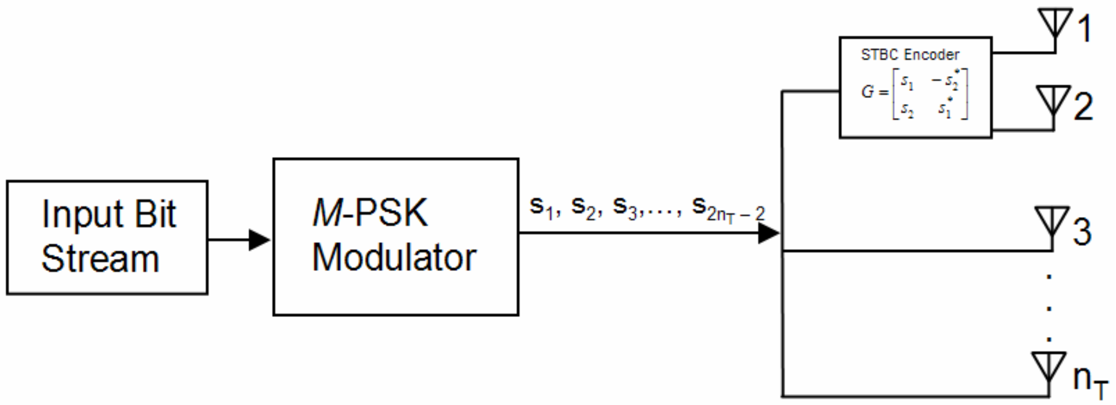

Figure 1-1 LSTBC Encoder

The remaining bits are split into two blocks and forwarded on to the $n_T - 2$, 2^k-PSK modulators, over two time periods. The $n_T \times 2$ transmitted column vector is given as

$$\mathbf{X} = \begin{bmatrix} x_1, x_2, \ldots, x_{n_T,1} \\ -x_2^*, x_1^*, \ldots, x_{n_T,2} \end{bmatrix} \qquad \text{Eq 1-2}$$

where $[\bullet]^T$ denotes the transpose operation and $x_{n_T,t}$ is the output of the n_T^{th} modulator at time t. As an example let $l = 1$, $k = 2$ and $n_T = 4$. Hence, the total number of bits arriving at the transmitter is 10. The first two bits are BPSK modulated and denoted as symbols x_1 and x_2. The remaining 8 bits are QPSK modulated and denoted as symbols x_3, x_4, x_5, x_6. The resulting transmitted vector is given as

$$\mathbf{X} = \begin{bmatrix} x_1 & -x_2^* \\ x_2 & x_1^* \\ x_3 & x_5 \\ x_4 & x_6 \end{bmatrix} \qquad \text{Eq 1-3}$$

providing an effective spectral efficiency of 5 bits/sec/Hz.

5.3.2 Receiver Model

Detection at the receiver is performed over two symbol periods. The received $n_R \times 1$ column vector at time t is given as

$$r_t = \mathbf{H}\mathbf{X}_t + N_t \qquad \text{Eq 1-4}$$

where $\mathbf{X}_t$ is the t^{th} column of $\mathbf{X}$ and N_t is the $n_R \times 1$ noise column vector at time t. The components of N_t are independent samples of a zero-mean complex Gaussian random variable with variance $N_o/2$ per dimension. For simplicity K is defined as the number of independent transmit antennas ($K = n_T - 2$).

In the first iteration an attempt is made to remove the interference in the STBC layer caused by the signals from the K independent antennas. We obtain a $K \times 1$ column vector y_t, which contains the soft estimates of the signals from the K independent layers at time t given as

$$y_t = \mathbf{w} r_t \qquad \textbf{Eq 1-5}$$

where **w** a is a $K \times n_R$ matrix similar to Eq 3-23 given as

$$\mathbf{w} = \left(\mathbf{h}^H \mathbf{h} + \frac{N_o}{2} I_{n_T} \right)^{-1} \mathbf{h}^H \qquad \textbf{Eq 1-6}$$

and **h** is a matrix containing the last K columns of matrix **H** given as

$$\mathbf{h} = \begin{bmatrix} & h_{1,3} & h_{1,4} & \cdots & h_{1,n_T} \\ \bullet\bullet & h_{2,3} & h_{2,4} & \cdots & h_{2,n_T} \\ & \vdots & \ddots & \ddots & \vdots \\ & h_{n_R,3} & h_{n_R,4} & \cdots & h_{n_R,n_T} \end{bmatrix} \qquad \textbf{Eq 1-7}$$

A new vector $\mathbf{z}_t$ is formed, which represents a soft estimate of the summed interference from the K independent layers at time t, and is given as

$$\mathbf{z}_t = \mathbf{h} y_t \qquad \textbf{Eq 1-8}$$

This interference is subtracted from the received vectors r_t to form the vectors $r_t^{'}$ as follows

$$r_t^{'} = r_t - \mathbf{z}_t \qquad \textbf{Eq 1-9}$$

The signals $r_t^{'}$ at times $t=1$ and 2 are forwarded to the STBC decoder given in [19] for the $\mathbf{G_2}$ matrix with n_R receive antennas, to produce the outputs $\hat{x}_1$ and $\hat{x}_2$, which are the estimates of the signals transmitted on antennas l and 2 respectively. To detect $\hat{x}_1$, the following decision metric is minimized

$$\left\| \left[\sum_{i=1}^{n_R} \left(r_1^i h_{1,i}^* + \left(r_2^i \right)^* h_{2,i} \right) \right] - x_1 \right|^2 + \left(-1 + \sum_{i=1}^{n_R} \sum_{j=1}^{2} \left| h_{i,j} \right|^2 \right) \left| x_1 \right|^2, \qquad \textbf{Eq 1-10}$$

over all possible values of x_1. Similarly to detect $\hat{x}_2$, the following decision metric is minimize

$$\left\| \left[\sum_{i=1}^{n_R} \left(r_1^i h_{2,i}^* - \left(r_2^i \right)^* h_{1,i} \right) \right] - x_2 \right|^2 + \left(-1 + \sum_{i=1}^{n_R} \sum_{j=1}^{2} \left| h_{i,j} \right|^2 \right) \left| x_2 \right|^2, \qquad \textbf{Eq 1-11}$$

over all possible values of x_2.

In the second iteration, the interference from the STBC layer is recreated and subtracted from the original received vector r_t as follows

$$r_t^{''} = r_t - \bar{\mathbf{h}} v, \quad v = \begin{cases} \begin{bmatrix} \hat{x}_1 & \hat{x}_2 \end{bmatrix}^T & \forall t = 1 \\ \begin{bmatrix} -\hat{x}_2^* & \hat{x}_1^* \end{bmatrix}^T & \forall t = 2 \end{cases} \qquad \textbf{Eq 1-12}$$

where $\bar{\mathbf{h}}$ is a matrix made up of the first two columns of **H** and is given as

$$\bar{\mathbf{h}} = \begin{bmatrix} h_{1,1} & h_{1,2} \\ h_{2,1} & h_{2,2} \\ \vdots & \vdots \\ h_{n_R,1} & h_{n_R,2} \end{bmatrix}. \qquad \textbf{Eq 1-13}$$

$\mathbf{r}'' = \begin{bmatrix} r_1'' \\ r_2'' \end{bmatrix}$ is the new received vector without the interference from the STBC layer. The algorithm then repeats steps Eq 5-5 and Eq 5-6, replacing r_t with r_t'' to produce a $K\times1$ column vector y_t'' given as

$$y_t' = \mathbf{w}r_t'', \qquad t=1,2. \qquad \textbf{Eq 1-14}$$

y_t' is then decoded using the 2^k-PSK demodulators to give us the recovered symbols $\hat{x}_{3,t} \ldots \hat{x}_{M_T,t}$ transmitted on the K in independent layers at time t. The bits received are obtained by a simple constellation de-mapping operation on the recovered symbols.

5.4 Performance Analysis of LSTBC

For a MIMO system it is assumed that the transmitter sends out a codeword $S^{(i)}$ over two time periods (t = 2) and the receiver mistakes it for another codeword $S^{(j)}$. The pairwise error probability (PEP), given knowledge of the channel realization is [51],

$$P\left(S^{(i)} \to S^{(j)}\right) \le \frac{1}{\left(\prod_{k=1}^{r(G_{i,j})} l_k\left(G_{i,j}\right)\right)^{n_R}} \left(\frac{r}{4n_T}\right)^{-r(G_{i,j})n_R} \qquad \textbf{Eq 1-15}$$

where $l_k\left(G_{i,j}\right)\left(k = 1,2,\ldots,r\left(G_{i,j}\right)\right)$ are the non-zero eigenvalues of $G_{i,j} = E_{i,j}E_{i,j}^{\ H}$ and $E_{i,j} = S^{(i)} - S^{(j)}$ is an $n_T \times 2$ codeword difference matrix. r is the SNR and $r\left(G_{i,j}\right)$ is the rank of $G_{i,j}$.

Receivers for spatial multiplexing (SM) systems treat each received signal vector as a codeword, i.e. t = 1, and perform ML decoding over every symbol vector. The performance analysis for any strictly SM systems follows easily once it is noticed that the codeword difference matrix $E_{i,j}$, is now a $n_T \times 1$ vector and $E_{i,j}E_{i,j}^{H}$ is a rank 1 matrix. From Eq 5-15, the PEP for SM can be written as

$$P\left(S^{(i)} \rightarrow S^{(j)}\right) \leq \frac{1}{l\left(G_{i,j}\right)^{M_R}} \left(\frac{r}{4n_T}\right)^{-n_R} \quad \textbf{Eq 1-16}$$

where $l\left(G_{i,j}\right) = E_{i,j}{}^{H} E_{i,j}$ since $G_{i,j}$ is rank 1. Hence the maximum diversity order for SM is n_R.

The codeword matrix S over two time periods transmitted by a 4 × 4 LSTBC scheme is

$$S = \begin{bmatrix} s_1 & s_5 \\ s_2 & s_6 \\ s_3 & -s_4^* \\ s_4 & s_3^* \end{bmatrix} \quad \textbf{Eq 1-17}$$

and the codeword difference matrix $E_{i,j}$ is of the form

$$E_{i,j} = S^{(i)} - S^{(j)} = \begin{bmatrix} e_1 & e_5 \\ e_2 & e_6 \\ e_3 & -e_4^* \\ e_4 & e_3^* \end{bmatrix}. \quad \textbf{Eq 1-18}$$

$E_{i,j}$ is not an orthogonal matrix as it does not satisfy the condition $E_{i,j}E_{i,j}^{H} = a\,\mathbf{I}$. Furthermore, $G_{i,j}$ is now a rank 2 matrix. Hence the PEP can now be derived from Eq 5-15 as

$$P\left(S^{(i)} \to S^{(j)}\right) \leq \frac{1}{\left(\prod_{k=1}^{r(G_{i,j})} l_k\left(G_{i,j}\right)\right)^{n_R}} \left(\frac{r}{4n_T}\right)^{-2n_R} \quad \textbf{Eq 1-19}$$

5.5 Simulation Results for LSTBC

In order to demonstrate the performance gains provided by the new proposed LSTBC architecture, initial comparisons are made with the Genie system presented in Chapter 4. Next, performance results of various LSTBC schemes are shown with different spectral efficiencies. This demonstrates the kind of rate diversity trade-off offered by LSTBC. Finally, comparisons are made with existing layered space-time architectures to demonstrate improvement in diversity order and trade-offs are applied to exploit the diversity gain in order to increase the spectral efficiency and match the spectral efficiency of the original V-BLAST architecture.

5.5.1 LSTBC Performance Compared with Genie

A LSTBC system with $n_T = n_R = 4$ and QPSK modulation on each layer is compared with an equivalent Genie system. The spectral efficiency provided by the LSTBC system, given $l = 2$ and $k = 2$, is 6b/s/Hz. The spectral efficiency of the Genie system is 8b/s/Hz. The results shown in Figure 5-2 compare the layer by layer performance of the two systems.

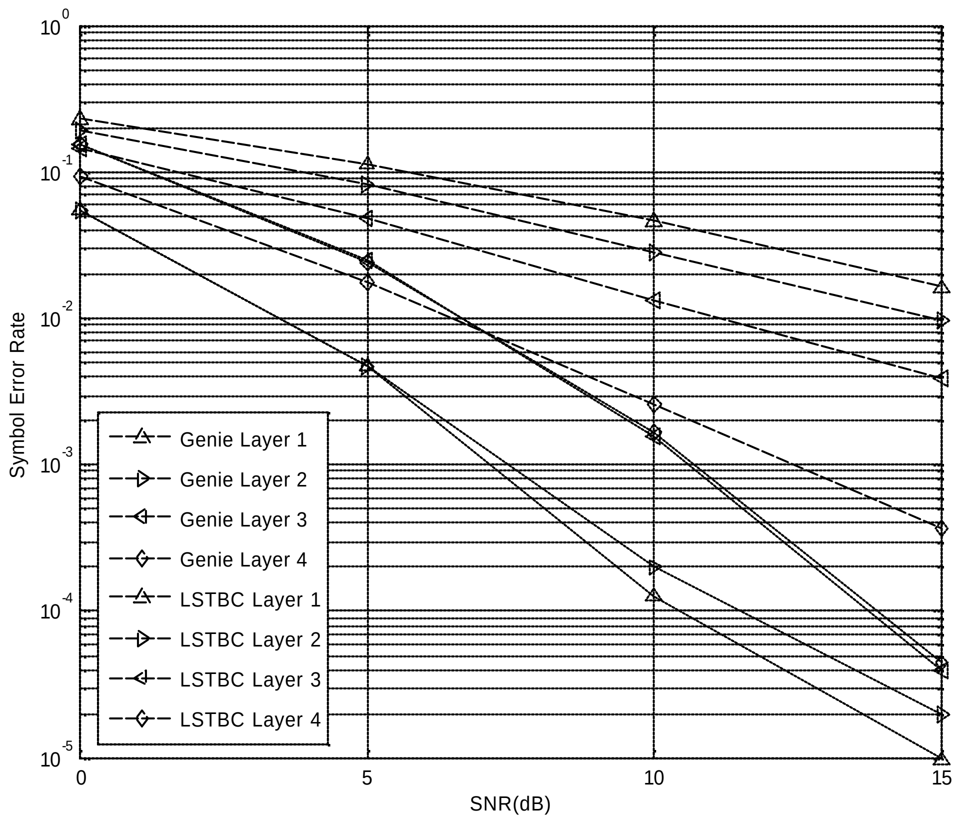

Figure 1-2 LSTBC vs. Genie $n_T = n_R = 4$, QPSK on each layer

The results in Figure 5-2 demonstrate that the performance of the first and second layers, which deploy STBC codes is significantly higher than those of the Genie system. The improvement in the second layer of Genie is as a result of perfect cancellation of the signal on layer 1. The LSTBC results for layers 1 and 2 suggests that in the first iteration, the interference of the independent layers is suppressed quite significantly. This in turn provides the overall system with high diversity first layer (the STBC layer) which yields the high diversity order for the upper layers. It can be observed that the diversity order and performance of the independent layers is the same.

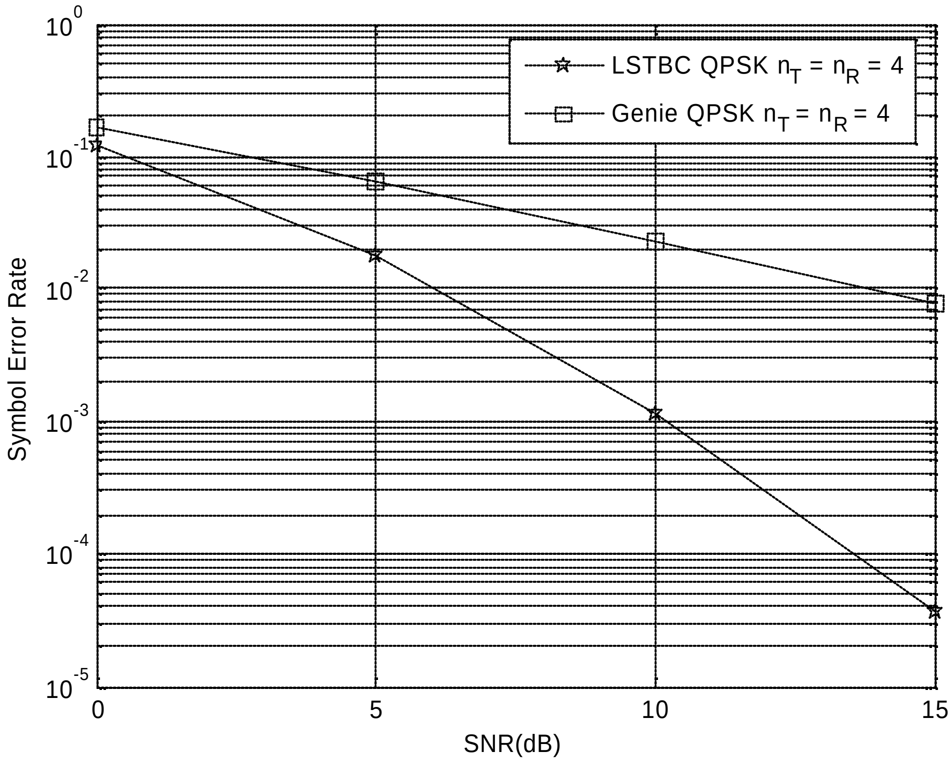

Figure 1-3 System Error Rate Performance LSTBC vs. Genie

Figure 5-3 compares the overall system SER performance of both schemes with increasing SNR. At a SER of 1×10^{-2}, LSTBC is seen to outperform Genie by approximately 8dB. Although the diversity order of the Genie system improves as the decoder steps from layer 1 through to 4, it can be seen that the performance of the first layer dominates the overall system performance yielding a diversity order of slightly less than 2. For LSTBC on the other hand, the high level of diversity order achieved by the STBC layer yields an overall system diversity order of 4.

5.5.2 Exploiting Diversity Gain to Improve Throughput

In Figure 5-4, a comparison is made between two LSTBC schemes with spectral efficiencies of 8b/s/Hz and the Genie scheme mentioned in Section 5.2.1 for $n_T = n_R = 4$. LSTBC scheme (a) deploys 16 QAM on the STBC layer and QPSK on the independent layers. LSTBC scheme (b) deploys QPSK on the STBC layer and 8-PSK on the independent layers. Both the schemes outperform Genie in the medium to high SNR region as shown. Although (a) and (b) provide the same spectral efficiency, the 2dB loss in performance of (a) over (b) at an SER of 1×10^{-3} is due to the higher modulation level used in the STBC layer.

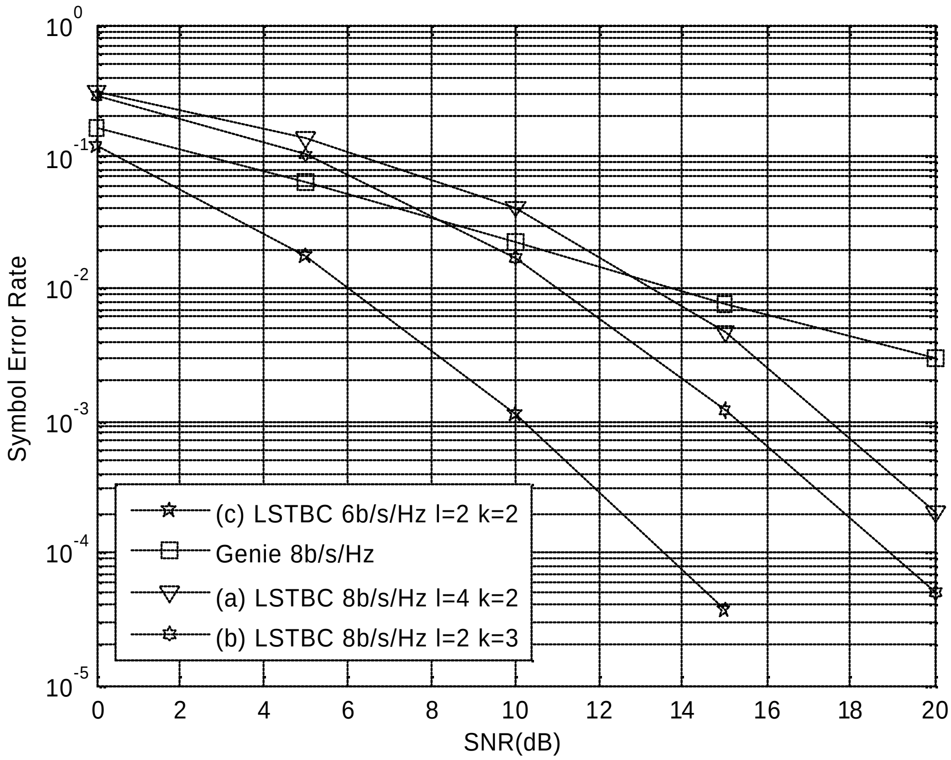

Figure 1-4 Increasing modulation level on STBC layer to improve rate at expense of diversity

In Figure 5-5, the SER performance of three LSTBC schemes is shown for $n_T = n_R = 4$. Here the modulation level on the STBC layer is kept constant using BPSK while that on the independent layer is stepped up from QPSK to 16QAM. At a SER of 1×10^{-2} it can be observed that for a stepped increase in spectral efficiency of 2b/s/Hz, the loss in SNR is approximately 5dB each. This means that although the diversity order of the system is significantly higher that the original V-BLAST architecture, the performance of the constellation in the independent layers results in poor performance.

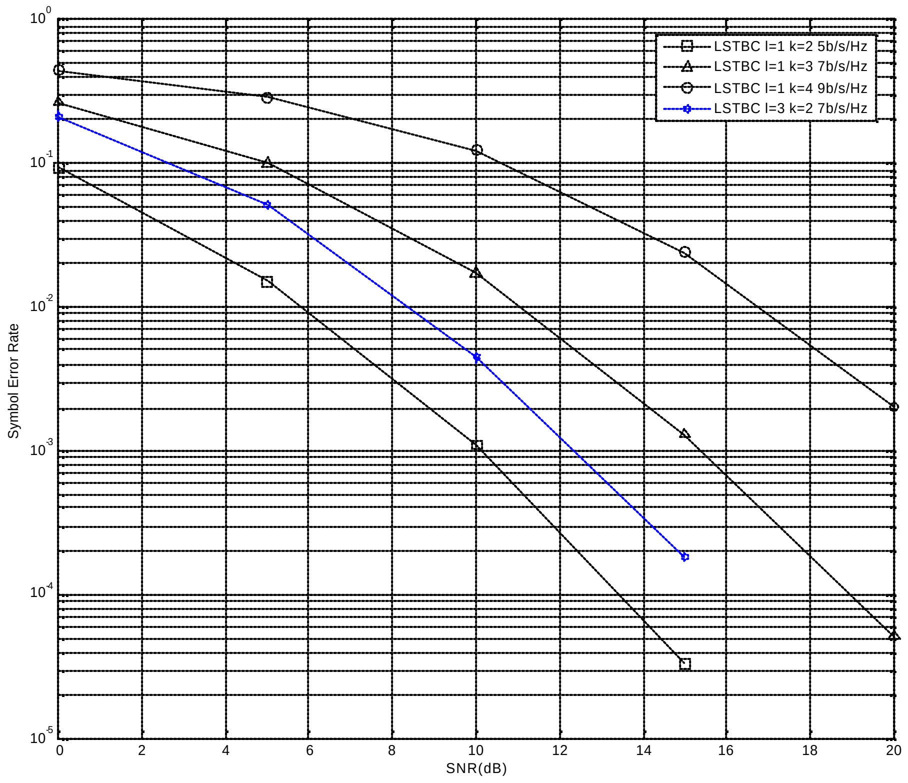

Figure 1-5 Performance comparison of independent layers to STBC layers for $n_T = n_R = 4$

Alternatively, the trade-off capabilities of LSTBC can be exploited in a different manner. A spectral efficiency of 7b/s/Hz can be achieved using two configurations, namely, $l = 1$ and $k = 3$ or $l = 3$ and $k = 2$. The first configuration puts the burden of the higher constellation on the independent layers while the latter places this burden on the STBC layer. As shown in Figure 5-5, the second option reduces the loss in SNR by approximately 2.5 dB at a SER of 1×10^{-2}.

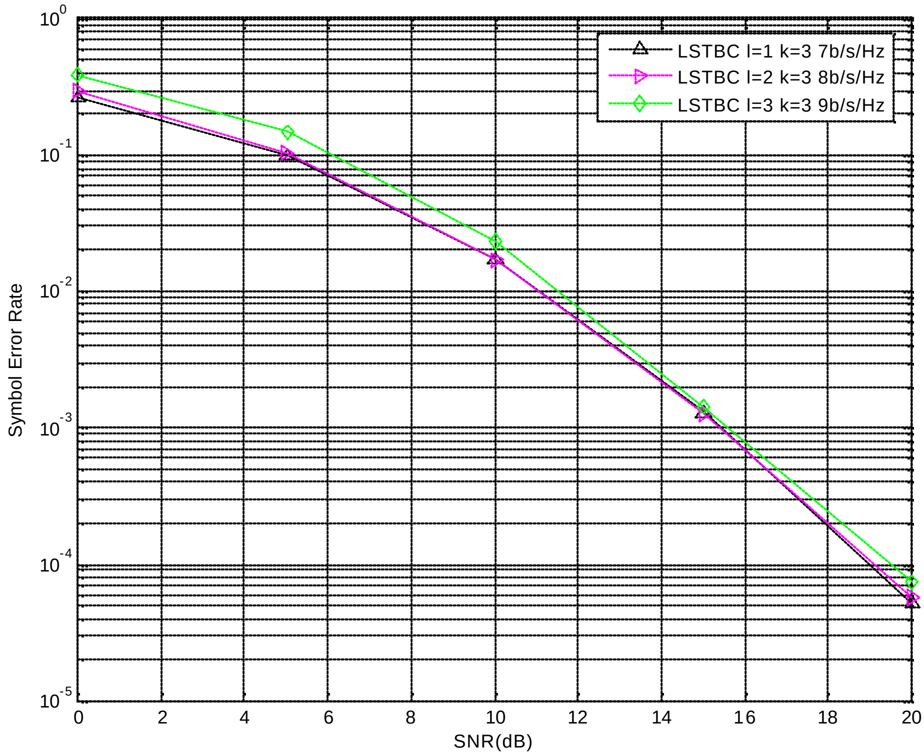

Figure 1-6 Increasing spectral efficiency using STBC layer only

Figure 5-6 demonstrates the performance of three LSTBC schemes with $n_T = n_R = 4$. In each scheme the modulation level on the independent layers is kept at 8-PSK while the modulation level of the STBC layer is stepped up from BPSK to 8PSK. As observed, the diversity gain and SER of all three schemes approximately match each other.

5.5.3 Comparison of LSTBC with existing BLAST schemes

In this section the performance of LSTBC with three layered space-time variants is compared with $n_T = n_R = 4$ providing more or less the same spectral efficiency. Figure 5-7 shows the SER performance curves. Conventional V-BLAST using QPSK on all four antennas and with a spectral efficiency of 8 bits/sec/Hz provides the lowest order of diversity. A modification of V-BLAST, suggested in [29] and [30] with BPSK on the first detected layer and QPSK on the remaining three layers with a spectral efficiency of 7 bits/sec/Hz is shown with an improvement of 2dB at a BER of 1×10^{-2}. However, the slope of the curve does not suggest any significant improvement in the diversity order achieved. The second technique suggested [26] using ordered successive interference cancellation with a spectral efficiency of 8 bits/sec/Hz and using ordered detection based on SNR is demonstrates a further improvement of 4dB. The slope of the curve demonstrates a negligible improvement in the diversity order. Finally we consider an LSTBC system with $l = 2$ and $k = 3$. The total number of bits transmitted over the two time periods given by $2l + 2k(n_T - 2)$ is 16. This provides an effective spectral efficiency of 8 bits/sec/Hz. The result shows an improvement of 4dB over the previous scheme at a BER of 1×10^{-2}. The performance curve indicates a significant improvement in the diversity order depicted by the increasing steepness of the slope with increasing SNR.

Figure 5-8 compares the performance of LSTBC with an increasing number of transmit and receive antennas with $n_T = n_R = 4$, 5 and 6. The improvement in performance is seen to be negligible and similar to that shown for MMSE V-BLAST in Figure 3-9. However, as the number of transmit and receive antennas increase proportionately, a slight loss in performance is observed.

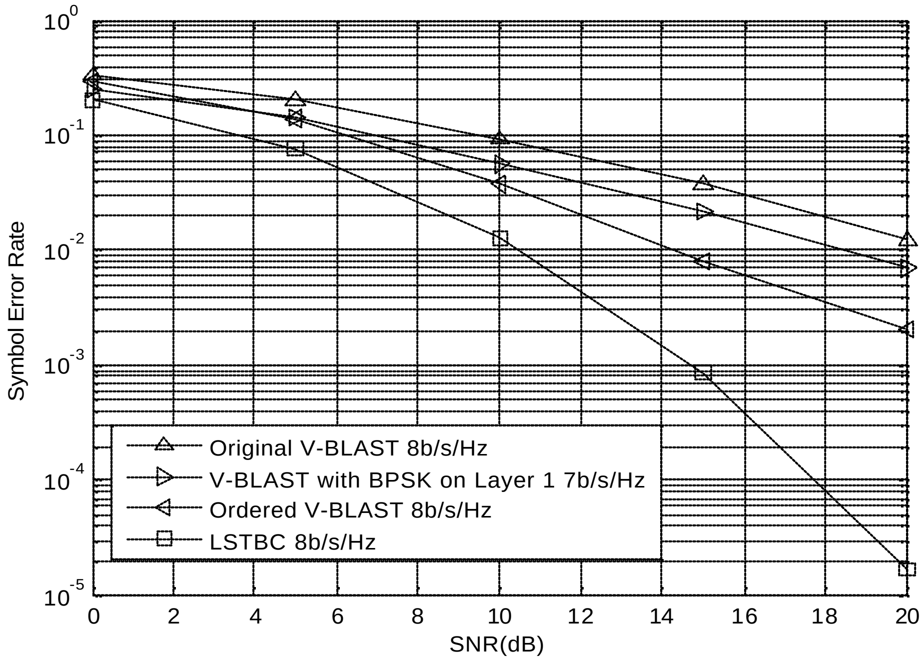

Figure 1-7 Comparison of LSTBC with existing BLAST architectures

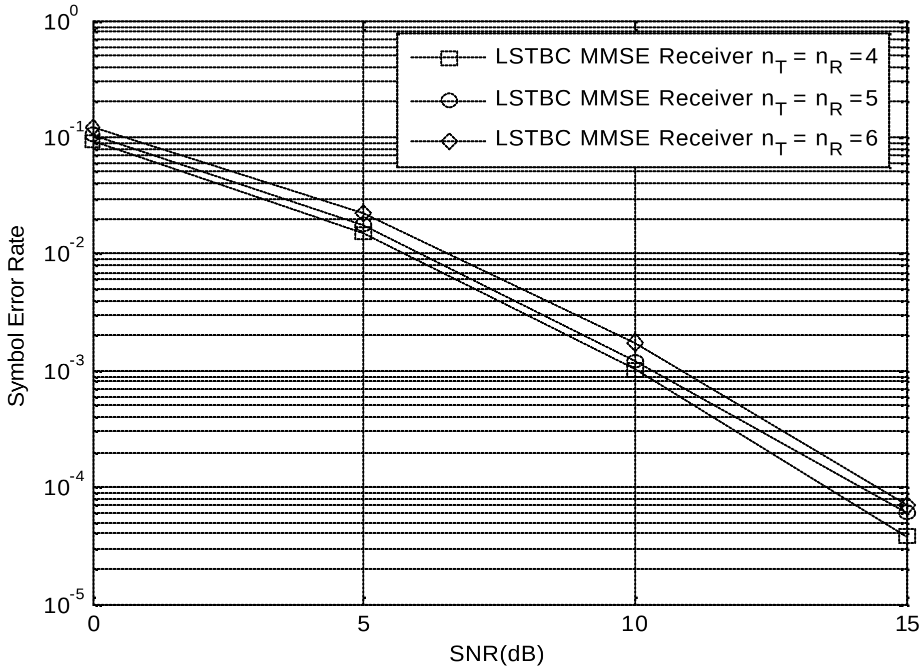

Figure 1-8 Diversity Performance of LSTBC for $n_T = n_R$ = 4, 5 and 6

5.6 MUD Performance of LSTBC

The iterative nulling and canceling approach used in V-BLAST and LSTBC is reminiscent of the successive interference canceling (SIC) proposed for multiuser detection (MUD) in CDMA receivers. In fact, any proposed MUD algorithm can be recast in the MIMO context if the input to the MIMO system are seen as virtual users. Here, we define an ideal MUD scenario as a number of co-located virtual single antenna users communicating with a common base station (BS). The BS is assumed to be equipped with the same number of antennas as the number of virtual users. It is further assumed that the BS has knowledge of all the users channels between the users and the BS. Another assumption taken into account is that all the users are equidistant from the BS and transmit at the same power. The results compare the performance of a V-BLAST scheme with a LSTBC scheme in this MUD scenario. Figure 5-9 shows the symbol error rate (SER) performance versus the number of users at an operating transmit power of 20 dB. The number of users indicates the number of antennas for each scheme.

A property of layered space-time systems is that the error rate performance at a fixed power remains steady with increasing number of antennas [52]. This indicates a linear growth in capacity with increasing number of users. The design of LSTBC as a layered space-time architecture was aimed at replicating this performance. In Figure 8, the performance of V-BLAST is shown to achieve an almost steady performance between 4×10^{-3} and 6×10^{-3}. Initial results of LSTBC in this MUD scenario demonstrated a very significant improvement in performance for a low number of users.

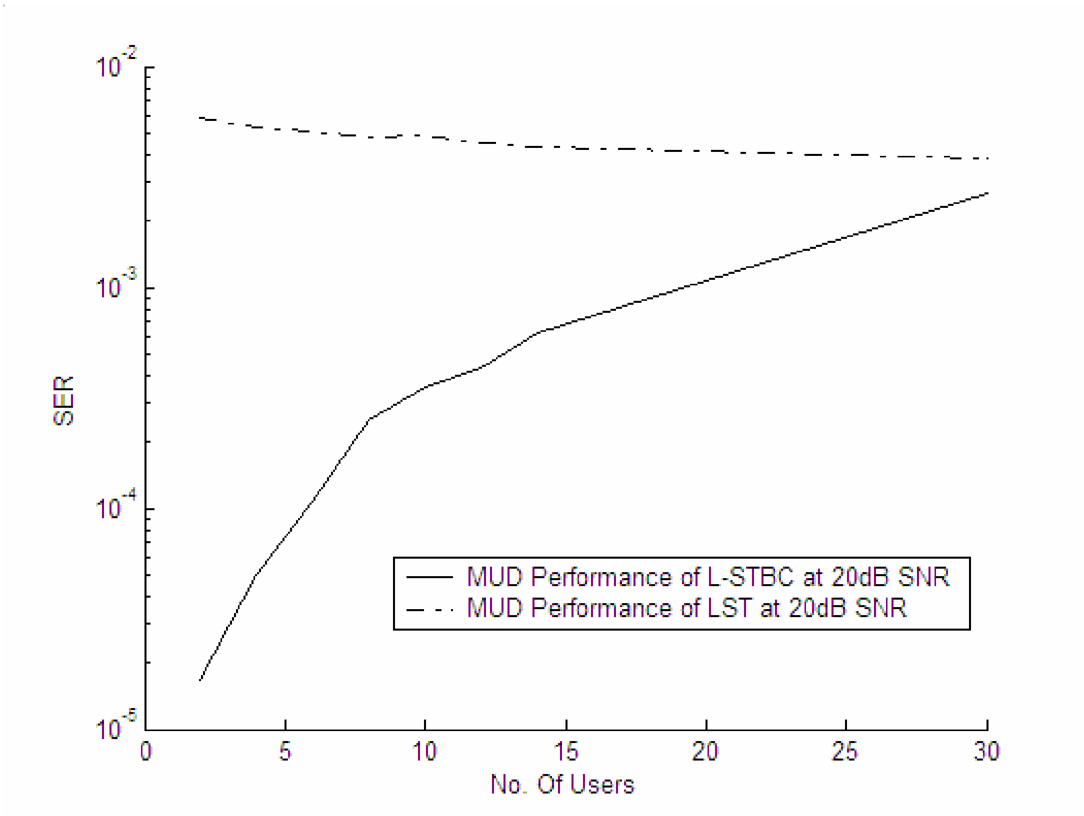

Figure 1-9 MUD performance comparison of LSTBC vs. LST for high number of users at 20 dB SNR

However, for an increasing number of users, the performance of LSTBC is seen to approach that of V-BLAST. This is because the STBC layer is overwhelmed with the increasing number of interferers and the system can no longer maintain the orthogonality of the STBC code. This effect is also depicted in Figure 5-8.

To counter this effect, this thesis now proposes a revised receiver algorithm. The revised algorithm constitutes two major changes to the weighting filter and the cancellation order. An optimized version of the weighting filter **w** in Eq 5-6 is given as

$$\mathbf{w}_{opt} = \left(\mathbf{h}^H \mathbf{h} + N_o I_{n_{T-2}}\right)^{-1} \mathbf{h}^H . \qquad \textbf{Eq 1-20}$$

The second change is the use of Ordered Successive Cancellation (OSUC) with the MMSE scheme. This detects the layer with the highest SNR, reducing the effects of EP. OSUC-MMSE is now used in the first iteration to detect the signals on the K independent layers. The estimates are therefore based on more robust soft decisions, providing improved interference cancellation by preserving the orthogonality of the STBC code. This is depicted by the slight improvement in performance of with increasing number of transmit and receive antennas as shown in Figure 5-10. The new MUD performance of LSTBC in Figure 5-11 achieves a significantly lower error floor than V-BLAST and with a similar steady performance between 2×10^{-5} and 6×10^{-5}. The OSUC-MMSE algorithm applied to LSTBC is shown in Table 5-1.

Initilizat ion
$i \leftarrow 1$
$\mathbf{r}_1 = \mathbf{r}$
$\mathbf{w}_1 = \left(\mathbf{h}^H \mathbf{h} + N_o I_{n_{T-2}}\right)^{-1} \mathbf{h}^H$
$k_1 = \arg\min_j \left\Vert (\mathbf{w}_i)_j \right\Vert^2$
Recursion
$\mathbf{w}_{k_1} = (\mathbf{w}_i)_{k_i}$
$y_{k_i} = \mathbf{w}_{k_i}^T \mathbf{r}_i$
$\hat{a}_{k_i} = Q(y_{k_i})$
$\mathbf{r}_{j+1} = \mathbf{r}_i - \hat{a}_{k_i} (\mathbf{h})_{k_i}$
$\mathbf{w}_{i+1} = \left(\mathbf{h}_i{}^H \mathbf{h}_i + N_o I_{n_{T-2}}\right)^{-1} \mathbf{h}_i{}^H$
$k_{i+1} = \arg \min_{j \notin (k_1,\ldots,k_i)} \left\Vert (\mathbf{w}_{i+1})_j \right\Vert^2$
$i = i+1$

Table 1-1 OSUC – MMSE Algorithm to improve MUD performance of LSTBC

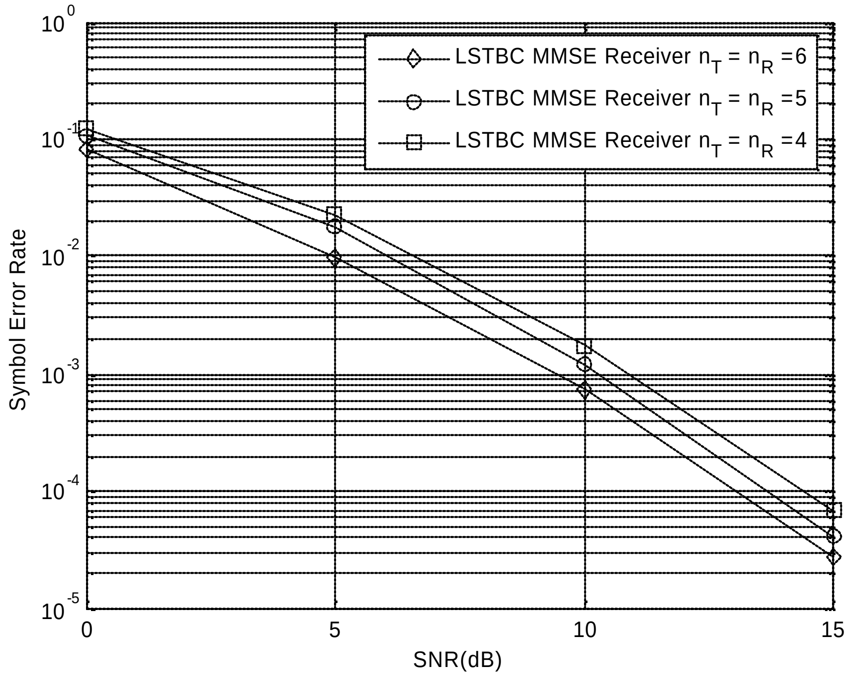

Figure 1-10 Improved Diversity Performance of LSTBC for $n_T = n_R = 4$, 5 and 6

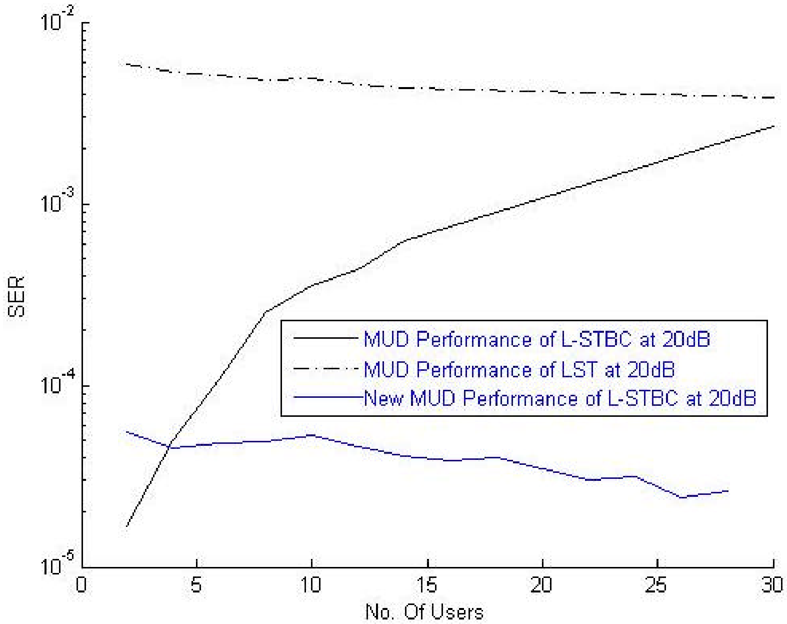

Figure 1-11 Improved MUD performance: Comparison of LSTBC vs. LST for high number of users at 20 dB SNR

CHAPTER 6 CONCLUSION

The concept of rate-diversity trade-offs was presented in the introduction to this thesis. The poor diversity performance of layered space-time codes was pointed out followed by the low rate structure of space-time block codes. Suggestions were made on combining both these schemes such that both rate and diversity benefits can be obtained using a common platform.

The contribution to this thesis starts from the study of the effect of interference on STBC as presented in Chapter 2. In this chapter simulation results were used to first demonstrate the improvement in diversity obtained with multiple transmit antennas and a single receive antenna, and was shown to be greater than that of a SISO link with 1 transmit and 1 receive antenna. Further, the improvement in diversity order with more than 1 receive antenna was presented. To study the coexistence of STBC with LST, an initial study to evaluate the performance of STBC in the presence of an interferer was carried out. Results in Chapter 2 demonstrated a degradation in performance due to loss of orthogonality of the STBC codes due to the interferer transmitting at the same power as the STBC antennas. These results were based on simulations with a single receive antenna. A further study with a higher number of receive antennas were demonstrated a very insignificant improvement in performance. Hence, it can be concluded from here that a more robust technique would need to be developed which would preserve the orthogonality of the space-time block codes.

In an attempt to exploit the spatial multiplexing gain provided by LST, the V-BLAST architecture which was applied later in this thesis was thoroughly explored. Receiver architectures using the ZF and MMSE algorithms were presented, and it was shown that the performance of the MMSE algorithm superseded that of the ZF algorithm. An ordered successive interference cancellation algorithm which starts by decoding the layer with the highest SNR was simulated and its performance compared to both the ZF and MMSE algorithms. An ordered successive interference cancellation algorithm which starts by decoding the layer with the highest SNR was simulated and its performance compared to both the ZF and MMSE algorithms. A modest improvement in performance was obtained. Overall it was demonstrated that the diversity order of layered space time block codes at best was given as $n_R - n_T + 1$. Further, the performances of the ZF and MMSE schemes were evaluated with increasing number of transmit and receive antennas. It was observed that the MMSE algorithm provided a modest increase in the performance. From here it was concluded that in layered space-time architectures the diversity order for square MIMO configurations was approximately 1.

Having demonstrated the poor diversity performance of LST, Chapter 4 presented a study on the effects of error propagation in layered space-time architectures as the root cause of poor diversity, with the relevant comparisons to the concept of a Genie BLAST system. It was shown that the limitation to the diversity order achieved by layered space-time architectures was based on the diversity order achieved by the first decoded layer. Also the effects of channel estimation errors in terms of error propagation enhancement were presented. A novel approach called KIL-BLAST was presented which the numerical results show guarantees a high diversity order for the first decoded layer. It was shown that even in the presence of channel estimation errors the effects of error propagation were suppressed significantly. From this study two conclusions were drawn:

- Firstly, improving the diversity order of the first decoded layer will yield very high performance results and the diversity order of layered space-time architectures is limited by error propagation.
- Secondly, the loss in throughput for KIL-BLAST motivated the research into a more sophisticated layered architecture which would not only improve the diversity order by suppressing error propagation, but also reduce the loss in throughput.

Finally, in Chapter 5, the drive to develop a system which would provide good rate-diversity trade-offs was presented. The motivation for this was that in a system combining the concepts of LST and STBC, the sacrifice in spectral efficiency should be kept to a minimum. Therefore, it is important to formulate a scheme in which STBC can co-exist with LST such that the capacity improvement of the latter can be exploited with a higher diversity performance supported by the former. A thorough literature review was presented which outlined some of the most relevant research work published. The aim was to demonstrate the work carried out until the contribution of Layered Space-Time Block Codes was published. Differences between existing schemes and that contributed in this thesis were made clear. The concept of LSTBC was introduced followed by the system model and performance results. It was shown through extensive numerical results that the performance of LSTBC was comparable to that of Genie BLAST in that the diversity order achieved by all layers reached the maximum diversity order of n_T. Further results were presented which demonstrated the use of LSTBC to provide good rate-diversity trade-offs, benefiting from the diversity order of the first decoded layer. Finally, an analogy to MUD scenarios in CDMA systems was made and it was shown that the performance of LSTBC systems was not suitable for an increasing number of users transmitting sychronously. An optimised model was developed and simulation results presented to show performance improvement in the presence of interferers.

The concepts of KIL-BLAST and LSTBC proposed in this thesis require the deployment of more than one antenna at the receiver. Currently these schemes could be thought of being more applicable to Point-2-Point MIMO systems. With stringent requirements on minimum antenna separation to reduce the effects of correlation, mobile devices would need to be of a relatively large size compared to handsets used today in order to obtain the benefits of these schemes. Therefore, reducing the number of receive antennas is a promising line of development. However, it is to be noted that reducing the number of receive antennas also means reducing the receive diversity advantage. From a performance point of view, this would provide an error rate performance curve with a slightly lower gradient as discussed in Chapters 4 and 5. As shown in Figure 5-7, LSTBC has an SNR advantage over conventional V-BLAST of approximately 10dB. A receiver algorithm which can accommodate a smaller number of receive antennas while preserving the orthogonality of the STBC code could possibly provide an SNR advantage less than 10dB, but still significantly greater than the other schemes compared in Figure 5-7.

6.1 Future Work

Following the publication of the research work contributed in this thesis, a significant amount of work related to the concept of LSTBC has been published, mostly driven by the same concepts and requirements as presented in Chapter 5. However, prior to the submission of this thesis, further proposals for future work were made based on the contributions of Chapters 4 and 5 and are summarised as follows:

- Development of a KIL-BLAST receiver architecture which can reduces the constraint on the number of receive antennas from $n_R \geq n_T$ to $n_R < n_T$.
- Development of LSTBC receiver architecture which can reduces the constraint on the number of receive antennas from $n_R \geq n_T$ to $n_R < n_T$.
- Improvement of diversity order for LSTBC using rate ¾ space-time block codes developed for three and four transmit antennas. This would require a slight increase in complexity at the receiver as decoding would have to be done over a longer number of symbol intervals.
- Application of quasi-orthogonal space-time block codes to LSTBC.
- Study of performance gains provided by LSTBC in the presence of a channel EVM as presented in Chapter 4.
- Application of more than one STBC layer to LSTBC and a study of the rate-diversity trade-offs achieved

LIST OF PUBLICATIONS

[1] Raza A., Schormans J. and Chen X., "KIL-BLAST: Diversity and Multiplexing Trade-Off for Layered Space-Time Architectures with Channel Estimation Error," *IEE Elec. Letters*, Vol. 40, No. 12, Pg: 747 – 749, 10th June 2004.

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