Iowa State University
Digital Repository

Retrospective Theses and Dissertations | Iowa State University Capstones, Theses and Dissertations

2007

Iterative post-SIC processing schemes in V-BLAST wireless MIMO communication systems

Wei-Tan Hsu
Iowa State University

Follow this and additional works at: https://lib.dr.iastate.edu/rtd

Part of the Electrical and Electronics Commons

Recommended Citation
Hsu, Wei-Tan, "Iterative post-SIC processing schemes in V-BLAST wireless MIMO communication systems" (2007). *Retrospective Theses and Dissertations*. 14552.
https://lib.dr.iastate.edu/rtd/14552

This Thesis is brought to you for free and open access by the Iowa State University Capstones, Theses and Dissertations at Iowa State University Digital Repository. It has been accepted for inclusion in Retrospective Theses and Dissertations by an authorized administrator of Iowa State University Digital Repository. For more information, please contact digirep@iastate.edu.

المنارة للاستشارات

www.manaraa.com

Iterative post-SIC processing schemes in V-BLAST wireless MIMO communication systems

by

Wei-Tan Hsu

A thesis submitted to the graduate faculty

in partial fulfillment of the requirements for the degree of

MASTER OF SCIENCE

Major: Electrical Engineering

Program of Study Committee:
Sang W. Kim, Major Professor
Zhengdao Wang
Johnny S. Wong

Iowa State University

Ames, Iowa

2007

Copyright © Wei-Tan Hsu, 2007. All rights reserved.

www.manaraa.com

UMI Number: 1443085

UMI Microform 1443085

Copyright 2007 by ProQuest Information and Learning Company.
All rights reserved. This microform edition is protected against
unauthorized copying under Title 17, United States Code.

ProQuest Information and Learning Company
300 North Zeeb Road
P.O. Box 1346
Ann Arbor, MI 48106-1346

للاستشارات

www.manaraa.com

DEDICATION

For my parents, who offered me unconditional love and support throughout the course of this thesis.

www.manaraa.com

TABLE OF CONTENTS

للاستشارات المنارة

www.manaraa.com

المنارة للاستشارات

www.manaraa.com

LIST OF TABLES

www.manaraa.com

LIST OF FIGURES

المنارة للاستشارات

www.manaraa.com

المنارة للاستشارات

www.manaraa.com

ACKNOWLEDGEMENTS

I am thankful to a lot of people who helped me with various aspects of conducting research and the writing of this thesis.

It is difficult to overstate my gratitude to my advisor, Dr. Sang W. Kim. With his enthusiasm, inspiration and great efforts, he helped to make research fun for me. Throughout my thesis-writing period, he provided encouragement, sound advice, good teaching, good company, and lots of good ideas. I would have been lost without him. "Thank you" does not seem to be sufficient but it is said with appreciation and respect.

I would like to express my thanks to Dr. Zhengdao Wang who was very patient and supportive in helping me solving problems encountered when writing this thesis; I have benefited a great deal from our discussions. Special thanks for his wonderful course on digital communication which enkindled my interest in this area as well.

I would also like to thank Dr. Johnny S. Wong for serving on my POS committee and being supportive.

The discussion with Dr. Aditya Ramamoorthy also helped me a lot. I appreciate his time and patience.

Last and not the least, warm thanks to my colleagues in the communication and signal processing research area of Iowa State University who assisted me during my graduate study and in finishing this thesis: Samarjit Das, Lei Ke, Naveen Kumar, Wei Lu, Kun Qiu, Jing Wang, Peng Yu and Dongbo Zhang.

المنارة للاستشارات

www.manaraa.com

ABSTRACT

In the conventional Vertical Bell Laboratory Layered Space-Time (V-BLAST) receiver with successive interference cancellation (SIC) decoding, the diversity order for the first detected symbol is the lowest, hence its error probability dominates the overall average error probability. In this thesis, a new SIC scheme was presented, called iterative post SIC (IP-SIC) that can increase the diversity order to a fixed desired value for *all* symbols, thereby significantly reduce the overall average error probability. The key to the technique is that after the interference from all substreams is subtracted from the received vector (the resulting vector will be referred to as the modified received vector), the detected symbol times its channel vector is added to the modified received vector one at a time and the symbol is detected again. Important features of the proposed approach are the increase in diversity order for those symbols detected earlier and the flexibility of balancing the increase in diversity order and the suppression of remaining interference. The latter feature can be used to further reduce the average error probability. The proposed technique is applied to the V-BLAST and space-time block coded V-BLAST system and its performance and computational complexity are analyzed.

المنارة للاستشارات

www.manaraa.com

CHAPTER 1. INTRODUCTION

This chapter presents some general introduction of the multiple-input multiple-output (MIMO) wireless communication systems and literature reviews. Contributions and outlines of this thesis will also be given in this chapter.

1.1 Overview of MIMO Wireless Communication Systems

Wireless communication is obviously the fastest growing area of the communication industry. In order to provide satisfactory services to customers with growing expectations and demands, wireless communication is always wished to be more reliable and have higher data rate. Nevertheless, many technical challenges remain in designing robust and fast wireless systems that deliver the performance necessary to support emerging applications, due to the fact that wireless channel is frequency selective, power-limited, susceptible to noise and interference. In addition, the radio wave propagation through the wireless channel may experience path loss caused by dissipation of transmit power and shadowing caused by obstacles between the transmitter and receiver that attenuate signal power through absorption, reflection, scattering and diffraction. Constructive and destructive addition of different multipath components may also be introduced by the wireless channel to form the fading effect, which is generally considered as a serious impairment to the wireless channel.

To fight against the effect of path loss, shadowing and multipath fading, MIMO systems utilizing multiple antennas at the transmit and/or receive end have been developed. The initial excitement of MIMO systems was sparked by the works of Winters [1], Foschini and Gans [2], and Telatar [3, 4]. Their works predicted the remarkable advantage of having more than single antenna – high spectral efficiencies. Multiple antennas form multiple channels be-

المنارة للاستشارات
www.manaraa.com

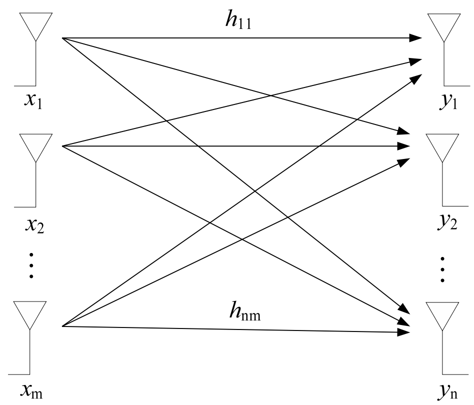

Figure 1.1 MIMO systems

tween the transmitter and receiver, thereby increasing signal *diversity* to make transmissions more reliable. *Multiplexing* can also be implemented using multiple antennas such that multiple data streams are able to be transmitted at the same time, resulting an increased data rate. To obtain these spectral efficiency improvements, we would often need knowledge of the channel condition, which is represented by the channel matrix. The cost of the performance enhancements achieved through MIMO techniques comes from deploying multiple antennas, the space and power requirements to install these extra antennas and the additional computing complexity to process multidimensional signals.

1.1.1 Mathematical Model of MIMO System

Here we consider a point-to-point wireless MIMO system with m transmitting and n receiving antennas, illustrated in Fig. 1.1. The following discrete-time model can be used to describe the system:

$$\begin{bmatrix} y_1 \\ \vdots \\ y_n \end{bmatrix} = \begin{bmatrix} h_{11} & \cdots & h_{1m} \\ \vdots & \ddots & \vdots \\ h_{n1} & \cdots & h_{nm} \end{bmatrix} \begin{bmatrix} x_1 \\ \vdots \\ x_m \end{bmatrix} + \begin{bmatrix} z_1 \\ \vdots \\ z_n \end{bmatrix} \tag{1.1}$$

Let $\mathbf{y} \triangleq [y_1, y_2, \cdots, y_n]^T$ be the $n \times 1$ received vector (^{T} denotes matrix transpose operation), $\mathbf{H} \triangleq [\mathbf{h}_1, \mathbf{h}_2, \cdots, \mathbf{h}_m]$ with $\mathbf{h}_i \triangleq [h_{1i}, h_{2i}, \cdots, h_{ni}]^T$ be the $n \times m$ matrix of channel gains whose element h_{ji} represents the gain from transmit antenna i to receive antenna j, $\mathbf{x} \triangleq$

المنارة للاستشارات
www.manaraa.com

$[x_1, x_2, \cdots, x_m]^T$ be the $m \times 1$ transmitted symbol, $\mathbf{z} \triangleq [z_1, z_2, \cdots, z_n]^T$ be the $n \times 1$ noise vector, then the model is rewritten as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{z} \tag{1.2}$$

1.1.2 Diversity in MIMO Systems

Many methods are available to achieve independent fading paths in a wireless system, including polarization diversity, directional (or angle) diversity, frequency diversity and spatial diversity [5]. MIMO systems are employed to obtain spatial diversity, that is, multiple antennas are separated in distance and the distance is large enough such that the fading amplitudes corresponding to each antenna are approximately independent, at the transmitter side and/or the receiver side [6]. By having independently faded signal replicas, more reliable reception than cases without diversity can be realized. For example, in a Rayleigh fading environment, if d independent antennas are used at the receiver end and single antenna is used at the transmitter side, then we could have a maximum (receiver) diversity gain of d and

$$d = -\lim_{\gamma \to \infty} \frac{\log P_e(\gamma)}{\log \gamma}, \tag{1.3}$$

where $P_e(\gamma)$ denotes the average error probability of a transmission scheme with signal-to-noise ratio (SNR) γ. Therefore, for high SNR, the average error probability decays as $1/\gamma^d$.

1.1.2.1 Receiver diversity in MIMO systems

For receiver diversity in MIMO systems, the independent fading paths corresponding to the multiple antennas are combined to get a signal which can be demodulated as usual. In most combining techniques, which we refer to as linear combining, the output of the combiner is just a weighted sum of the different fading paths. Fig. 1.2 represents this combining procedure. Some common techniques include:

- selection combining (SC)
- threshold combining (TC)

المنارة للاستشارات
www.manaraa.com

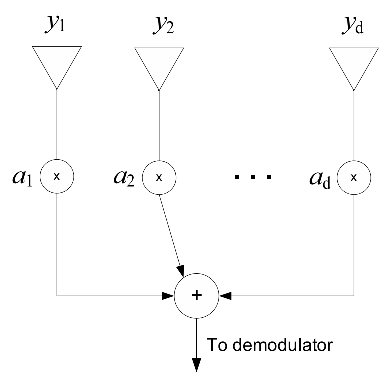

Figure 1.2 Linear combiner

- maximal-ratio combining (MRC)
- equal-gain combining (EGC)

When combining is made, the phase of every path or branch is removed, called co-phasing, to ensure the combiner output is coherent. After combining, the total received SNR γ_{Σ} would follow a more favorable distribution than the case with only single path is present; this would lead to a lower average error rate and thus the effect of fading is mitigated.

1.1.2.2 Transmitter diversity in MIMO systems

In transmit diversity when there are multiple transmit antennas, the transmit power is divided among these antennas. Transmitter diversity is often realized in systems that more space and power are available at the transmitter side than at the receiver end. The design of transmit diversity would depends on whether or not the transmitter knows the channel side information (CSI); if CSI is known to the transmitter, the case becomes quite similar to that of receiver diversity: the signal is weighted individually (multiplied by different complex gains) before sending to each antenna to transmit, in order to balance the channel fadings. Because the average total transmit energy is fixed, the sum of gains or weighting factors is subject to the total power constraint. Similar to linear combining in receiver diversity, the weighting factors are chosen to maximize the received SNR at the receiver side. However, when the CSI

www.manaraa.com

is not known at the transmitter, it is not possible to determine the different transmit gains for each transmit antenna to compensate fadings occurred in channels. To have transmit diversity in this case, it would require a blend of space and time diversity through the technology of space-time block or trellis codes. *Alamouti scheme* is one of the frequently used space-time block codes (STBC) to achieve transmit diversity for no CSI is available to the transmitter; a scheme involving this technique is presented in Chapter 4 of this thesis.

1.1.3 Spatial Multiplexing in MIMO Systems

By decomposing a MIMO system into several independent parallel spatial channels, spatial multiplexing is able to be implemented. We can get a data rate increase in comparison with a system with just one antenna at the transmitter and receiver respectively, by multiplexing independent data onto these independent channels [7].

Consider again a MIMO channel with $n \times m$ channel matrix $\mathbf{H}$: by performing singular value decomposition (SVD) on $\mathbf{H}$ we can obtain

$$\mathbf{H} = \mathbf{U}\mathbf{\Lambda}\mathbf{V}^H \tag{1.4}$$

where ^{H} denotes complex conjugate transpose, $n \times n$ matrix $\mathbf{U}$ and $m \times m$ matrix $\mathbf{V}$ are unitary matrices and $\mathbf{\Lambda}$ is an $n \times m$ diagonal matrix of singular values $\{\sqrt{\lambda_i}\}$ of $\mathbf{H}$; $r_{\mathbf{H}}$ of these singular values are nonzero, where $r_{\mathbf{H}}$ is the rank of matrix $\mathbf{H}$. Recall that $\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n}$, we could have

$$\begin{aligned} \tilde{\mathbf{y}} &= \mathbf{U}^H\mathbf{y} = \mathbf{U}^H(\mathbf{H}\mathbf{x} + \mathbf{n}) = \mathbf{U}^H(\mathbf{U}\mathbf{\Lambda}\mathbf{V}^H\mathbf{x} + \mathbf{n}) \\ &= \mathbf{U}^H(\mathbf{U}\mathbf{\Lambda}\mathbf{V}^H\mathbf{V}\tilde{\mathbf{x}} + \mathbf{n}) = \mathbf{\Lambda}\tilde{\mathbf{x}} + \tilde{\mathbf{n}} \end{aligned} \tag{1.5}$$

where the transformation of $\mathbf{x} = \mathbf{V}\tilde{\mathbf{x}}$ is called transmit precoding and the transformation of $\tilde{\mathbf{y}} = \mathbf{U}^H\mathbf{y}$ is called receiver shaping. Since $\tilde{\mathbf{n}} = \mathbf{U}^H\mathbf{n}$ and the original $\mathbf{n}$ are identically distributed, we obtain $r_{\mathbf{H}}$ parallel channels, and the ith channel can be represented by

$$\tilde{y}_i = \sqrt{\lambda_i}\tilde{x}_i + \tilde{n}_i \tag{1.6}$$

The resulting parallel channels do not interfere with each other, hence these channels are independent, linked only through the total power constraint. By sending independent data

المنارة للاستشارات
www.manaraa.com

across each of the parallel channels, the MIMO channel can support $r_{\mathbf{H}}$ times the data rate of a single transmit antenna single receive antenna system, yielding a multiplexing gain of $r_{\mathbf{H}}$.

1.1.4 Diversity and Multiplexing Trade-offs in MIMO Systems

Previous two subsections have introduced the two advantageous of MIMO systems over the traditional single antenna systems: more diversity and more multiplexing gain. Multiplexing different data streams onto equivalent decomposed parallel channels will produce an improved data rate, nevertheless the SNR related to these individual parallel channels depend on the singular values of the channel matrix $\mathbf{H}$; practical signaling strategies of these channels will typically bear poor reliability. Alternatively, we could utilize the multiple antennas for diversity such that independently faded channels are combined to create a more robust synthesized channel with low error probabilities.

Use multiple antennas or space dimensions purely for either diversity or multiplexing is not necessary; some resources can be devoted for diversity and the remaining for multiplexing. A trade-off between these two aspects is therefore established and this trade-off has been extensively researched in the literature [8–11], in terms of theoretical or channel capacity point of view, or from the perspective of practical space-time code designs. For example, in [8] a simple characterization of this trade-off for block fading channel in the limit of asymptotically high SNR is given: for each multiplexing gain $r_{\mathbf{H}}$, let the optimal diversity gain $d_{\text{opt}}(r_{\mathbf{H}})$ be the maximum diversity gain that can be achieved; if the fading blocklength $L \geq m+n-1$, then

$$d_{\text{opt}}(r_{\mathbf{H}}) = (m - r_{\mathbf{H}})(n - r_{\mathbf{H}}),\ 0 \leq r_{\mathbf{H}} \leq \min(m, n). \tag{1.7}$$

This implies that if all transmit and receive antennas are utilized for diversity, a full diversity gain $d = mn$ can be achieved and we can decrease the diversity gain to have some multiplexing gain.

Numerous models exist for MIMO systems; differences of them come from the way they adjust the diversity and multiplexing trade-off. Systems with bad channel condition would favor models with high diversity gain to ensure that the data transmission is reliable enough,

www.manaraa.com

whereas systems having good channel states may employ more antennas to boost up the data rate. Vertical Bell Laboratory Layered Space-Time (V-BLAST), the proposed schemes of this thesis are based on, is a model that provides very nice balance of diversity and multiplexing with relatively low complexity; we shall examine this model in next chapter.

1.2 Thesis Contribution

In this thesis, an extension procedure to the popular low-complexity successive interference cancellation (SIC) decoding algorithm for V-BLAST system is proposed; we name it iterative post-SIC (IP-SIC) processing. By applying the proposed IP-SIC scheme assuming no error propagation in SIC, the diversity order for *all* symbols can be increased to a fixed desired value to reduce the overall error probability significantly.

After interference from all substreams are subtracted from the received vector in SIC decoding, a modified received vector consisting of only noise and residues is obtained. The proposed IP-SIC scheme can then be applied by adding the product of SIC-detected symbol and its channel vector to the modified received vector one at a time; the symbol is detected again. Neglecting the error propagation, the diversity order for those symbols detected earlier can be increased and their detection reliability is enhanced, compared with SIC decoding only. On the other hand, when error propagation is taken into account, since the error probability of a specific substream is reduced, the error propagation effect on the remaining substreams is alleviated; therefore, the IP-SIC scheme is propitious for the last several substreams processed in SIC, although their diversity improvement is not dramatic.

It is demonstrated by analytical and simulation results that the proposed IP-SIC scheme is able to provide large power gains over the system decoded by SIC only. The additional complexity associated with the IP-SIC is however very small – the number of operations required for each substream is only several times of the number of receiving antennas.

Two alternatives of the IP-SIC scheme are also proposed in this thesis. One alternative balances the increase in diversity order and the suppression of remaining interference. Instead of using all degrees of freedom for diversity combining in IP-SIC, we utilize some degrees

المنارة للاستشارات
www.manaraa.com

of freedom to suppress several strong residues; the average error probability can be further reduced by this method. Another alternative applies the proposed IP-SIC scheme on joint ML/SIC V-BLAST system. An overall system improvement can be achieved by joint ML detecting the "bottleneck" substreams; the remaining substreams are to be SIC decoded and IP-SIC can be applied afterwards.

IP-SIC on STBC V-BLAST systems is also investigated in this thesis. Simulation results show that the IP-SIC scheme is also able to offer significant gains in these systems with group SIC, at a low additional complexity.

1.3 Thesis Outline

The remainder of this thesis is organized as follows. In Chapter 2, we first study the model of V-BLAST, two flavors of receivers and corresponding ordering schemes. Chapter 3 presents the proposed algorithm, its performance and complexity analysis and the two alternative schemes. An extension to the proposed scheme to include transmitter diversity involving STBC is included in Chapter 4. In Chapter 5 the results of the proposed schemes are given along with explanations of simulation strategies and parameters used. Conclusions and possible future directions are stated in Chapter 6.

المنارة للاستشارات
www.manaraa.com

CHAPTER 2. V-BLAST SYSTEM MODEL AND DECODING ALGORITHMS

2.1 Introduction

Among the plentiful schemes that have been proposed to exploit the high spectral efficiency of MIMO channels, Diagonal Bell Laboratory Layered Space-Time (D-BLAST) algorithm was invented by Foschini [12] to achieve a substantial part of the MIMO capacity. However, the high complexity of implementation is its major drawback. A simplified version of D-BLAST, V-BLAST, by Golden *et al.* [13], is relatively easy to implement while still able to reap a large portion of the high spectral efficiency. Research showed that V-BLAST system is equivalent to a decision feedback equalizer and is optimal in achieving the channel capacity [14, 15].

The main idea [16] of the V-BLAST architecture is to split the information bit stream into several substreams and transmit them in parallel using a set of transmit antennas at the same time and frequency; the number of substreams equals to the number of transmit antennas. At the receiver end, receive antennas obtain the substreams, which are mixed and superimposed by noise, due to the nature of the wireless propagation channel. Applying proper signal processing procedure, the receiver can separate the transmitted substreams so that the matrix wireless channel is transformed into a set of virtual parallel independent channels, given that multipath is rich enough. Although there exists an optimal detection scheme, maximum likelihood (ML) detection that can produce higher diversity gain by detecting multiple substreams simultaneously and minimize the error probability, its implementation complexity grows exponentially with the transmit symbols' constellation size and the number of transmit substreams. Successive interference cancellation (SIC), or sequential interference suppression, is another decoding approach that is more popular since it provides a reasonable balance between performance and

المنارة للاستشارات
www.manaraa.com

complexity. This SIC algorithm will be used in deriving the proposed schemes throughout this thesis.

2.2 System Model and Decoding Algorithm

2.2.1 System Model

We consider a V-BLAST system based on the general MIMO system model with m transmit antennas and n receive antennas ((1.1) or (1.2)), and the number of receive antennas is larger than or equal to the number of transmit antennas ($n \geq m$) so that the transmitted data substreams can be separated. At the transmitter side, the data stream is passed through a serial-to-parallel converter to be transformed into m substreams; each substream is sent through a different transmit antenna. All n receive antennas each receives the mixed signals from all m transmit antennas.

This model can also be interpreted as m users each has one antenna; all these users are transmitting data streams at the same time in the same frequency band to a base station, which is equipped with n receive antennas.

At each time, we have a received vector $\mathbf{y} = [y_1, y_2, \cdots, y_n]^T$ where the entry y_i corresponds to the obtained signal from receive antenna i. The $n \times m$ channel matrix $\mathbf{H}$ can be represented as $\mathbf{H} = [\mathbf{h}_1, \mathbf{h}_2, \cdots, \mathbf{h}_m]$ where $\mathbf{h}_i$ denotes the ith column of $\mathbf{H}$, $\mathbf{h}_i = [h_{1i}, h_{2i}, \cdots, h_{ni}]^T$. Elements of $\mathbf{H}$, h_{ji}, are modeled as independent identically distributed (i.i.d.) complex Gaussian random variables with zero mean and unit variance ($h_{ji} \sim \mathcal{CN}(0, 1)$); thus the fading environment is Rayleigh rich-scattering. The channel is assumed to be quasi-static random, that is, h_{ji} is fixed for every frame of information bits but varying from frame to frame. We also assume the channel matrix is perfectly known at the receiver. $\mathbf{x} = [x_1, x_2, \cdots, x_m]^T$ is the transmitted signal vector; each component x_i of $\mathbf{x}$ is an M-ary modulated symbol, or $x_i \in \mathcal{S}$, $\mathcal{S} = \{s_1, s_2, \cdots, s_M\}$; the average energy contained in each symbol is assumed to be E_s. The noise vector $\mathbf{z} = [z_1, z_2, \cdots, z_n]^T$ is modeled as white Gaussian – entries of $\mathbf{z}$ are i.i.d. complex Gaussian random variables with mean being zero and variance equal to $N_0/2$ per dimension ($z_i \sim \mathcal{CN}(0, N_0)$). The transmitted data symbols x_i, the channel gains h_{ji} and the noise z_i are

المنارة للاستشارات
www.manaraa.com

independent of each other. Altogether, the model is given by

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{z} = \sum_{i=1}^{m} x_i \mathbf{h}_i + \mathbf{z} \tag{2.1}$$

Two types of receivers are available for the SIC decoding algorithm, namely, zero-forcing (ZF) and minimum mean square error (MMSE) receivers; each is associated with various ordering schemes. These receivers along with their ordering schemes will be discussed in the remainder of this chapter.

2.2.2 ZF Receiver

2.2.2.1 Suppression and cancellation

The idea of ZF receiver is to create a suppression vector from the known channel matrix such that applying this suppression vector to the received vector will completely remove interference signals of all other substreams except the substream of interest; the additive noise vector will however be enhanced by the suppression vector as well. Let

$$\mathbf{W} = (\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H = [\mathbf{w}_1, \mathbf{w}_2, \cdots, \mathbf{w}_m]^T \tag{2.2}$$

be the pseudo-inverse of $\mathbf{H}$ and $\mathbf{w}_i = [w_{i1}, w_{i2}, \cdots, w_{in}]$ is the ith row of $\mathbf{W}$. $\mathbf{w}_i$ will be the suppression vector for data symbol x_i since it satisfies

$$\mathbf{w}_i\mathbf{h}_j = \delta_{ij}, \tag{2.3}$$

where δ_{ij} is the Kronecker delta function,

$$\delta_{ij} = \begin{cases} 1 & \text{if } i = j \\ 0 & \text{if } i \neq j \end{cases} \tag{2.4}$$

Therefore applying $\mathbf{w}_i$ on received vector $\mathbf{y}$ yields decision statistic v_i for symbol x_i:

$$v_i = \mathbf{w}_i\mathbf{y} = x_i + \mathbf{w}_i\mathbf{z} \tag{2.5}$$

The enhanced noise component $\mathbf{w}_i\mathbf{z}$ is a complex Gaussian random variable with zero mean and variance $\|\mathbf{w}_i\|^2 N_0/2$ per dimension ($\mathbf{w}_i\mathbf{z} \sim \mathcal{CN}(0, \|\mathbf{w}_i\|^2 N_0)$). To obtain an estimation of

المنارة للاستشارات
www.manaraa.com

x_i, $\hat{x}_i$, we need to apply the quantization or slicing function $f(\cdot)$ on v_i, based on the maximum *a posteriori* (MAP) probability decision. For example, if x_i's are binary phase shift keying (BPSK) modulated and $\pm\sqrt{E_s}$ are transmitted with equal probability, the quantization function will give

$$\hat{x}_i = f(v_i) = \begin{cases} +\sqrt{E_s} & \text{if } v_i \geq 0 \\ -\sqrt{E_s} & \text{if } v_i < 0. \end{cases} \tag{2.6}$$

After detecting x_i, the signal of this substream is cancelled from the received vector $\mathbf{y}$, producing a modified received vector $\mathbf{y}_1$,

$$\mathbf{y}_1 = \mathbf{y} - \hat{x}_i \mathbf{h}_i \tag{2.7}$$

and the channel matrix will be correspondingly modified, by removing the ith column, as

$$\mathbf{H}_1 = [\mathbf{h}_1, \cdots, \mathbf{h}_{i-1}, \mathbf{h}_{i+1}, \cdots, \mathbf{h}_m] \tag{2.8}$$

$\mathbf{H}_1$ will be used to calculate the suppression vector for the next substream. Detection and cancellation based on $\mathbf{y}_1$ will be similar, and we perform this SIC procedure for every substream until all the substreams have been detected.

It is clear that for the kth substream, applying suppression vector to the received vector will suppress the remaining $m-k$ substreams and combine $n-m+k$ diversity paths to generate the decision statistic. Indeed, by reducing the number of columns of the channel matrix by one every time we cancel a substream and assuming all substreams are perfectly detected and cancelled, the kth substream will have a diversity order of $n-m+k$, as shown in [17].

When processing the kth substream, we need to lower the probability of $P_{e,k} \triangleq \Pr(\hat{x}_k \neq x_k)$, the probability that the detection of $\hat{x}_k$ is in error; because if $\hat{x}_k \neq x_k$, the modified received vector will be affected and the detection of the remaining $m-k$ substreams will be influenced, resulting an effect called *error propagation*. Hence, a proper ordering in detecting substreams is desired for SIC, to ensure that the error propagation is minimized.

2.2.2.2 Ordering

<u>SNR based ordering</u>

المنارة للاستشارات
www.manaraa.com

The ordering determined based on SNR is the original ordering scheme invented along with the V-BLAST algorithm [13]. Since the error probability $P_{e,i}$ decreases with increasing SNR and the substream with the highest SNR introduces the largest interference on the remaining substreams, the substreams are detected and cancelled in order or largest SNR, *i.e.* at stage k of SIC, the substream with the highest SNR among all remaining $m-k+1$ substreams is detected and cancelled first.

For ZF receiver, it follows from (2.5) that the SNR γ_i for symbol x_i is

$$\gamma_i = \frac{|x_i|^2}{\|\mathbf{w}_i\|^2 N_0}. \tag{2.9}$$

For equi-energy signaling, we have $|x_i|^2 = E_s$ for all substreams. Hence the SNR ordered SIC proceeds in order of $1/\|\mathbf{w}_i\|^2$, *i.e.* the substream with the smallest $\|\mathbf{w}_i\|^2$ is cancelled first.

LLR based ordering

Log-Likelihood Ratio (LLR) based ordering technique for SIC was introduced in [18]. The major difference from the SNR-based ordering is that *a posteriori* information is taken into account. The decision statistic v_i after suppressing interference substreams will provide additional information. Therefore by utilizing the additional *a posteriori* information, the LLR-based ordering will outperform the SNR-based ordering.

The pairwise LLR of the symbol $\hat{x}_i$ to the t-th symbol in the symbol alphabet, s_t, $s_t \in \mathcal{S}$, is

$$\beta_{i,t} = \ln \frac{\Pr(x_i = \hat{x}_i | v_i)}{\Pr(x_i = s_t | v_i)}. \tag{2.10}$$

With the equality

$$\sum_{t=1}^{M} \Pr(x_i = s_t | v_i) = 1, \tag{2.11}$$

it is possible to express the probability of symbol error given v_i as

$$\Pr(x_i \neq \hat{x}_i | v_i) = 1 - \frac{1}{\sum_{t=1}^{M} \exp(-\beta_{i,t})}. \tag{2.12}$$

The probability $\Pr(x_i \neq \hat{x}_i | v_i)$ decreases along with decreasing $\sum_{t=1}^{M} \exp(-\beta_{i,t})$, therefore the LLR order would process the substream with the smallest $\sum_{t=1}^{M} \exp(-\beta_{i,t})$ value first. Thus similar to the SNR ordering case, at stage k of SIC, we compute $\sum_{t=1}^{M} \exp(-\beta_{i,t})$ for all

المنارة للاستشارات
www.manaraa.com

$m-k+1$ substreams yet to be processed and the one with smallest $\sum_{t=1}^{M}\exp(-\beta_{i,t})$ will be selected and cancelled.

If every symbol in $\mathcal{S}$ is to be transmitted with the same probability, then the pairwise LLR as in (2.10) can be rewritten as

$$\beta_{i,t} = \ln\frac{\Pr(v_i|x_i=\hat{x}_i)}{\Pr(v_i|x_i=s_t)}. \tag{2.13}$$

Since the conditional probability density of v_i given $x_i = s_t$ is

$$\Pr(v_i|x_i=s_t) = \frac{\exp(-\frac{|v_i-s_t|^2}{\|\mathbf{w}_i\|^2 N_0})}{\pi\|\mathbf{w}_i\|^2 N_0}, \tag{2.14}$$

the pairwise LLR is then

$$\beta_{i,t} = \frac{|v_i-s_t|^2 - |v_i-\hat{x}_i|^2}{\|\mathbf{w}_i\|^2 N_0}. \tag{2.15}$$

For example, for BPSK signals of energy E_s and assuming no error propagation from previous SIC stages, the *a posteriori* symbol error probability can be represented by

$$\Pr(x_i \neq \hat{x}_i|v_i) = \frac{1}{1+\exp(|\mu_i|)} \tag{2.16}$$

where

$$\mu_i = \ln\frac{\Pr(x_i=+\sqrt{E_s}|v_i)}{\Pr(x_i=-\sqrt{E_s}|v_i)} = \frac{4\sqrt{E_s}\,\mathfrak{Re}\{v_i\}}{\|\mathbf{w}_i\|^2 N_0} \tag{2.17}$$

and $\mathfrak{Re}\{v_i\}$ denotes the real part of v_i. According to the LLR ordering, we will choose the substream with the largest $|\mu_i|$, or equivalently, $|\mathfrak{Re}\{v_i\}|/\|\mathbf{w}_i\|^2$ to process.

2.2.3 MMSE Receiver

2.2.3.1 Detection and cancellation

Different from ZF receiver, the MMSE receiver does not completely remove the interference substreams; however, it does not enhance the additive noise part either, which is a main deficiency of ZF receiver. Rather, the MMSE receiver is aiming at minimizing the mean square error (MSE) $\mathbb{E}[(x_i-\hat{x}_i)^2]$, where $\mathbb{E}[\cdot]$ stands for the expected value. In general, MMSE receiver provides a good tradeoff between the noise and the interference and has better reliability performance over the ZF receiver [19].

المنارة للاستشارات
www.manaraa.com

For MMSE receiver, both the channel gain matrix $\mathbf{H}$ and the noise variance N_0 should be known at the receiver. The interference suppression matrix $\mathbf{W}$ is expressed as [15]

$$\mathbf{W} = [\mathbf{H}^H\mathbf{H} + \frac{N_0}{E_s}\mathbf{I}]^{-1}\mathbf{H}^H = [\mathbf{w}_1, \mathbf{w}_2, \cdots, \mathbf{w}_m]^T \tag{2.18}$$

where $\mathbf{I}$ is the identity matrix whose size is identical to that of $\mathbf{H}^H\mathbf{H}$. Unlike the ZF receiver, $\mathbf{w}_i$'s do not satisfy (2.3); in other words, applying $\mathbf{w}_i$'s on $\mathbf{y}$ will not completely remove other substreams, but they are able to yield $\min\{\mathbb{E}[(x_i - \mathbf{w}_i\mathbf{y})^2]\}$.

Similar to the ZF receiver, we perform MAP detection on $v_i = \mathbf{w}_i\mathbf{y}$ to get $\hat{x}_i$ for x_i, then cancel the decision $\hat{x}_i$ from $\mathbf{y}$ to yield $\mathbf{y}_1$, as in (2.7). The corresponding column $\mathbf{h}_i$ will be removed from $\mathbf{H}$, producing a modified channel matrix $\mathbf{H}_1$. The SIC proceeds in this manner until all substreams have been detected. Similar to the ZF receiver, the error propagation effect also exists in MMSE SIC scheme and thus ordering is necessary to minimize the error propagation.

2.2.3.2 Ordering

SNR based ordering

As in ZF receiver, the SNR γ_i for symbol x_i can be expressed as in (2.9). Hence, for equi-energy signaling, the SNR-ordered SIC cancels the substream with the smallest $\|\mathbf{w}_i\|^2$ value first.

SINR based ordering

An improved version, called signal to interference-plus-noise ratio (SINR) based ordering, is described in [20]. Rather than noise only, this scheme considers the joint effect of noise and interference. Applying $\|\mathbf{w}_i\|^2$ to the received vector gives

$$v_i = \mathbf{w}_i\mathbf{y} = \mathbf{w}_i(\mathbf{H}\mathbf{x} + \mathbf{z}) = x_i\mathbf{w}_i\mathbf{h}_i + \mathbf{w}_i\mathbf{z} + \sum_{j\neq i} x_j\mathbf{w}_i\mathbf{h}_j \tag{2.19}$$

Therefore the SINR ζ_i for the ith substream, assuming no error propagation from previous stages, can be written as

$$\zeta_i = \frac{E_s|\mathbf{w}_i\mathbf{h}_i|^2}{\sum_{j\neq i}|\mathbf{w}_i\mathbf{h}_j|^2 E_s + \|\mathbf{w}_i\|^2 N_0} \tag{2.20}$$

المنارة للاستشارات
www.manaraa.com

Since the symbol error probability decreases as the SINR increases, we cancel the substream with the largest ζ_i value first, to minimize the error probability.

LLR based ordering

A posteriori information can also be utilized for ordering MMSE SIC scheme [21]. Consider the simple case where BPSK modulation is used and the number of substreams at the transmitter side is $m = 2$. The LLR μ_i of the ith substream is given by

$$\mu_i = \ln \frac{\Pr(x_i = +\sqrt{E_s}|v_i)}{\Pr(x_i = -\sqrt{E_s}|v_i)} \tag{2.21}$$

Applying Baye's Rule, μ_i can be expressed as

$$\mu_i = \ln \frac{\Pr(v_i|x_i = +\sqrt{E_s})}{\Pr(v_i|x_i = -\sqrt{E_s})} + \ln \frac{\Pr(x_i = +\sqrt{E_s})}{\Pr(x_i = -\sqrt{E_s})} \tag{2.22}$$

The conditional probability density of v_i given $x_i \pm \sqrt{E_s}$ and $x_j \pm \sqrt{E_s}$, $i = 1, 2$, $j = 1, 2$, $i \neq j$, is

$$\Pr(v_i|x_i, x_j) = \frac{\exp(-\frac{|v_i \mp \mathbf{w}_i\mathbf{h}_i\sqrt{E_s} \mp \mathbf{w}_i\mathbf{h}_j\sqrt{E_s}|^2}{\|\mathbf{w}_i\|^2 N_0}) + \exp(-\frac{|v_i \mp \mathbf{w}_i\mathbf{h}_i\sqrt{E_s} \pm \mathbf{w}_i\mathbf{h}_j\sqrt{E_s}|^2}{\|\mathbf{w}_i\|^2 N_0})}{2\pi\|\mathbf{w}_i\|^2 N_0}, \tag{2.23}$$

Hence, for equi-probable source, μ_i is given by

$$\mu_i = \ln \frac{\exp(-\frac{|v_i - \mathbf{w}_i\mathbf{h}_i\sqrt{E_s} - \mathbf{w}_i\mathbf{h}_j\sqrt{E_s}|^2}{\|\mathbf{w}_i\|^2 N_0}) + \exp(-\frac{|v_i - \mathbf{w}_i\mathbf{h}_i\sqrt{E_s} + \mathbf{w}_i\mathbf{h}_j\sqrt{E_s}|^2}{\|\mathbf{w}_i\|^2 N_0})}{\exp(-\frac{|v_i + \mathbf{w}_i\mathbf{h}_i\sqrt{E_s} - \mathbf{w}_i\mathbf{h}_j\sqrt{E_s}|^2}{\|\mathbf{w}_i\|^2 N_0}) + \exp(-\frac{|v_i + \mathbf{w}_i\mathbf{h}_i\sqrt{E_s} + \mathbf{w}_i\mathbf{h}_j\sqrt{E_s}|^2}{\|\mathbf{w}_i\|^2 N_0})}, \tag{2.24}$$

with $i = 1, 2$, $j = 1, 2$, $i \neq j$. This expression will grow exponentially as m increases; thus some approximation is desired. For large m, the condition

$$\sum_{j \neq i} \mathbb{E}[(\mathfrak{Re}\{\mathbf{w}_i\mathbf{h}_j\})^2] \gg \mathbb{E}[(\mathfrak{Re}\{\mathbf{w}_i\mathbf{h}_i\})^2] \tag{2.25}$$

is true in general, hence the real part of the random variable $\sum_{j \neq i} x_i\mathbf{w}_i\mathbf{h}_j$, which appeared in (2.19), can be approximated as Gaussian with zero mean and variance $E_s \sum_{j \neq i}(\mathfrak{Re}\{\mathbf{w}_i\mathbf{h}_j\})^2$. Thus, the random variable $\mathfrak{Re}\{v_i\}$ given $x_i = \pm\sqrt{E_s}$ can be approximated as a Gaussian random variable with mean $\pm\sqrt{E_s}\mathbf{w}_i\mathbf{h}_i$ and variance $E_s \sum_{j \neq i}(\mathfrak{Re}\{\mathbf{w}_i\mathbf{h}_j\})^2 + \|\mathbf{w}_i\|^2 N_0/2$. Therefore (2.24) boils down to [21]

$$\mu_i = \frac{4\mathfrak{Re}\{v_i\}\mathbf{w}_i\mathbf{h}_i\sqrt{E_s}}{2E_s \sum_{j \neq i}(\mathfrak{Re}\{\mathbf{w}_i\mathbf{h}_j\})^2 + \|\mathbf{w}_i\|^2 N_0} \tag{2.26}$$

المنارة للاستشارات
www.manaraa.com

Similar to the ZF receiver case, the substream with the largest $|\mu_i|$ value cancelled first.

For general M-ary signaling, the pairwise LLR will be defined the same way as in ZF SIC (2.13). Using approximations similar to the BPSK case, the pairwise LLR can be shown as [21]

$$\beta_{i,t} = \frac{|v_i - s_t \mathbf{w}_i \mathbf{h}_i|^2 - |v_i - \hat{x}_i \mathbf{w}_i \mathbf{h}_i|^2}{E_s \sum_{j \neq i} |\mathbf{w}_i \mathbf{h}_j|^2 + \|\mathbf{w}_i\|^2 N_0} \tag{2.27}$$

Then, the substream with the smallest $\sum_{t=1}^{M} \exp(-\beta_{i,t})$ value is cancelled first.

المنارة للاستشارات
www.manaraa.com

CHAPTER 3. ITERATIVE POST-SIC

In this chapter, the proposed scheme for V-BLAST system with SIC decoding and alternatives are presented. The key idea of the proposed scheme is that after the interference from all substreams is removed from the received vector, the detected symbol times its channel vector is added to the modified received vector and we detect the symbol again. Assuming no error propagation, the diversity order of all symbols can be increased to a fixed desired value not exceeding the number of receive antennas through the proposed scheme; the overall average error probability can thereby be significantly reduced.

3.1 System Model

If the conventional SIC decoding algorithm is employed to separate the transmitted substreams, and the substreams are processed in the order of $(1), (2), \cdots, (m)$, then at the kth stage of SIC, $m-k$ interfering substream is to be suppressed by the ZF receiver, or to be suppressed by the MMSE receiver; the remaining $n-m+k$ diversity paths are combined to yield $v_{(k)}$ that is used to detect $\hat{x}_{(k)}$ based on MAP decision. Hence $v_{(k)}$ can be expressed as the following, assuming ZF suppression:

$$\begin{aligned} v_{(k)} &= \mathbf{w}_{(k)}\mathbf{y}_{(k-1)} = \mathbf{w}_{(k)}(\mathbf{y} - \sum_{i=1}^{k-1}\hat{x}_{(i)}\mathbf{h}_{(i)}) \\ &= x_{(k)} + \mathbf{w}_{(k)}\mathbf{z} + \mathbf{w}_{(k)}\sum_{i=1}^{k-1}(x_{(i)} - \hat{x}_{(i)})\mathbf{h}_{(i)} \end{aligned} \tag{3.1}$$

We refer $\mathbf{u}_{(j)} \triangleq (x_{(j)} - \hat{x}_{(j)})\mathbf{h}_{(j)}$ to as the *residue* of SIC stage j. After all m substreams are detected and subtracted from $\mathbf{y}$, we obtain

$$\mathbf{y}_{(m)} = \mathbf{y} - \sum_{i=1}^{m}\hat{x}_{(i)}\mathbf{h}_{(i)} = \mathbf{z} + \sum_{i=1}^{m}\mathbf{u}_{(i)}, \tag{3.2}$$

www.manaraa.com

which consists of noise and residues. If we add the contribution of the first detected substream, $\hat{x}_{(1)}\mathbf{h}_{(1)}$, to $\mathbf{y}_{(m)}$, we obtain

$$\mathbf{y}_{(1')} = \mathbf{y}_{(m)} + \hat{x}_{(1)}\mathbf{h}_{(1)} = x_{(1)}\mathbf{h}_{(1)} + \mathbf{z} + \sum_{i=2}^{m}\mathbf{u}_{(i)}, \tag{3.3}$$

which contains the contribution from substream (1) plus noise and residues of substreams (2) to (m). It should be noted that the residue of substream (1) is removed. Let

$$\mathbf{W}_{(1')} = (\mathbf{h}_{(1)}^H\mathbf{h}_{(1)})^{-1}\mathbf{h}_{(1)}^H = \frac{\mathbf{h}_{(1)}^H}{\|\mathbf{h}_{(1)}\|^2} \tag{3.4}$$

be the suppression vector (ZF type) for $\mathbf{h}_{(1)}$. Applying $\mathbf{w}_{(1')}$ to $\mathbf{y}_{(1')}$ yields

$$\begin{aligned} v_{(1')} &= \mathbf{W}_{(1')}\mathbf{y}_{(1')} = \frac{\mathbf{h}_{(1)}^H\mathbf{h}_{(1)}}{\|\mathbf{h}_{(1)}\|^2}x_{(1)} + \frac{\mathbf{h}_{(1)}^H}{\|\mathbf{h}_{(1)}\|^2}(\mathbf{z}+\sum_{i=2}^{m}\mathbf{u}_{(i)}) \\ &= x_{(1)} + \frac{\mathbf{h}_{(1)}^H}{\|\mathbf{h}_{(1)}\|^2}(\mathbf{z}+\sum_{i=2}^{m}\mathbf{u}_{(i)}) \end{aligned} \tag{3.5}$$

Assuming there is no residue ($\mathbf{u}_{(i)} = 0$), $v_{(1')}$ is of diversity order n, which is the same as the (m)th (last) substream in SIC. Therefore, compared to $v_{(1)}$ which has diversity order only $n-m+1$ [17], $\hat{x}_{(1')}$ will be more reliable than $\hat{x}_{(1)}$; in other words,

$$P_{e,(1)} > P_{e,(1')}. \tag{3.6}$$

After substream (1) is re-processed and re-detected, we can substract the new estimation from received vector $\mathbf{y}_{(\mathbf{1'})}$ and add the SIC-detected signal of substream (2), $\hat{x}_{(2)}\mathbf{h}_{(2)}$, to obtain

$$\begin{aligned} \mathbf{y}_{(2')} &= \mathbf{y}_{(1')} - \hat{x}_{(1')}\mathbf{h}_{(1)} + \hat{x}_{(2)}\mathbf{h}_{(2)} \\ &= x_{(2)}\mathbf{h}_{(2)} + (x_{(1)} - \hat{x}_{(1')})\mathbf{h}_{(1)} + \sum_{i=3}^{m}\mathbf{u}_{(i)} + \mathbf{z} \\ &= x_{(2)}\mathbf{h}_{(2)} + \mathbf{u}_{(1')} + \sum_{i=3}^{m}\mathbf{u}_{(i)} + \mathbf{z} \end{aligned} \tag{3.7}$$

where $\mathbf{u}_{(1')} = (x_{(1)} - \hat{x}_{(1')})\mathbf{h}_{(1)}$ is the new residue. Applying the suppression vector $\mathbf{W}_{(2')} = \frac{\mathbf{h}_{(2)}^H}{\|\mathbf{h}_{(2)}\|^2}$ to $\mathbf{y}_{(2')}$, we can obtain $v_{(2')}$ to estimate $\hat{x}_{(2')}$. Similar to $v_{(1')}$, the diversity order of $v_{(2')}$ is n assuming zero residues, thereby providing a more accurate estimation of $x_{(2)}$. Similar procedure can be applied to all remaining substreams. In what follows, this will be referred to as iterative post-SIC (IP-SIC) processing.

المنارة للاستشارات
www.manaraa.com

In general, the decision statistic $v_{(k')}$ can be expressed as

$$v_{(k')} = \mathbf{W}_{(k')}\mathbf{y}_{(k')} = x_{(k)} + \frac{\mathbf{h}_{(k)}^H}{\|\mathbf{h}_{(k)}\|^2}(\mathbf{z} + \sum_{i=1}^{k-1}\mathbf{u}_{(i')} + \sum_{i=k+1}^{m}\mathbf{u}_{(i)}) \tag{3.8}$$

The diversity order of all substreams can be improved from $n-m+k$ to n, assuming there is no residue; hence the increase in diversity order is more significant for smaller k, *i.e.* substreams detected earlier.

Although extra residues $\sum_{i=k+1}^{m}\mathbf{u}_{(i)}$ are present in $v_{(k')}$ as compared to $v_{(k)}$ in (3.1), the increase in diversity order can further reduce the error probability – the fading effects mitigated by increase in diversity order will result in a reliability improvement more significant than the performance degradation caused by these extra residues. This is confirmed by the simulation results presented later in this section and in Chapter 5.

The last several substreams detected in SIC will not get diversity increase substantially through IP-SIC, but their reliability is still improved because of the reduced residue of earlier detected substreams. The error probabilities of the substreams detected earlier are reduced by IP-SIC, therefore the variances of their residues are shrunk:

$$\text{var}(\mathbf{u}_{(k')}) < \text{var}(\mathbf{u}_{(k)}). \tag{3.9}$$

The reduced residue variances of earlier detected substreams will lead to larger SINR for the later detected substreams and consequently lower error probabilities. Specifically for the last detected substream (m'), we have

$$v_{(m)} = x_{(m)} + \frac{\mathbf{h}_{(m)}^H}{\|\mathbf{h}_{(m)}\|^2}(\mathbf{z} + \sum_{i=1}^{m-1}\mathbf{u}_{(i)}) \tag{3.10}$$

and

$$v_{(m')} = x_{(m)} + \frac{\mathbf{h}_{(m)}^H}{\|\mathbf{h}_{(m)}\|^2}(\mathbf{z} + \sum_{i=1}^{m-1}\mathbf{u}_{(i')}), \tag{3.11}$$

the diversity orders of $v_{(m)}$ and $v_{(m')}$ are the same; however since

$$\text{var}(\sum_{i=1}^{m-1}\mathbf{u}_{(i')}) < \text{var}(\sum_{i=1}^{m-1}\mathbf{u}_{(i)}), \tag{3.12}$$

the estimation $\hat{x}_{(m')}$ will be more accurate than $\hat{x}_{(m)}$.

www.manaraa.com

Because of the residue variance reductions, running IP-SIC more than one round may further reduce the error probability; but additional gain is small from running IP-SIC more than one round.

Since the residue of a specific substream will affect the reliability of remaining substreams, the optimal ordering of IP-SIC to minimize error propagation is: at stage k, choose the substream with largest residue variance reduction, *i.e.* substream i with the largest $\mathrm{var}(\mathbf{u}_{(i)}) - \mathrm{var}(\mathbf{u}_{(i')})$ value from the remaining $m - k + 1$ substreams, to process. However, because the residue variances are difficult to calculate and the benefit of optimal ordering for IP-SIC is small, we can follow the original order of SIC in applying IP-SIC to avoid additional complexity.

MMSE type interference suppression can also be applied to the IP-SIC with suppression vector

$$\mathbf{W}_{(k')} = (\mathbf{h}_{(k)}^H \mathbf{h}_{(k)} + \frac{N_0}{E_s})^{-1} \mathbf{h}_{(k)}^H = \frac{\mathbf{h}_{(k)}^H}{\|\mathbf{h}_{(k)}\|^2 + \frac{N_0}{E_s}} \tag{3.13}$$

Fig. 3.1 through 3.3 show the substream error counts of conventional SIC and IP-SIC for $m = n = 8$ V-BLAST systems, BPSK modulation, obtained at $E_s/N_0 = 0$ dB per transmit antenna and 10^5 transmission trials with MATLAB simulation. Fig. 3.1 is for no ordering ZF receiver, Fig. 3.2 is for LLR ordering ZF receiver and Fig. 3.3 is for LLR ordering MMSE receiver. It can be seen that compared to SIC, the substream error counts are reduced through IP-SIC. The substreams that are detected earlier have a more error count reduction compared to the substreams that are detected later. This is mainly due to the increase in diversity order. The error count reductions of the last several substreams are mainly due to a decrease in residue.

In short, the IP-SIC algorithm can be summarized as

1. Perform the conventional SIC decoding on all substreams and save the ordering list.

2. For $k = (1)$ to (m)

   (a) add the detected signal $\hat{x}_{(k)}\mathbf{h}_{(k)}$ to modified received vector $\mathbf{y}_{(m)}$ (if $k = 1$) or $\mathbf{y}_{(\{k-1\}')}$ (if $k > 1$), yielding $\mathbf{y}_{(k')}$

المنارة للاستشارات
www.manaraa.com

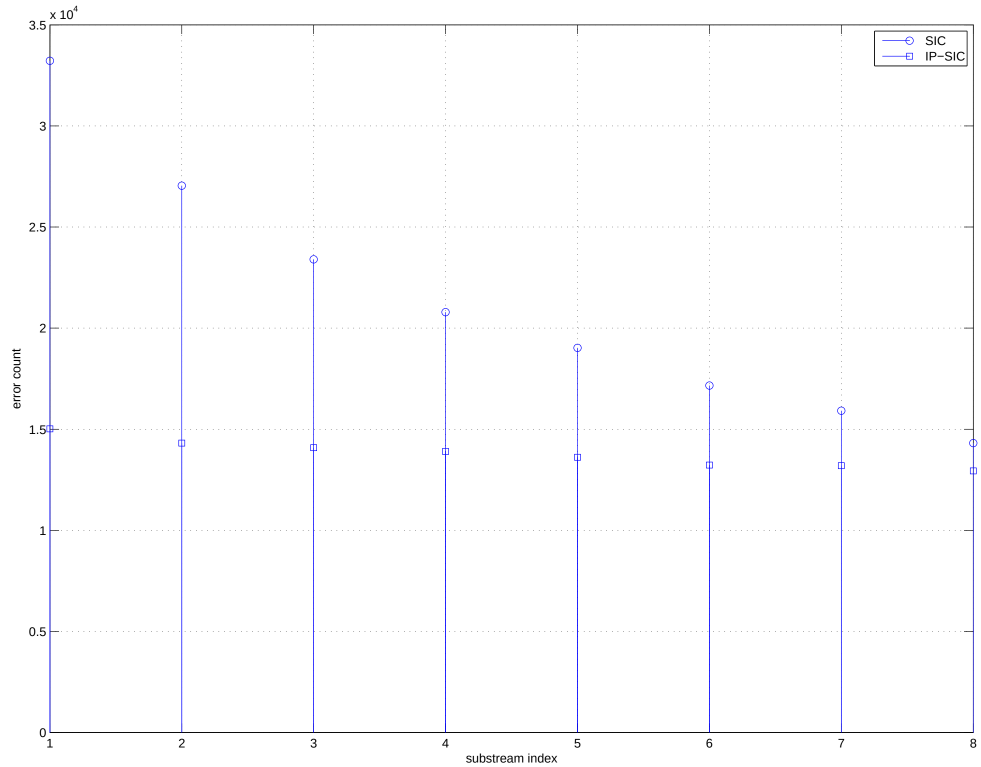

Figure 3.1 Substream error counts, $E_s/N_0 = 0$ dB, $m = n = 8$, BPSK, no ordering ZF, number of transmission trials $=10^5$.

(b) form suppression (diversity combining) vector $\mathbf{W}_{(k')} = \frac{\mathbf{h}_{(k)}^H}{\|\mathbf{h}_{(k)}\|^2}$ (ZF) or $\mathbf{W}_{(k')} = \frac{\mathbf{h}_{(k)}^H}{\|\mathbf{h}_{(k)}\|^2 + \frac{N_0}{E_s}}$ (MMSE)

(c) apply vector $\mathbf{W}_{(k')}$ to $\mathbf{y}_{(k')}$, make MAP decision $\hat{x}_{(k')}$ based on $v_{(k')} = \mathbf{W}_{(k')}\mathbf{y}_{(k')}$

(d) subtract the new signal estimation $\hat{x}_{(k')}\mathbf{h}_{(k)}$ from $\mathbf{y}_{(k')}$

3. Additional rounds of IP-SIC can be applied for further improvement of reliability.

3.2 Performance Analysis

Since the performance analysis of IP-SIC is quite complicated if ordering is employed, we will only consider the unordered case with ZF type suppression.

www.manaraa.com

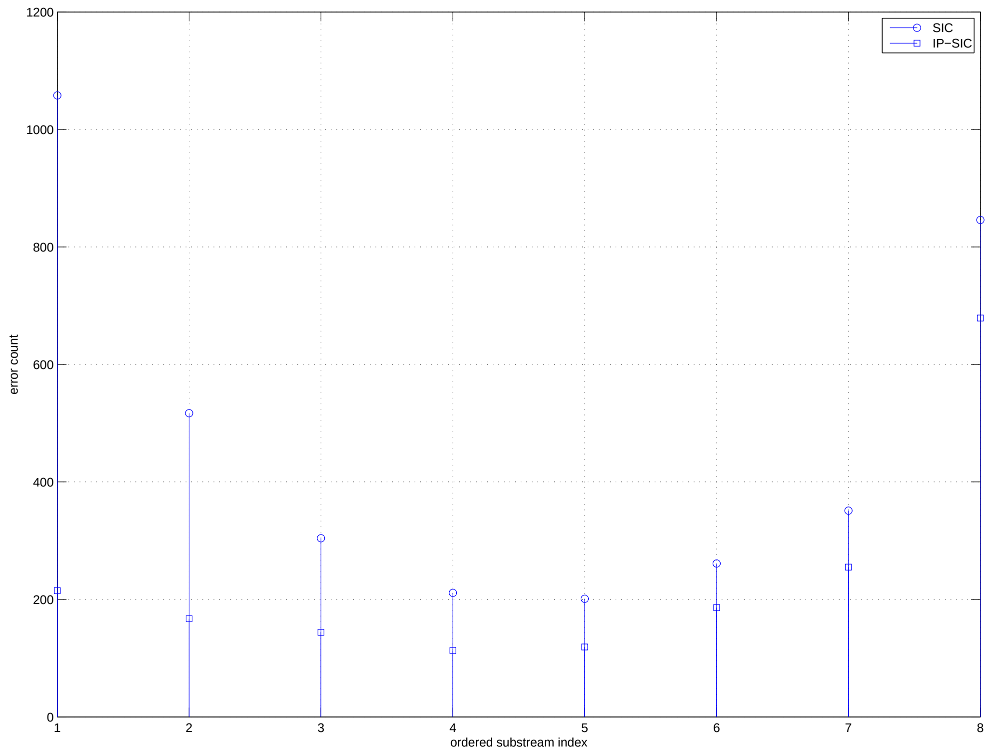

Figure 3.2 Substream error counts, $E_s/N_0 = 0$ dB, $m = n = 8$, BPSK, LLR ordering ZF, number of transmission trials $=10^5$.

3.2.1 Performance Analysis of SIC

We define the event

$$R_{i,\{1,2,\cdots,k-1\}} \triangleq \{i \text{ errors occurred in detecting substreams } 1,2,\cdots,k-1\} \tag{3.14}$$

By total probability theorem, the error probability of the substream k can be expressed as

$$P_{e,k} = \sum_{i=0}^{k-1} \Pr(\hat{x}_k \neq x_k | R_{i,\{1,2,\cdots,k-1\}}) \Pr(R_{i,\{1,2,\cdots,k-1\}}) \tag{3.15}$$

Therefore we are interested in finding $\Pr(\hat{x}_k \neq x_k | R_{i,\{1,2,\cdots,k-1\}})$ and $\Pr(R_{i,\{1,2,\cdots,k-1\}})$.

From Chapter 2 we know that the modified received vector $\mathbf{y}_k$ after the kth substream has been cancelled can be expressed as

$$\mathbf{y}_k = \sum_{j=k+1}^{m} x_j \mathbf{h}_j + \mathbf{n} + \sum_{j=1}^{k} \mathbf{u}_j, \tag{3.16}$$

المنارة للاستشارات
www.manaraa.com

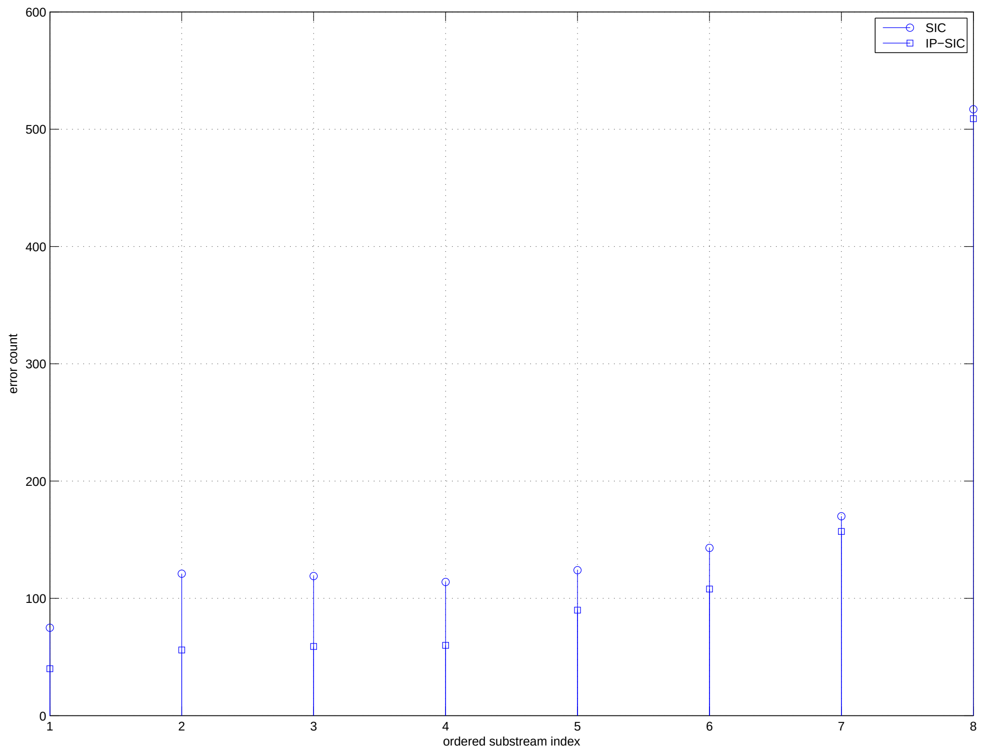

Figure 3.3 Substream error counts, $E_s/N_0 = 0$ dB, $m = n = 8$, BPSK, LLR ordering MMSE, number of transmission trials $=10^5$.

which contains uncancelled substreams, noise and residues from cancelled substreams. Let

$$\mathbf{n}_k \triangleq \mathbf{n} + \sum_{j=1}^{k} \mathbf{u}_j \tag{3.17}$$

be the sum of noise and residue. For BPSK modulation, the discrete random variable $x_k - \hat{x}_k$ can take values 0 with probability $1 - P_{e,k}$, or $\pm 2\sqrt{E_s}$ with probability $P_{e,k}/2$ respectively, assuming $\pm\sqrt{E_s}$ are transmitted with equal probability. Assuming i errors occurred on substream 1 through $k-1$, the mean $\mathbb{E}[\mathbf{n}_k | R_{i,\{1,2,\cdots,k-1\}}]$ will be zero and the covariance matrix of $\mathbf{n}_k$ given $R_{i,\{1,2,\cdots,k-1\}}$ is

$$\mathbf{C}_{\mathbf{n}_k | R_{i,\{1,2,\cdots,k-1\}}} = (N_0 + 4iE_s)\mathbf{I} \tag{3.18}$$

where identity matrix $\mathbf{I}$ is of size $n \times n$. The entries of $\mathbf{n_k}$ given $R_{i,\{1,2,\cdots,k-1\}}$ can be approximated as white Gaussian distributed [22]; hence we can utilize the error probability formula

www.manaraa.com

for Rayleigh fading BPSK signal with diversity order d and SNR γ [23]:

$$P_e(d,\gamma) = [\frac{1}{2}(1-\sqrt{\frac{\gamma}{1+\gamma}})]^d \sum_{t=0}^{d-1} \begin{pmatrix} d+t-1 \\ t \end{pmatrix} [\frac{1}{2}(1+\sqrt{\frac{\gamma}{1+\gamma}})]^t \tag{3.19}$$

Substituting $d = n-m+k$ and $\gamma = \frac{E_s}{N_0+4iE_s}$ into (3.19) yields

$$\Pr(\hat{x}_k \neq x_k | R_{i,\{1,2,\cdots,k-1\}}) = P_e(n-m+k, \frac{E_s}{N_0+4iE_s}) \tag{3.20}$$

On the other hand, we have

$$\begin{aligned} \Pr(R_{i,\{1,2,\cdots,k-1\}}) &= \Pr(\hat{x}_{k-1} \neq x_{k-1} | R_{i-1,\{1,2,\cdots,k-2\}}) \Pr(R_{i-1,\{1,2,\cdots,k-2\}}) + \\ &\quad \Pr(\hat{x}_{k-1} = x_{k-1} | R_{i,\{1,2,\cdots,k-2\}}) \Pr(R_{i,\{1,2,\cdots,k-2\}}). \end{aligned} \tag{3.21}$$

All entries in Table 3.1 can be calculated recursively using (3.21) with initial values

$$\begin{aligned} \Pr(R_{0,\{1\}}) &= 1 - P_e(n-m+1, E_s/N_0) \\ \Pr(R_{1,\{1\}}) &= P_e(n-m+1, E_s/N_0) \end{aligned} \tag{3.22}$$

and these entries can be applied into (3.15) to compute $P_{e,k}$ for all substreams.

Table 3.1 Residue probabilities for computing substream error probabilities in SIC

$\Pr(R_{0,\{1\}})$	$\Pr(R_{0,\{1,2\}})$	$\cdots$	$\Pr(R_{0,\{1,\cdots,k\}})$
$\Pr(R_{1,\{1\}})$	$\Pr(R_{1,\{1,2\}})$	$\cdots$	$\vdots$
	$\Pr(R_{2,\{1,2\}})$	$\cdots$	$\vdots$
		$\cdots$	$\Pr(R_{k,\{1,\cdots,k\}})$

3.2.2 Performance Analysis of IP-SIC

For IP-SIC, we first consider the case of $m = 2$ and $n \geq 2$. If we let $P_{e,k}$ be the bie error probability of the substream k, the bit error probabilities of substream 1 and 2, $P_{e,1'}$ and $P_{e,2'}$ with IP-SIC are given by

$$\begin{aligned} P_{e,1'} &= P_{e,2} P_e(n, \frac{E_s}{N_0+4E_s}) + (1-P_{e,2}) P_e(n, \frac{E_s}{N_0}) \\ P_{e,2'} &= P_{e,1'} P_e(n, \frac{E_s}{N_0+4E_s}) + (1-P_{e,1'}) P_e(n, \frac{E_s}{N_0}). \end{aligned} \tag{3.23}$$

المنارة للاستشارات
www.manaraa.com

Similarly, the bit error probabilities with 2 rounds IP-SIC are given by

$$\begin{aligned} P_{e,1''} &= P_{e,2'}P_e(n, \frac{E_s}{N_0+4E_s}) + (1-P_{e,2'})P_e(n, \frac{E_s}{N_0}) \\ P_{e,2''} &= P_{e,1''}P_e(n, \frac{E_s}{N_0+4E_s}) + (1-P_{e,1''})P_e(n, \frac{E_s}{N_0}) \end{aligned} \quad (3.24)$$

In general, for $m > 2$ and $n \geq m$, we have

$$P_{e,k'} = \sum_{i=0}^{m-1} P_e(n, \frac{E_s}{N_0+4iE_s}) \Pr(R_{i,\{1',\cdots,(k-1)',k+1,\cdots,m\}}) \quad (3.25)$$

and

$$P_{e,k''} = \sum_{i=0}^{m-1} P_e(n, \frac{E_s}{N_0+4iE_s}) \Pr(R_{i,\{1'',\cdots,(k-1)'',(k+1)',\cdots,m'\}}), \quad (3.26)$$

etc. For $m \geq 3$, general expressions for $\Pr(R_{i,\{1',\cdots,(k-1)',k+1,\cdots,m\}})$ and $\Pr(R_{i,\{1'',\cdots,(k-1)'',(k+1)',\cdots,m'\}})$ are difficult to derive and the complexities to compute them grows exponentially with increasing m. Derivation of $\Pr(R_{i,\{2,3\}})$ for $i = 0, 1, 2$, which will be used in finding $P_{e,1'}$ when $m = 3$, is provided in the appendix.

3.3 Complexity Analysis

3.3.1 Complexity of SIC

The complexity of SIC without ordering lies in constructing and applying suppression vectors, MAP detection and cancellation of estimated signal. Among these the MAP detection is modulation dependent (not related to n and m) and will not be included in the following analysis.

Let $f_\times$ and f_+ denotes the number of complex multiplications and complex additions for a specific step respectively. The number of operations needed to obtain the suppression matrix $\mathbf{W} = (\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H$ for ZF case is approximately $f_\times = f_+ = 2(m-k+1)^2n + 2(m-k+1)^3/3$ while for MMSE it is approximately $f_\times = 2(m-k+1)^2n + 2(m-k+1)^3/3$, $f_+ = 2(m-k+1)^2n + 2(m-k+1)^3/3 + (m-k+1)$, if the Gaussian elimination algorithm is applied for the matrix inversion. Applying suppression vector on received vector and estimated signal removal would require $f_\times = f_+ = 2n$ for both ZF and MMSE. Hence the complexity for SIC is on the order of $\mathcal{O}(m^2n)$.

المنارة للاستشارات
www.manaraa.com

3.3.2 Complexity of Ordering

For ZF or MMSE suppression with SNR ordering, the operations needed would be to compute $m-k+1$ substreams' $\|\mathbf{w}_i\|^2$ values which takes $f_\times = f_+ = (m-k+1)n$ and select the substream with the smallest $\|\mathbf{w}_i\|^2$ to process at each stage of SIC; therfore the complexity of SNR ordering for each substream is on the order of $\mathcal{O}(mn)$. For MMSE SINR ordering, approximately extra $f_\times = f_+ = (m-k+1)n$ operations are needed than SNR ordering at each stage of SIC, to calculate $|\mathbf{w}_i\mathbf{h}_j|^2$, but the complexity order stays the same as in the SNR ordering, $\mathcal{O}(mn)$.

For ZF LLR ordering, at each stage of SIC, operations needed are from computing

- $\|\mathbf{w}_i\|^2$ values, $f_\times = f_+ = (m-k+1)n$;
- v_i values, $f_\times = f_+ = (m-k+1)n$;
- $|v_i - s_t|^2$ values, $f_\times = f_+ = (m-k+1)M$;
- $\sum_{t=1}^{M} \exp(-\beta_{i,t})$ values, $f_+ = (m-k+1)M$ and $(m-k+1)M$ exponential operations;
- browse through $m-k+1$ LLR values to pick the maximum;
- plus other miscellaneous additions and multiplications operations whose number is some multiples of $(m-k+1)$.

Therefore the complexity order would be $\mathcal{O}(m\max(M,n))$. For MMSE LLR ordering, derivation is similar to ZF case but computing $|\mathbf{w}_i\mathbf{h}_j|^2$ would require additional $f_\times = f_+ = (m-k+1)n$ and the complexity order is again $\mathcal{O}(m\max(M,n))$.

3.3.3 Complexity of IP-SIC

Applying IP-SIC on each substream requires

- adding estimated signal of interested substream back to the received vector, $f_\times = f_+ = n$;
- deriving $\mathbf{W}_{(i')} = \frac{\mathbf{h}_{(i)}^H}{\|\mathbf{h}_{(i)}\|^2}$, $f_\times = 2n$, $f_+ = n$ (ZF) or $\mathbf{W}_{(i')} = \frac{\mathbf{h}_{(i)}^H}{\|\mathbf{h}_{(i)}\|^2 + \frac{N_0}{E_s}}$, $f_\times = 2n$, $f_+ = n+1$ (MMSE);

www.manaraa.com

- applying vector $\mathbf{W}_{(i')}$ on received vector, $f_\times = f_+ = n$;
- removing the new signal estimation from received vector, $f_\times = f_+ = n$.

Therefore the complexity order of IP-SIC processing for each substream is $\mathcal{O}(n)$, which is much smaller than that of SIC with ordering. Appending IP-SIC to SIC with any ordering scheme will not change the overall complexity order, which is dominated by the complexity order of SIC, $\mathcal{O}(m^2 n)$.

3.4 IP-SIC with Partial Residue Supression

After SIC decoding is complete in V-BLAST systems, the total degrees of freedom (DoF) for interference suppression and diversity combining is m. Therefore for IP-SIC, we could utilize q DoF for interference suppression and $m-q$ DoF to combine $n-q$ diversity paths. A trade-off exists in interference suppression and diversity combining: more DoF used for interference suppression causes less DoF available for diversity combining, and vice versa.

Residue is one of the major factors that determines the performance of V-BLAST system with SIC decoding because the energy of a non-zero residue can often be higher than the energy of the original symbol. For example, in BPSK signaling, if a substream is detected incorrectly, the energy of its residue will be $4E_s$, which is four times the energy of the original substream. Hence the interference introduced by residue can degrade the detection reliability of the remaining substreams significantly; it is always desired to cancel as many residues as possible. On the other hand, the performance improvement provided by increasing diversity order saturates: beyond certain diversity order, increasing diversity order furthermore yields little performance enhancement. Therefore, in IP-SIC, it is possible to use some DoF to suppress some (preferably the strongest) residues at the cost of lowered diversity order, to further improve the system performance.

To suppress q ($q \leq m-1$) residues in IP-SIC processing (ordered) substream (k'), we need to include $q+1$ columns of channel gains in a matrix $\mathbf{H}_{(k')}$; the first column of $\mathbf{H}_{(k')}$ will be $\mathbf{h}_{(k)}$ and the remaining q columns are going to be the channel vectors corresponding to the q residues. Then we can form the $(1+q) \times n$ suppression matrix $\mathbf{W}_{(k')} = (\mathbf{H}^H_{(k')}\mathbf{H}_{(k')})^{-1}\mathbf{H}^H_{k'}$

www.manaraa.com

(ZF) or $\mathbf{W}_{(k')} = (\mathbf{H}_{(k')}^H \mathbf{H}_{(k')} + \frac{N_0}{E_s}\mathbf{I})^{-1}\mathbf{H}_{(k')}^H$ (MMSE); applying the first row of $\mathbf{W}_{(k')}$ on the modified received vector will suppress the q residues and $v_{(k')}$ is obtained by combining $n-q$ diversity paths. Since the q residues are suppressed at the expense of diversity order lowered by q, the value q should be chosen for every (k') in order to optimize the substream estimation reliability.

As an example, consider a system with all substreams processed with no ordering ZF SIC algorithm. We see from Fig. 3.1 that when IP-SIC processing substream k', on average, the strongest (largest variance) q residues would be $\mathbf{u}_{k+1}, \mathbf{u}_{k+2}, \cdots, \mathbf{u}_{k+q}$ for $k+q \leq m$, $\mathbf{u}_{k+1}, \cdots,$ $\mathbf{u}_m, \mathbf{u}_{1'}, \cdots, \mathbf{u}_{\{q-m+k\}'}$ for $k+q > m, k \neq m$ and $\mathbf{u}_{1'}, \mathbf{u}_{2'}, \cdots, \mathbf{u}_{q'}$ for $k = m$. Hence we can form

$$\mathbf{H}_{k'} = \begin{cases} [\mathbf{h}_k, \mathbf{h}_{k+1}, \cdots, \mathbf{h}_{k+q}], & k+q \leq m \\ [\mathbf{h}_k, \mathbf{h}_{k+1}, \cdots, \mathbf{h}_m, \mathbf{h}_1, \cdots, \mathbf{h}_{q-m+k}], & k+q > m, k \neq m \\ [\mathbf{h}_m, \mathbf{h}_1, \cdots, \mathbf{h}_q], & k = m, \end{cases} \tag{3.27}$$

and let $\mathbf{W}_{k'} = (\mathbf{H}_{k'}^H \mathbf{H}_{k'})^{-1}\mathbf{H}_{k'}^H$ be the suppression matrix; applying the first row of $\mathbf{W}_{k'}$ on the modified received vector will suppress the q residues and yield the decision statistic $v_{k'}$.

For the cases other than ZF with no ordering, we can still apply this residue suppression idea in IP-SIC processing, but to find the indices of the q strongest residues is another challenge. From Fig. 3.2 and 3.3 we see that the error counts for SIC and IP-SIC are not monotonically decreasing with ordered substream index; the q strongest residues when IP-SIC processing substream (i') may not come from (circularly) next q substreams as in the no ordering ZF case. Hence it would be difficult to analytically find the strongest residues to suppress and consequently yield the optimal results.

3.5 Joint ML/SIC Combined Decoding and IP-SIC

In SIC decoding algorithm, due to the error propagation effect, the detection reliability of the first substream to process will limit the overall performance of V-BLAST system: in SIC the kth processed substream bears an diversity order of $n-m+k$ assuming there is no residue, but when residue is taken into account, the actual diversity order will stay at $n-m+1$ for

المنارة للاستشارات
www.manaraa.com

all substreams [22, 24]. In other words, the actual diversity order for all substreams and the average reliability performance will be the same, $n - m + 1$. The first substream to process in SIC is the bottleneck of the overall performance and its diversity order will determine that of all remaining substreams.

To enhance the overall performance, extending the diversity order of the first substream beyond $n-m+1$ is critical. One good solution will be carrying out joint ML detection on the first g substreams [25], instead of using SIC. In this manner the first g substreams will have diversity order incresed to $n - m + g$, as well as the remaining $m - g$ substreams.

To joint ML detecting g substreams, removing the remaining $m - g$ substreams utilizing group interference suppression technique [26] is necessary. At the first stage of SIC, we can select g out of m substreams with largest SNR or LLR, according to criteria described in Chapter 2 about ordering in ZF SIC. The indices with these ordered substreams would be $(1), (2), \cdots, (g)$. Let

$$\bar{\mathbf{H}} = [\mathbf{h}_{(1)}, \mathbf{h}_{(2)}, \cdots, \mathbf{h}_{(g)}] \tag{3.28}$$

and $\tilde{\mathbf{H}}$ consist of columns of $\mathbf{H}$ that are not contained in $\bar{\mathbf{H}}$. If we denote the null space of $\tilde{\mathbf{H}}$ as $\tilde{\mathcal{H}}$, then the dimension of $\tilde{\mathcal{H}}$ and the rank of $\tilde{\mathbf{H}}$ have the relationship

$$\dim(\tilde{\mathcal{H}}) + \text{rank}(\tilde{\mathbf{H}}) = n, \tag{3.29}$$

it follows that $\dim(\tilde{\mathcal{H}}) \geq n - m + g$ since $\text{rank}(\tilde{\mathbf{H}}) \leq m - g$. Hence we can find a set of orthonormal vectors $\{\mathbf{l}_1, \mathbf{l}_2, \cdots, \mathbf{l}_{n-m+g}\}$; each vector is of size $1 \times n$. The matrix

$$\mathbf{L} = [\mathbf{l}_1, \mathbf{l}_2, \cdots, \mathbf{l}_{n-m+g}]^T \tag{3.30}$$

would be the suppression matrix of $\tilde{\mathbf{H}}$ satisfying $\mathbf{L}\tilde{\mathbf{H}} = \underline{0}$. Applying this suppression matrix to the received vector will produce

$$\mathbf{Ly} = \mathbf{LHx} + \mathbf{Lz} = \mathbf{L}(\sum_{i=1}^{g} x_{(i)}\mathbf{h}_{(i)}) + \mathbf{Lz} \tag{3.31}$$

Therefore to jointly ML detect $x_{(1)}, x_{(2)}, \cdots, x_{(g)}$ assuming each symbol is transmitted with equal probability, we substitute M^g different vectors $\hat{\mathbf{x}}_g = [\hat{x}_{(1)}, \hat{x}_{(2)}, \cdots, \hat{x}_{(g)}]^T, \hat{x}_{(i)} \in \mathcal{S}$ into $\mathbf{L}\bar{\mathbf{H}}\hat{\mathbf{x}}_g$ and decide the one making $\mathbf{L}\bar{\mathbf{H}}\hat{\mathbf{x}}_g$ closest to $\mathbf{Ly}$ as the transmitted symbol vector.

المنارة للاستشارات
www.manaraa.com

After jointly ML detecting $x_{(1)}, x_{(2)}, \cdots, x_{(g)}$, we modify the received vector as

$$\mathbf{y}_{(g)} = \mathbf{y} - \sum_{i=1}^{g} \hat{x}_{(i)} \mathbf{h}_{(i)} \tag{3.32}$$

and let $\mathbf{H}_{(g)} = \tilde{\mathbf{H}}$. The remaining $m - g$ substreams will be processed by conventional ZF or MMSE SIC algorithm. The IP-SIC procedure can be applied after all stages of SIC have been completed, by adding the joint ML/SIC detected symbol times its channel vector one at a time and re-detect the symbol, to yield an even better reliability performance.

It is clear that the computational complexity of joint ML detection grows exponentially with increasing M or g; therefore it would be practical to apply this method when the modulation constellation size is not large and only on a small portion of substreams.

المنارة للاستشارات

www.manaraa.com

CHAPTER 4. IP-SIC IN V-BLAST SYSTEMS COMBINED WITH STBC

4.1 Model Description of a STBC/V-BLAST Combined System

It has been mentioned in Chapter 1 that utilizing space-time block codes (STBC) at the transmitter side of a MIMO wireless communication system is a way of achieving transmitter diversity gain. In this chapter we consider a V-BLAST system that incorporates STBC at the transmitter end, to obtain transmitter diversity gain as well as multiplexing gain [18, 26]. We apply the proposed IP-SIC in the combined STBC/V-BLAST system.

We assume again the transmitter side is equipped with m antennas while n antennas are available at the receiver end. The m transmit antennas are partitioned into Ψ groups; the ψth group G_ψ consists of ρ_ψ antennas, thus we have $\sum_{i=1}^{\Psi} \rho_i = m$. Each group G_ψ is encoded by a space-time block encoder B_ψ; $1 \leq \psi \leq \Psi$. The encoder B_ψ produces a codeword consisting ρ_ψ symbols at each time τ and these ρ_ψ symbols are multiplexed to the ρ_ψ transmit antennas of this group and transmitted simutaneously. Altogether, we have a total of m symbols transmitted simutaneously from antennas $1, 2, \cdots, m$, with each antenna transmitting one symbol.

Let $x^{\tau}_{\{i,\psi\}}$ be the transmitted symbol from the ith transmit antenna in the ψth group G_ψ at time τ, and $h_{j,\{i,\psi\}}$ be the path gain from the ith transmit antenna in the ψth group G_ψ to the jth receive antenna. The signal obtained by the jth receive antenna at time τ can be represented as

$$y^{\tau}_{j} = \sum_{\psi=1}^{\Psi} \sum_{i=1}^{\rho_\psi} h_{j,\{i,\psi\}} x^{\tau}_{\{i,\psi\}} + z^{\tau}_{j} \tag{4.1}$$

where z^{τ}_{j} denotes the additive noise at the jth receive antenna at time τ. Same as in the V-BLAST model described in Chapter 2, the channel gains $h_{j,\{i,\psi\}}, 1 \leq i \leq \rho_\psi, 1 \leq \psi \leq \Psi, 1 \leq$

المنارة للاستشارات

www.manaraa.com

$j \le n$ are modeled as i.i.d. complex Gaussian random variables with zero mean and variance 1; the noise samples $z_j^\tau, 1 \le j \le n$ are assumed to be i.i.d. Gaussian distributed having mean equals zero and $N_0/2$ per dimension variance; transmitted symbols $x^\tau_{\{i,\psi\}}, 1 \le i \le \rho_\psi, 1 \le \psi \le \Psi$ are M-ary modulated. If we let

$$\mathbf{y}^\tau = [y_1^\tau, y_2^\tau, \cdots, y_n^\tau]^T, \tag{4.2}$$

$$\mathbf{z}^\tau = [z_1^\tau, z_2^\tau, \cdots, z_n^\tau]^T, \tag{4.3}$$

$$\mathbf{x}^\tau_\psi = [x^\tau_{\{1,\psi\}}, x^\tau_{\{2,\psi\}}, \cdots, x^\tau_{\{\rho_\psi,\psi\}}]^T \tag{4.4}$$

and

$$\mathbf{A}_\psi = [\mathbf{h}_{\{1,\psi\}}, \mathbf{h}_{\{2,\psi\}}, \cdots, \mathbf{h}_{\{\rho_\psi,\psi\}}] \tag{4.5}$$

where

$$\mathbf{h}_{\{i,\psi\}} = [h_{1,\{i,\psi\}}, h_{2,\{i,\psi\}}, \cdots, h_{n,\{i,\psi\}}]^T, \tag{4.6}$$

then the received vector $\mathbf{y}^\tau$ at time τ can be rewritten in matrix form as

$$\mathbf{y}^\tau = \sum_{\psi=1}^{\Psi} \mathbf{A}_\psi \mathbf{x}^\tau_\psi + \mathbf{z}^\tau = \mathbf{H}\mathbf{X}^\tau + \mathbf{z}^\tau \tag{4.7}$$

where

$$\mathbf{H} = [\mathbf{A}_1, \mathbf{A}_2, \cdots, \mathbf{A}_\Psi] \tag{4.8}$$

denotes the entire channel matrix that is perfectly known at the receiver side and

$$\mathbf{X}^\tau = [\mathbf{x}_1^\tau, \mathbf{x}_2^\tau, \cdots, \mathbf{x}_\Psi^\tau]^T \tag{4.9}$$

is the transmitted symbol vector at time τ. Fig. 4.1 is a block diagram illustrating this system.

4.2 Group SIC and STBC Decoding

In order to decode the transmitted codeword of a particular group, we need to suppress the signals came from all other groups using group interference suppression technique [26] as we did in section 3.5. In detecting the signal vector of the ψth group, we form matrix

$$\tilde{\mathbf{H}}_\psi = [\mathbf{A}_1, \cdots, \mathbf{A}_{\psi-1}, \mathbf{A}_{\psi+1}, \cdots, \mathbf{A}_\Psi] \tag{4.10}$$

المنارة للاستشارات
www.manaraa.com

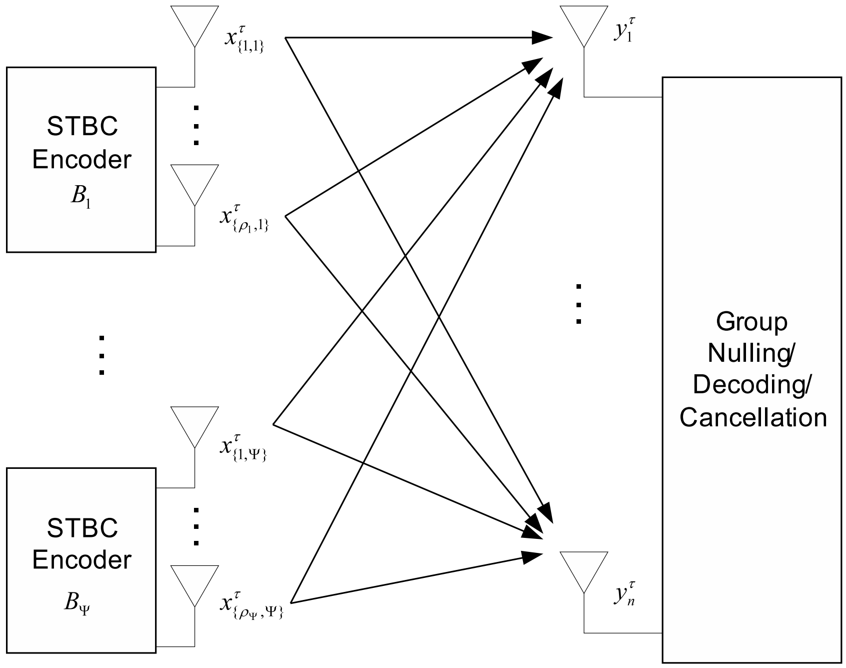

Figure 4.1 V-BLAST system combined with STBC

that contains channel gains of the interference groups. Assuming the number of receive antennas satisfy $n \geq m + 1 - \rho_\psi$, the suppression matrix $\mathbf{L}_\psi$ for $\tilde{\mathbf{H}}_\psi$ can be represented as

$$\mathbf{L}_\psi = [\mathbf{l}_1, \mathbf{l}_2, \cdots, \mathbf{l}_{n-m+\rho_\psi}]^T, \tag{4.11}$$

where $\mathbf{l}_i, i = 1, 2, \cdots, n - m + \rho_\psi$, each having size $1 \times n$, are the orthonormal vectors for $\mathbf{A}_j, j = 1, \cdots, \psi - 1, \psi + 1, \cdots, \Psi$. Hence we have $\mathbf{L}_\psi \tilde{\mathbf{H}}_\psi = \underline{0}$; that is, applying $\mathbf{L}_\psi$ on received vector $\mathbf{y}^\tau$ gives

$$\begin{aligned} \mathbf{L}_\psi \mathbf{y}^\tau &= \mathbf{L}_\psi \mathbf{H} \mathbf{X}^\tau + \mathbf{L}_\psi \mathbf{z}^\tau \\ &= \mathbf{L}_\psi \mathbf{A}_\psi \mathbf{x}^\tau_\psi + \mathbf{L}_\psi \mathbf{z}^\tau \end{aligned} \tag{4.12}$$

To decode the transmitted codeword from group G_ψ, we pass vector $\mathbf{L}_\psi \mathbf{y}^\tau$ to the STBC decoder D_ψ for group G_ψ. In the sequel, we simplify the problem by assuming the 2×2 Alamouti STBC [27] is employed in all Ψ transmit groups; thus every group ψ has $\rho_\psi = 2$

www.manaraa.com

transmit antennas and

$$\begin{aligned}\mathbf{x}_\psi^1 &= [x_{\{1,\psi\}}^1, x_{\{2,\psi\}}^1]^T \\ \mathbf{x}_\psi^2 &= [x_{\{1,\psi\}}^2, x_{\{2,\psi\}}^2]^T = [-(x_{\{2,\psi\}}^1)^*, (x_{\{1,\psi\}}^1)^*]^T \end{aligned} \tag{4.13}$$

where * denotes complex conjugate. Assume channel gains do not change during a block of 2 time slots, we obtain

$$\begin{aligned}\mathbf{L}_\psi \mathbf{y}^1 &= \mathbf{L}_\psi \mathbf{A}_\psi \mathbf{x}_\psi^1 + \mathbf{L}_\psi \mathbf{z}^1 = \mathbf{L}_\psi [\mathbf{h}_{\{1,\psi\}}\ \mathbf{h}_{\{2,\psi\}}]\mathbf{x}_\psi^1 + \mathbf{L}_\psi \mathbf{z}^1 \\ &= \begin{bmatrix} \mathbf{l}_1\mathbf{h}_{\{1,\psi\}} & \mathbf{l}_1\mathbf{h}_{\{2,\psi\}} \\ \vdots & \vdots \\ \mathbf{l}_{n-m+2}\mathbf{h}_{\{1,\psi\}} & \mathbf{l}_{n-m+2}\mathbf{h}_{\{2,\psi\}} \end{bmatrix} \begin{bmatrix} x_{\{1,\psi\}}^1 \\ x_{\{2,\psi\}}^1 \end{bmatrix} + \begin{bmatrix} \mathbf{l}_1\mathbf{z}^1 \\ \vdots \\ \mathbf{l}_{n-m+2}\mathbf{z}^1 \end{bmatrix} \end{aligned} \tag{4.14}$$

and similarly

$$\begin{aligned}\mathbf{L}_\psi \mathbf{y}^2 &= \mathbf{L}_\psi \mathbf{A}_\psi \mathbf{x}_\psi^2 + \mathbf{L}_\psi \mathbf{z}^2 = \mathbf{L}_\psi [\mathbf{h}_{\{1,\psi\}}\ \mathbf{h}_{\{2,\psi\}}]\mathbf{x}_\psi^2 + \mathbf{L}_\psi \mathbf{z}^2 \\ &= \begin{bmatrix} \mathbf{l}_1\mathbf{h}_{\{1,\psi\}} & \mathbf{l}_1\mathbf{h}_{\{2,\psi\}} \\ \vdots & \vdots \\ \mathbf{l}_{n-m+2}\mathbf{h}_{\{1,\psi\}} & \mathbf{l}_{n-m+2}\mathbf{h}_{\{2,\psi\}} \end{bmatrix} \begin{bmatrix} -(x_{\{2,\psi\}}^1)^* \\ (x_{\{1,\psi\}}^1)^* \end{bmatrix} + \begin{bmatrix} \mathbf{l}_1\mathbf{z}^2 \\ \vdots \\ \mathbf{l}_{n-m+2}\mathbf{z}^2 \end{bmatrix} \end{aligned} \tag{4.15}$$

To decode $x_{\{1,\psi\}}^1$ and $x_{\{2,\psi\}}^1$, we linearly combine $\mathbf{L}_\psi \mathbf{y}^1$ and $\mathbf{L}_\psi \mathbf{y}^2$ to get

$$\begin{aligned} v_{\{1,\psi\}} &= (\mathbf{L}_\psi \mathbf{h}_{\{1,\psi\}})^H \mathbf{L}_\psi \mathbf{y}^1 + \mathbf{L}_\psi \mathbf{h}_{\{2,\psi\}} (\mathbf{L}_\psi \mathbf{y}^2)^H \\ &= \alpha_\psi x_{\{1,\psi\}}^1 + \eta_{\{1,\psi\}} \end{aligned} \tag{4.16}$$

and

$$\begin{aligned} v_{\{2,\psi\}} &= (\mathbf{L}_\psi \mathbf{h}_{\{2,\psi\}})^H \mathbf{L}_\psi \mathbf{y}^1 - \mathbf{L}_\psi \mathbf{h}_{\{1,\psi\}} (\mathbf{L}_\psi \mathbf{y}^2)^H \\ &= \alpha_\psi x_{\{2,\psi\}}^1 + \eta_{\{2,\psi\}}, \end{aligned} \tag{4.17}$$

where

$$\alpha_\psi = \sum_{i=1}^{n-m+2} \sum_{j=1}^{2} |\mathbf{l}_i \mathbf{h}_{\{j,\psi\}}|^2 \tag{4.18}$$

and $\eta_{\{1,\psi\}} = (\mathbf{L}_\psi \mathbf{h}_{\{1,\psi\}})^H \mathbf{L}_\psi \mathbf{n}^1 + \mathbf{L}_\psi \mathbf{h}_{\{2,\psi\}} (\mathbf{L}_\psi \mathbf{n}^2)^H$, $\eta_{\{2,\psi\}} = (\mathbf{L}_\psi \mathbf{h}_{\{2,\psi\}})^H \mathbf{L}_\psi \mathbf{n}^1 - \mathbf{L}_\psi \mathbf{h}_{\{1,\psi\}} (\mathbf{L}_\psi \mathbf{n}^2)^H$. $\eta_{\{1,\psi\}}$ and $\eta_{\{2,\psi\}}$ are independent Gaussian random variables with zero mean and variance

www.manaraa.com

$\alpha_\psi N_0$. The transmitted symbols of group G_ψ can be detected using MAP decision principle

$$\hat{x}^1_{\{i,\psi\}} = \arg\max_{s_t} \Pr(x^1_{\{i,\psi\}} = s_t | v_{\{i,\psi\}}),\ i = 1, 2 \tag{4.19}$$

After detection, we follow the SIC procedure to cancel the signal of group G_ψ. The received vectors will be modified as

$$\begin{aligned} \mathbf{y}^1_1 &= \mathbf{y}^1 - \mathbf{A}_\psi[\hat{x}^1_{\{1,\psi\}}, \hat{x}^1_{\{2,\psi\}}]^T = \mathbf{y}^1 - \mathbf{A}_\psi \hat{\mathbf{x}}^1_\psi \\ \mathbf{y}^2_1 &= \mathbf{y}^2 - \mathbf{A}_\psi[-(\hat{x}^1_{\{2,\psi\}})^*, (\hat{x}^1_{\{1,\psi\}})^*]^T = \mathbf{y}^2 - \mathbf{A}_\psi \hat{\mathbf{x}}^2_\psi \end{aligned} \tag{4.20}$$

and the channel matrix $\mathbf{A}_\psi$ of group G_ψ will be removed from $\mathbf{H}$ to produce modified channel matrix $\mathbf{H}_1 = \tilde{\mathbf{H}}_\psi$.

The group SIC/decoding continues until all groups have been processed. Assuming $x^1_{\{1,\psi\}}$ and $x^1_{\{2,\psi\}}$ are detected correctly for all groups, then there is no error propagation and the diversity order for symbols in the group processed at the kth stage of SIC is $2(2k + n - m)$.

Like in the plain V-BLAST system, the error propagation of group SIC algorithm can be reduced through proper ordering [18]. It follows from (4.16) and (4.17) that the SNR for the ψth group is $\alpha^2_\psi \|\mathbf{x}^1_\psi\|^2 / N_0$; therefore in the case of equi-energy signaling, the SNR based ordering will be to process the group with the largest α_ψ at stage k of SIC. On the other hand, if every symbol in the modulation alphabet $\mathcal{S}$ is transmitted with equal probability, then the pairwise LLR of group ψ is

$$\begin{aligned} \beta_{\psi,t} &= \ln \frac{\Pr(\mathbf{v}_\psi | \mathbf{x}^1_\psi = \hat{\mathbf{x}}^1_\psi)}{\Pr(\mathbf{v}_\psi | \mathbf{x}^1_\psi = \mathbf{s}_t)} \\ &= \frac{\|\mathbf{v}_\psi - \alpha_\psi \mathbf{s}_t\|^2 - \|\mathbf{v}_\psi - \alpha_\psi \hat{\mathbf{x}}^1_\psi\|^2}{\alpha_\psi N_0}, \end{aligned} \tag{4.21}$$

where $\mathbf{v}_\psi = [v_{\{1,\psi\}}, v_{\{2,\psi\}}]^T$ and $\mathbf{s}_t = [s_{\{1,t\}}, s_{\{2,t\}}]^T, s_{\{1,t\}}, s_{\{2,t\}} \in \mathcal{S}, t = 1, 2, \cdots, M^2$. Since $\Pr(\mathbf{x}^1_\psi \neq \hat{\mathbf{x}}^1_\psi | \mathbf{v}_\psi) = 1 - 1/\sum_{t=1}^{M^2} \exp(-\beta_{\psi,t})$ decreases along with decreasing $\sum_{t=1}^{M^2} \exp(-\beta_{\psi,t})$, the LLR based ordering would select the group with the smallest $\sum_{t=1}^{M^2} \exp(-\beta_{\psi,t})$ to process at the kth stage of SIC.

المنارة للاستشارات
www.manaraa.com

4.3 IP-SIC and Group SIC/STBC Decoding

Upon all stages of SIC are finished, we can start applying the IP-SIC procedure on the modified received vector, which consists of noise and residues. Assuming the groups are SIC processed in the order of $(1), (2), \cdots, (\Psi)$, then the resulting modified received vector can be expressed as

$$\begin{aligned} \mathbf{y}^{\tau}_{(\Psi)} &= \mathbf{y}^{\tau} - \sum_{\psi=1}^{\Psi} \mathbf{A}_{(\psi)} \hat{\mathbf{x}}^{\tau}_{(\psi)} \\ &= \mathbf{z}^{\tau} + \sum_{\psi=1}^{\Psi} \mathbf{A}_{(\psi)} [\mathbf{x}^{\tau}_{(\psi)} - \hat{\mathbf{x}}^{\tau}_{(\psi)}]. \end{aligned} \tag{4.22}$$

To IP-SIC process the first group $G_{(1)}$ in the ordered list, we add its estimated signal back to the modified received vector to obtain

$$\begin{aligned} \mathbf{y}^{\tau}_{(1')} &= \mathbf{y}^{\tau}_{(\Psi)} + \mathbf{A}_{(1)} \hat{\mathbf{x}}^{\tau}_{(1)} \\ &= \mathbf{A}_{(1)} \mathbf{x}^{\tau}_{(1)} + \mathbf{z}^{\tau} + \sum_{i=2}^{\Psi} \mathbf{u}^{\tau}_{(i)} \end{aligned} \tag{4.23}$$

where $\mathbf{u}_{(i)}$ is the SIC cancellation residue of group $G_{(i)}$. To obtain $\hat{\mathbf{x}}^{\tau}_{(1')}$, the IP-SIC estimation of $\mathbf{x}^{\tau}_{(1)}$, there is no need to construct a suppression matrix to suppress interference groups; we simply do diversity combining and STBC decoding. Again assume the 2×2 Alamouti STBC is used for all Ψ groups, $\tau = 1, 2$, then

$$\begin{aligned} v_{\{1,1'\}} &= (\mathbf{I}\mathbf{h}_{\{1,1'\}})^H \mathbf{I}\mathbf{y}^1_{(1')} + \mathbf{I}\mathbf{h}_{\{2,1'\}} (\mathbf{I}\mathbf{y}^2_{(1')})^H \\ v_{\{2,1'\}} &= (\mathbf{I}\mathbf{h}_{\{2,1'\}})^H \mathbf{I}\mathbf{y}^1_{(1')} - \mathbf{I}\mathbf{h}_{\{1,1'\}} (\mathbf{I}\mathbf{y}^2_{(1')})^H \end{aligned} \tag{4.24}$$

where is identity matrix $\mathbf{I}$ is of size $n \times n$. Then, we can get $\hat{\mathbf{x}}^{\tau}_{(1')}$ by using the MAP decision principle based on $\mathbf{v}_{(1')} = [v_{\{1,1'\}}, v_{\{2,1'\}}]^T$. The new estimated signal of group $G_{(1)}$ will then be removed from the received vectors, similar to (4.20). The IP-SIC for the rest groups $G_{(2)}, \cdots, G_{(\Psi)}$ in the ordered list will follow the same procedure: adding SIC estimated signals back, diversity combining, STBC decoding and cancelling the IP-SIC estimated signals from the received vectors. By IP-SIC, the diversity order of all groups can increase to $2n$ if no error propagation is present. The IP-SIC can also be performed more than one round to improve reliability furthermore.

www.manaraa.com

Alternatively, similar to the method described in section 3.4 for plain V-BLAST case, in applying IP-SIC, we can combine less diversity paths and utilize these saved DoF to suppress some residues. Suppose residues of q groups $G_{[1]}, G_{[2]}, \cdots, G_{[q]}$ are to be suppressed, we form matrix $\tilde{\mathbf{H}} = [\mathbf{A}_{[1]}\ \mathbf{A}_{[2]}\ \cdots\ \mathbf{A}_{[q]}]$ that contains channel gains of these groups and the suppression matrix $\mathbf{L}$ for $\tilde{\mathbf{H}}$; the suppression matrix $\mathbf{L}$ consists of $n - \sum_{i=1}^{q} \rho_{[i]}$ orthonormal vectors. Applying $\mathbf{L}$ on modified received vectors will remove the residues of groups $G_{[1]}, G_{[2]}, \cdots, G_{[q]}$; for the case that all groups encoded with 2×2 Alamouti STBC, we replace the identity matrix $\mathbf{I}$ with $\mathbf{L}$ in (4.24) for decoding. The result is that residues of q groups are cancelled, with $\sum_{i=1}^{q} \rho_{[i]}$ less diversity paths combined compared to IP-SIC without residue suppression.

المنارة للاستشارات

www.manaraa.com

CHAPTER 5. RESULTS AND DISCUSSIONS

Simulation results and discussions on the bit error rate (BER) are provided in this chapter. The simulations were conducted in Monte Carlo fashion using MATLAB software and terminated when the accumulated number of errors is 100. BPSK modulation is considered in all numerical results.

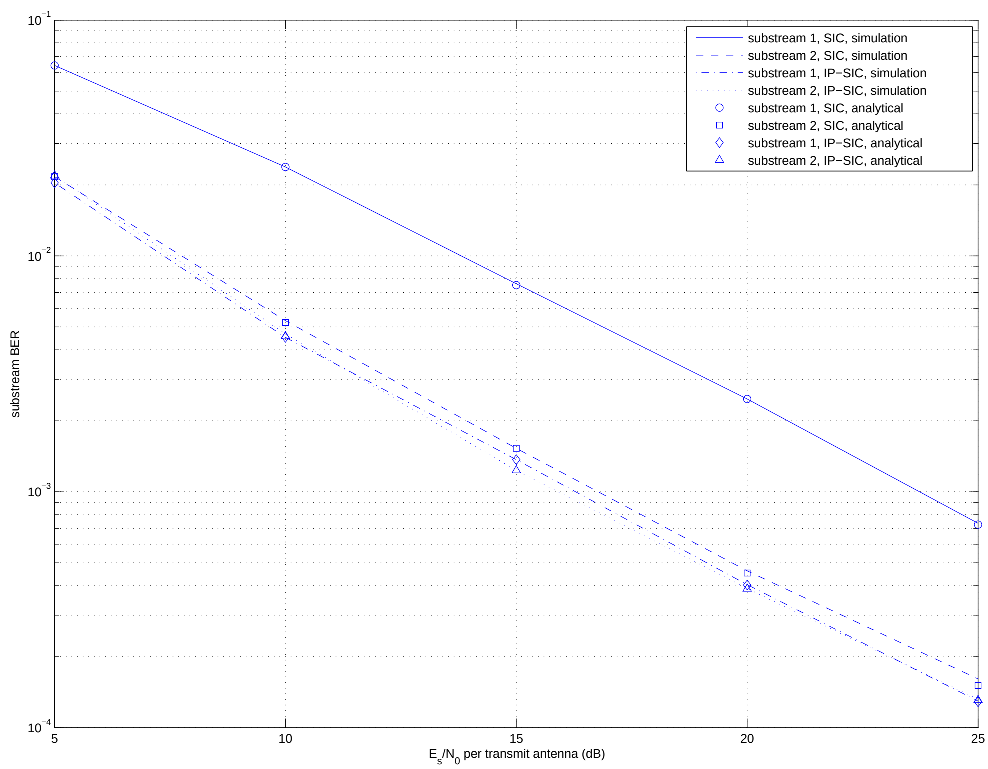

Figure 5.1 BER of different substreams versus E_s/N_0 per transmit antenna, $m = n = 2$; BPSK, ZF, no ordering.

Fig. 5.1 shows the BER of different substreams versus E_s/N_0 per transmit antenna for the

المنارة للاستشارات
www.manaraa.com

2×2 ZF V-BLAST system ($m = n = 2$) with SIC and IP-SIC, without ordering; the results are obtained by both simulation and analytical calculation. We find that the power gains provided by IP-SIC over SIC-only are 7 dB and 1 dB for the first and the second substream respectively, at BER=10^{-3}. The reason for the first substream to have larger gain than the second substream is that though IP-SIC, the diversity order for the first substream is increased from 1 to 2 assuming there is no residue, while for the second substream there is no diversity order improvement and its gain comes from residue reduction of the first substream only. The analytical equation (3.23) is seen to be quite accurate in calculating the substream error probabilities of the IP-SIC algorithm for ZF without ordering.

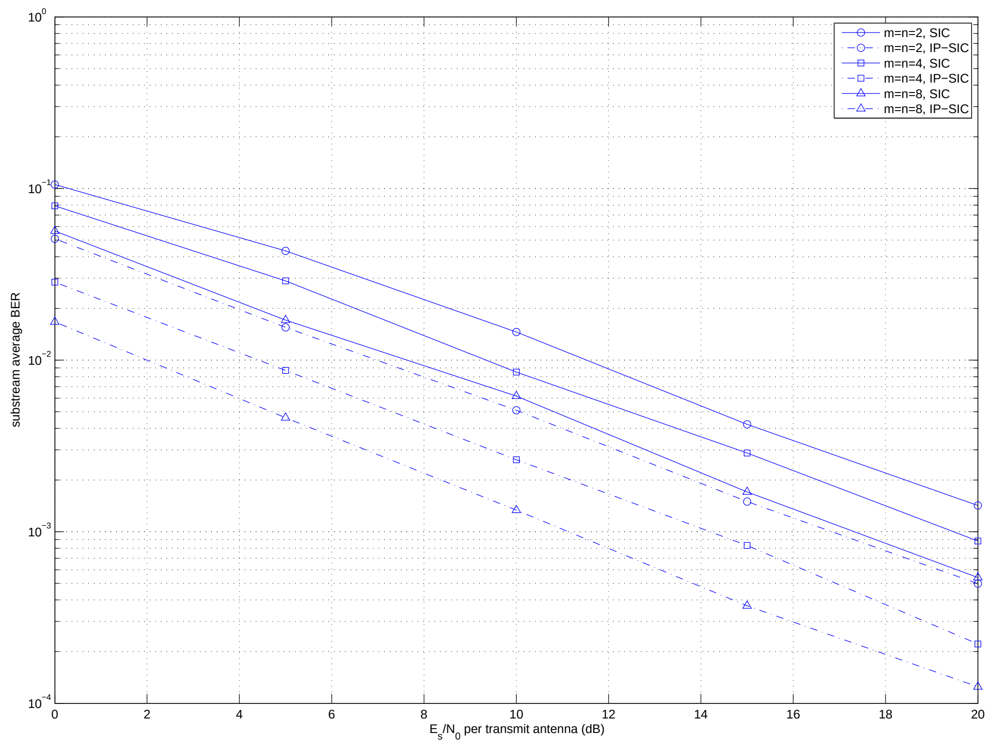

Figure 5.2 Average BER versus E_s/N_0 per transmit antenna, various numbers of antennas; BPSK, ZF, no ordering.

Fig. 5.2 compares average BER versus E_s/N_0 per transmit antenna for ZF systems with different number of transmit and receive antennas, with SIC and IP-SIC without ordering. At

المنارة للاستشارات
www.manaraa.com

the error rate of 10^{-2}, the power gain provided by IP-SIC over SIC only are 4.5 dB, 5.1 dB and 5.7 dB for $m = n = 2$, $m = n = 4$ and $m = n = 8$ respectively. We find that as the number of antennas increases, the power gain becomes larger; this is due to the fact that the diversity order increases as the number of antennas grows.

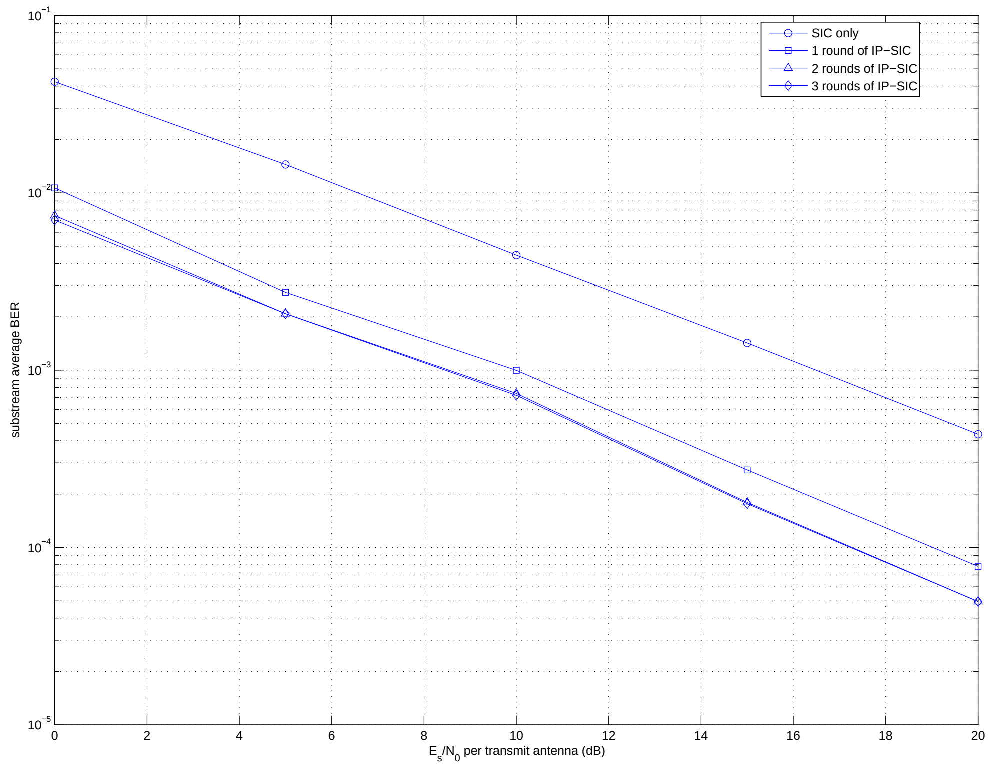

Figure 5.3 Average BER versus E_s/N_0 per transmit antenna, $m = n = 10$; multiple rounds of IP-SIC, BPSK, ZF, no ordering.

Fig. 5.3 shows BER when more than one round of IP-SIC is applied versus E_s/N_0 per transmit antenna. These simulations were performed on 10×10 ZF V-BLAST system without ordering. We find that at BER=10^{-3}, applying 1 round of IP-SIC provides a power gain of 6.4 dB over SIC only. 2 rounds of IP-SIC provides additional 1.5 dB gain. We find that no further gain is obtained beyond 2 rounds of IP-SIC.

Fig. 5.4 shows the average BER of IP-SIC applied on various numbers of substreams versus E_s/N_0 per transmit antennas. The system simulated is of 8×8 ZF without ordering. We find

المنارة للاستشارات
www.manaraa.com

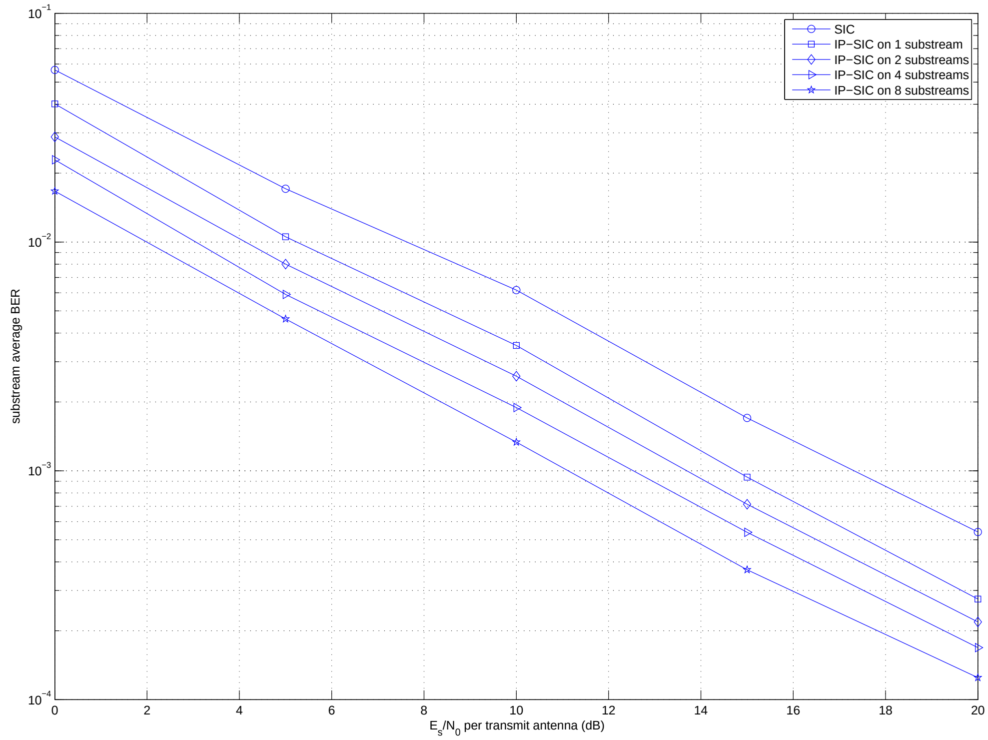

Figure 5.4 Average BER versus E_s/N_0 per transmit antenna, $m = n = 8$; IP-SIC on various numbers of substreams, BPSK, ZF, no ordering.

that at BER=10^{-3}, applying IP-SIC on first substream only provides a power gain of 2.6 dB over SIC only; by applying IP-SIC on the first two substreams, the power gain over SIC only is 3.5 dB; applying IP-SIC on the first four substreams, a power gain of 4.5 dB over SIC only is achieved; applying IP-SIC for one round (all eight substreams), we obtain a 5.7 dB gain over SIC only.

Fig. 5.5 shows the average BER versus E_s/N_0 per transmit antenna for ZF systems with various ordering schemes: no ordering, SNR based ordering and LLR based ordering. We can see that the power gains provided by the IP-SIC scheme for these ordering techniques at BER=10^{-3} are: 5.7 dB for no ordering, 4.4 dB for SNR ordering and 2.7 dB for LLR ordering. We also find that the performances of SNR-ordered SIC and IP-SIC for no ordering case are

المنارة للاستشارات
www.manaraa.com

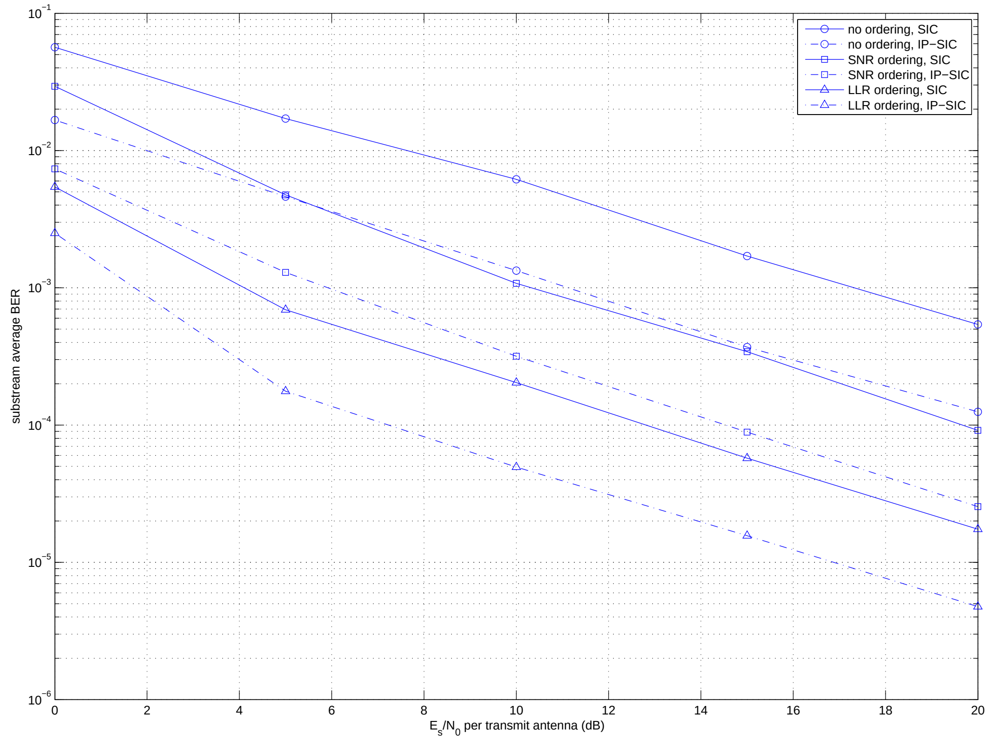

Figure 5.5 Average BER versus E_s/N_0 per transmit antenna, $m = n = 8$; BPSK, ZF, various ordering schemes.

comparable. However, the complexity of IP-SIC is only $\mathcal{O}(n)$, which is much less than that of SNR based ordering, $\mathcal{O}(mn)$.

Fig. 5.6 compares the BER with MMSE type interference suppression for various ordering schemes: no ordering, SNR based ordering and LLR based ordering. We find that at BER=10^{-3}, IP-SIC provides 2.2 dB gain over SIC for non-ordered case, 0.9 dB gain for SNR ordered case and 0.6 dB gain for LLR ordered case. Compared to ZF case, the gain of IP-SIC over SIC only for MMSE is smaller. This is because unlike ZF, the SIC detection errors of MMSE are concentrated on the last several substreams in the ordered list, which can be seen from Fig. 3.3. The diversity order increase due to IP-SIC is small for the last several substreams in the ordered list and thus small power gain results.

Fig. 5.7 shows the performance of the alternative IP-SIC scheme – IP-SIC with partial

المنارة للاستشارات
www.manaraa.com

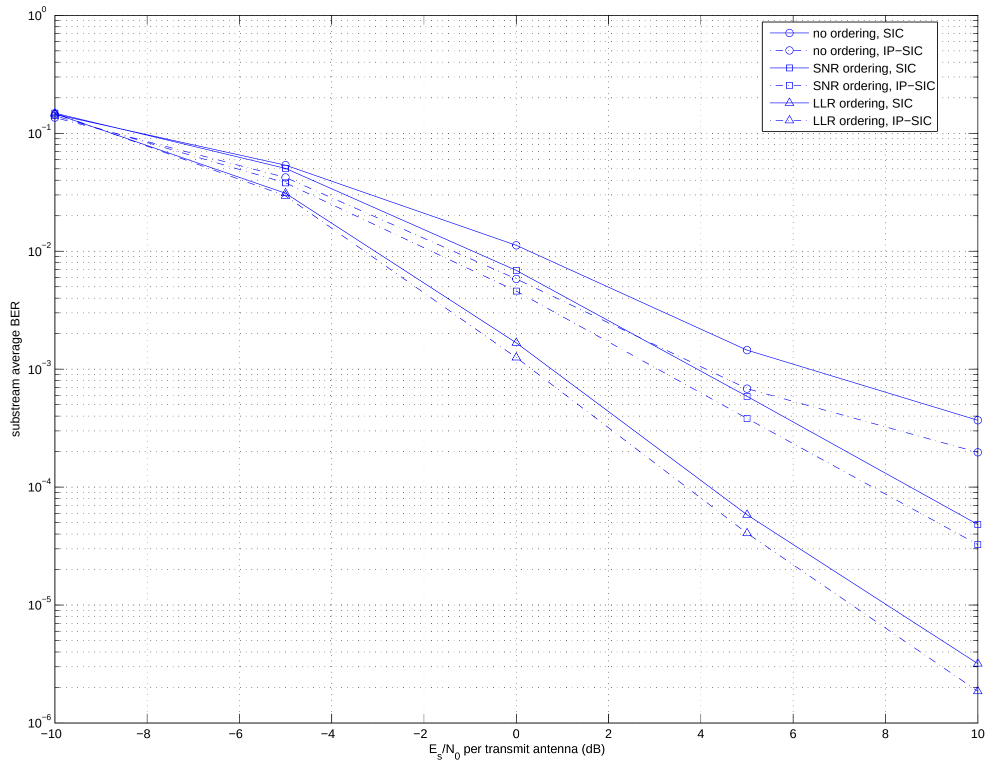

Figure 5.6 Average BER versus E_s/N_0 per transmit antenna, $m = n = 8$; BPSK, MMSE, various ordering schemes.

residue cancellation, discussed in section 3.4. The systems simulated are of ZF without ordering. The residues cancelled correspond to the circularly next q substreams, as in (3.27), after substream k', the substream being processed. We find that at BER=10^{-3}, for $q = 1$, the additional gain over no residue cancellation case is 1.1 dB; when $q = 2$, the additional gain drops to 0.6 dB. The implication is that for this particular case, suppressing $q = 1$ residues and increasing diversity order of all substreams to $n - q = 7$ in IP-SIC yields better performance compared to suppressing $q = 2$ residues and increasing diversity order of all substreams to $n - q = 6$.

Fig. 5.8 illustrates the performance of IP-SIC with partial residue cancellation for an LLR ordered MMSE system. From Fig. 3.3 we see that the residues are concentrated on the last several substreams in the ordered list, therefore we are motivated to choose residues of the last

www.manaraa.com

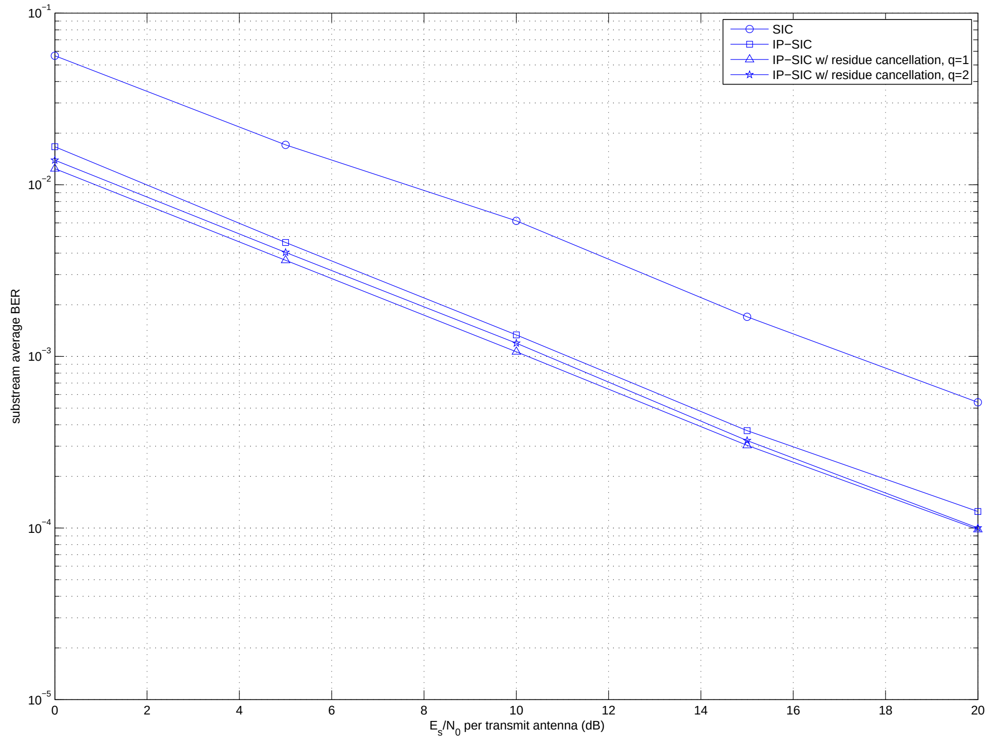

Figure 5.7 Average BER versus E_s/N_0 per transmit antenna, $m = n = 8$; BPSK, ZF, no ordering, various numbers of residue cancellation.

several substreams to suppress in IP-SIC. Specifically, for substreams $(1')$ to $(6')$, we fix the $q = 2$ residues to suppress to be from the last 2 substreams, *i.e.* $\mathbf{H}_{(k')} = [\mathbf{h}_{(k)}, \mathbf{h}_{(7)}, \mathbf{h}_{(8)}], k = 1, \cdots, 6$; for substream $(7')$, the $q = 1$ residue to suppress is chosen to be the from the last substream, *i.e.* $\mathbf{H}_{(7')} = [\mathbf{h}_{(7)}, \mathbf{h}_{(8)}]$; for substream $(8')$, no residue is suppressed, or $\mathbf{H}_{(8')} = \mathbf{h}_{(8)}$. We find that at BER=$10^{-5}$, IP-SIC with this residue suppression strategy provides 1.3 dB power gain over IP-SIC with no residue suppression.

Fig. 5.9 shows the performance of IP-SIC on joint ML/SIC combined system, discussed in section 3.5. The joint ML/SIC scheme detect the first two substreams using ML detection. We find that the diversity order with joint ML/SIC is 2. At BER=10^{-3}, the power gain provided by joint ML/IP-SIC over joint ML/SIC is 3.5 dB. Compared to the 5.7 dB gain for IP-SIC over SIC only, the gain in systems with joint ML/SIC is smaller; this is because the overall

المنارة للاستشارات
www.manaraa.com

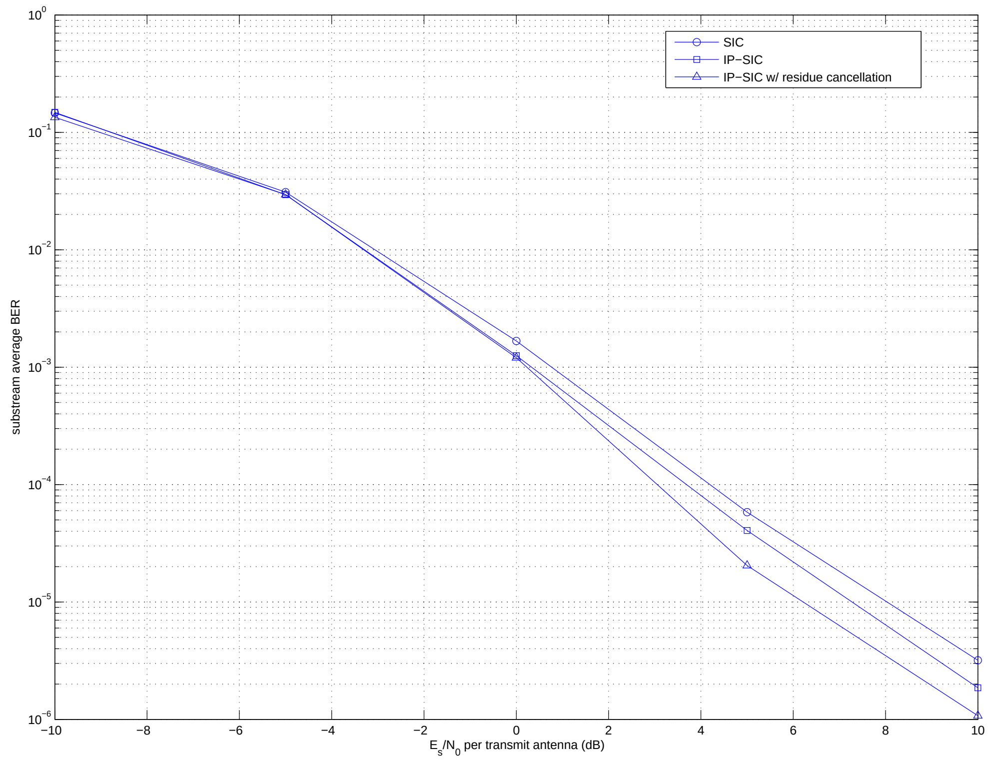

Figure 5.8 Average BER versus E_s/N_0 per transmit antenna, $m = n = 8$; BPSK, MMSE, LLR ordering, with and without residue suppression.

diversity order for the joint ML/SIC system is higher, which limits the diversity improvement from IP-SIC.

Fig. 5.10 shows the average BER of space-time block coded V-BLAST system, the transmit antennas are divided into $\Psi = 4$ groups, with each group forming 2×2 Alamouti STBC. Group-wise SIC is performed in order of SNR. We find that IP-SIC can provide a power gain of 3.2 dB at BER of 10^{-3}, over SIC only system.

المنارة للاستشارات
www.manaraa.com

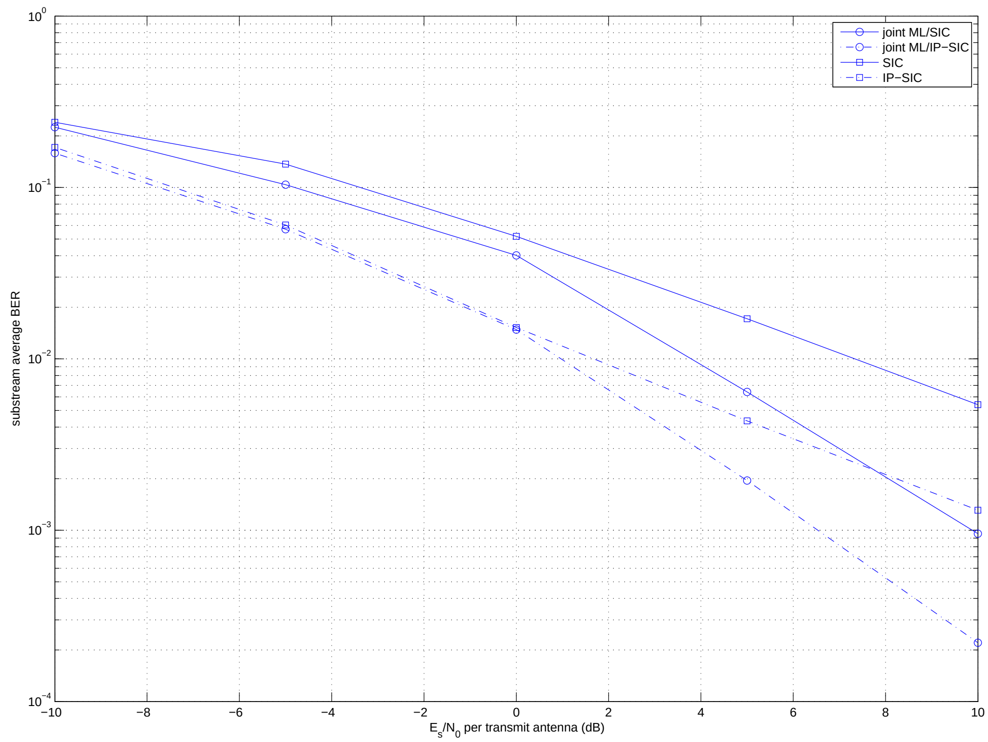

Figure 5.9 Average BER versus E_s/N_0 per transmit antenna, $m = n = 8$; BPSK, ZF, no ordering, joint ML/SIC.

المنارة للاستشارات
www.manaraa.com

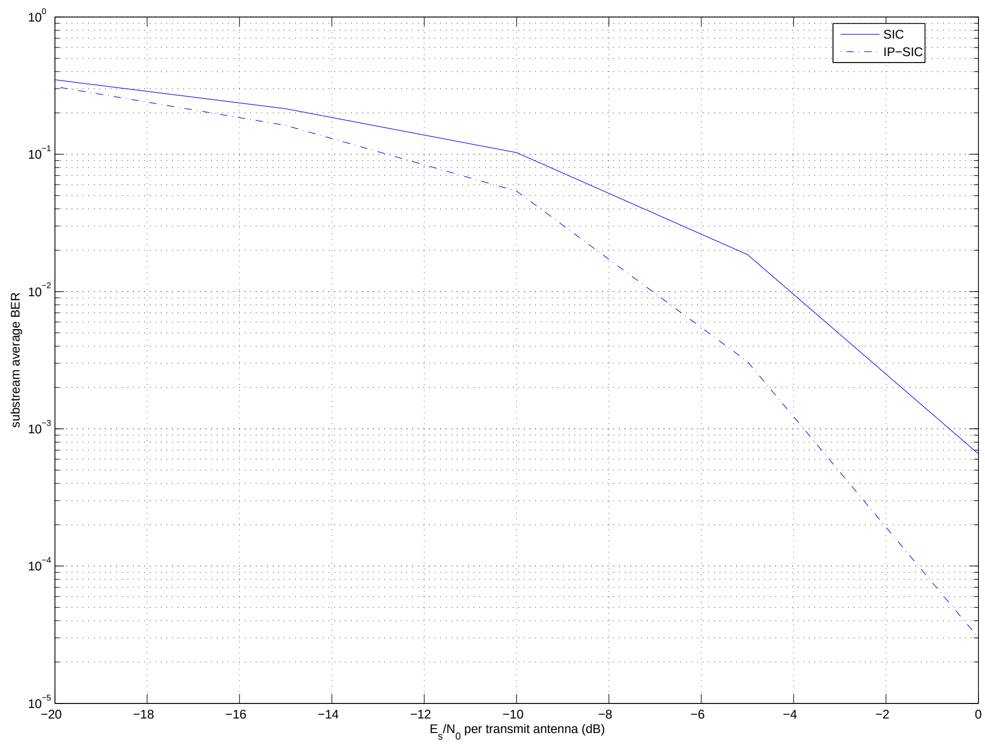

Figure 5.10 Average BER versus E_s/N_0 per transmit antenna, $m = n = 8$, four groups; 2×2 Alamouti STBC, BPSK, SNR ordering.

www.manaraa.com

CHAPTER 6. CONCLUSIONS AND FUTURE WORK

6.1 Conclusions

Novel scheme (IP-SIC) to enhance the reliability performance of the popular SIC decoding algorithm in V-BLAST system is developed in this thesis. Assuming there is no residues, the proposed scheme is able to increase the diversity order of all substreams to a fixed desired value to reduce the average error probability significantly. After the SIC process, interference from all substreams is removed from the received vector and a modified received vector consisting of only noise and residues results. In IP-SIC, the product of detected symbol and its channel vector is added to the modified received vector one at a time and the symbols are re-detected. Other than increasing the substream diversity order, the proposed approach is also capable of balancing the suppression of interference and the promotion in diversity order – specific residue can be suppressed at the expense of lowered diversity order; this feature can be used to further reduce the average error probability. The proposed technique is applied to joint ML/SIC as well as STBC encoded V-BLAST systems.

It is shown by analytical and simulation results that the decoding reliability of the proposed scheme significantly outperforms that of SIC process only. On the other hand, the additional computation complexity related to this new scheme is very small and will not affect the complexity order of the overall decoding. For example, in the case of 8×8 ZF V-BLAST system with BPSK modulation, the power gain provided by IP-SIC over no ordering SIC only is comparable to the power gain of SNR ordering SIC over no ordering SIC; nevertheless the complexity order of IP-SIC is only $\mathcal{O}(n)$, which is much less than the complexity order of SNR ordering, $\mathcal{O}(mn)$. The proposed scheme is promising in practice for the important performance enhancement it brings with low computational requirements.

www.manaraa.com

6.2 Future Work

A variety of works are available to be conducted in the future. In this thesis, only the performance analysis for no ordering ZF SIC with BPSK modulation is presented; generalizing the analysis to other modulation types, MMSE receiver and other ordering schemes will be of great interest. For instance, in the IP-SIC with partial residue suppression scheme discussed in section 3.4, to determine which residues are the strongest ones to suppress in order to yield optimized results, we would rely on the analytical substream error probability of SIC and IP-SIC to calculate the residues' variance.

As another advancement, the diversity/residue suppression trade-off can be further studied. How the degrees of freedom should be allocated for diversity paths combining and residue suppression, in order to result in optimized performance, needs to be thoroughly investigated.

For the STBC V-BLAST system, only the 2×2 Alamouti code is considered in this thesis. It is possible to employ other STBC [28] in each group to further promote the performance; how should the proposed scheme be modified to comply with the requirement of other STBC can be a topic for future study.

المنارة للاستشارات
www.manaraa.com

APPENDIX

Derivation of $\Pr(R_{i,\{2,3\}}), i = 0, 1, 2$ for Calculating $P_{e,1'}$ when $m = 3$

In this appendix, the probability $\Pr(R_{i,\{2,3\}}), i = 0, 1, 2$ that is used in (3.23) to calculate $P_{e,1'}$ when $m = 3$ is derived. Let events

$$\begin{aligned} A &\triangleq \{\text{substream 1 is detected in error during SIC}\}, \\ B &\triangleq \{\text{substream 2 is detected in error during SIC}\}, \\ C &\triangleq \{\text{substream 3 is detected in error during SIC}\}, \end{aligned} \tag{A.1}$$

then $\Pr(A) = P_{e,1}, \Pr(B) = P_{e,2}$ and $\Pr(C) = P_{e,3}$.

We can form a Venn's diagram as in Fig. A.1. The three event A, B and C divide the whole probability space into 8 regions, $K_1, K_2, \cdots, K_8$. Therefore we can construct

$$\Pr(A) = K_1 + K_2 + K_4 + K_5 \tag{A.2}$$

$$\Pr(B) = K_2 + K_3 + K_5 + K_6 \tag{A.3}$$

$$\Pr(C) = K_4 + K_5 + K_6 + K_7 \tag{A.4}$$

as well as

$$\Pr(B|A) = \Pr(R_{1,\{1\}}) = \frac{\Pr(BA)}{\Pr(A)} = \frac{K_2 + K_5}{K_1 + K_2 + K_4 + K_5} \tag{A.5}$$

$$\Pr(B|\bar{A}) = \Pr(R_{0,\{1\}}) = \frac{\Pr(B\bar{A})}{\Pr(\bar{A})} = \frac{K_3 + K_6}{K_3 + K_6 + K_7 + K_8} \tag{A.6}$$

$$\Pr(C|AB) = \Pr(R_{2,\{1,2\}}) = \frac{\Pr(CBA)}{\Pr(AB)} = \frac{K_5}{K_2 + K_5} \tag{A.7}$$

$$\Pr(C|\bar{A}B) = \Pr(R_{1,\{1,2\}})/2 = \frac{\Pr(CB\bar{A})}{\Pr(\bar{A}B)} = \frac{K_6}{K_3 + K_6} \tag{A.8}$$

$$\Pr(C|A\bar{B}) = \Pr(R_{1,\{1,2\}})/2 = \frac{\Pr(C\bar{B}A)}{\Pr(A\bar{B})} = \frac{K_4}{K_1 + K_4} \tag{A.9}$$

$$\Pr(C|\bar{A}\bar{B}) = \Pr(R_{0,\{1,2\}}) = \frac{\Pr(C\bar{B}\bar{A})}{\Pr(\bar{A}\bar{B})} = \frac{K_7}{K_7 + K_8} \tag{A.10}$$

المنارة للاستشارات
www.manaraa.com

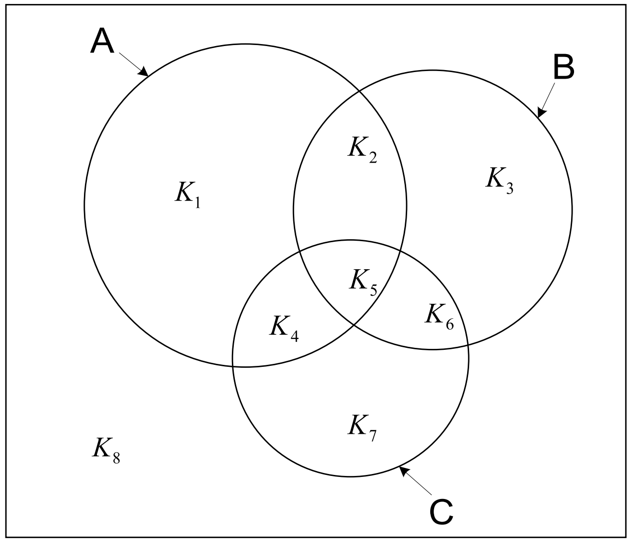

Figure A.1 Venn's diagram of events A,B and C

Using these equations, $K_1, K_2, \cdots, K_8$ can be solved respectively. Therefore the probabilities we are interested in,

$$\Pr(R_{0,\{2,3\}}) = K_1 + K_8 \tag{A.11}$$

$$\Pr(R_{1,\{2,3\}}) = K_2 + K_3 + K_4 + K_7 \tag{A.12}$$

$$\Pr(R_{2,\{2,3\}}) = K_5 + K_6 \tag{A.13}$$

can be obtained. Plug these values into (3.25), $P_{e,1'}$, the error probability of substream 1 after IP-SIC in an $m = 3$ V-BLAST system with no ordering SIC, is derived.

To get the probability of $\Pr(R_{i,\{1',3\}}), i = 0, 1, 2$, we need to form a fourth circle D in the Venn's diagram and follow a similar procedure.

المنارة للاستشارات
www.manaraa.com

BIBLIOGRAPHY

[1] J. Winters, "On the capacity of radio communications systems with diversity," *IEEE Journals on Selected Areas of Communication*, pp. 871–878, June 1987.

[2] G. J. Foschini and M. Gans, "On limits of wireless communications in a fading environment when using multiple antennas," *Wireless Personal Communications*, pp. 311–335, March 1998.

[3] E. Telatar, "Capacity of multi-antenna gaussian channels," *AT&T Bell Labs Internal Technical Memo*, June 1995.

[4] E. Telatar, "Capacity of multi-antenna gaussian channels," *European Transactions on Telecommunication*, pp. 585–596, November 1999.

[5] M. K. Simon and M.-S. Alouini, *Digital Communication over Fading Channels: A Unified Approach to Performance Analysis.* New York, NY: Wiley, 2000.

[6] A. Goldsmith, *Wireless Communications.* New York, NY: Cambridge University Press, 2005.

[7] A. M. Tulino and S. Verdu, "Random matrix theory and wireless communicaitons," *Foundations and Trends in Communications and Information Theory*, pp. 1–182, 2004.

[8] L. Zheng and D. N. Tse, "Diversity and multiplexing: A fundamental trade-off in multiple antenna channels," *IEEE Transactions on Information Theory*, pp. 1073–1096, May 2003.

[9] H. Yao and G. Wornell, "Structured space-time block codes with optimal diversity-multiplexing," in *Proc. IEEE Globecom '03*, pp. 1941–1945, December 2003.

المنارة للاستشارات
www.manaraa.com

[10] D. N. C. Tse, P. Viswanath, and L. Zheng, "Diversity-multiplexing tradeoff in multiple-access channels," *IEEE Transactions on Information Theory*, vol. 50, pp. 1859–1874, September 2004.

[11] H. Gamal, G. Caire, and M. Damon, "Lattice coding and decoding achieve the optimal diversity-multiplexing trade-off of MIMO channels," *IEEE Transactions on Information Theory*, pp. 968–985, June 2004.

[12] G. J. Foschini, "Layered space-time architecture for wireless communication in a fading environment when using multiple antennas," *Bell Labs Technical Journal*, vol. 1, no. 2, pp. 41–59, 1996.

[13] G. D. Golden, G. J. Foschini, R. A. Valenzuela, and P. W. Wolniansky, "Detection algorithm and initial laboratory results using V-BLAST space-time communication architecture," *Electron. Letters*, vol. 35, pp. 14–16, January 1999.

[14] G. Ginis and J. M. Cioffi, "On the relationship between V-BLAST and GDFE," *IEEE Communication Letters*, vol. 5, pp. 364–366, September 2001.

[15] M. Varanasi and T. Guess, "Optimum decision feedback multiuser equalization with successive decoding achieves the total capacity of the Gaussian multiple-access channel," *Conference Record of the Thirty-First Asilomar Conference on Signals, Systems and Computers*, vol. 2, pp. 1405–1409, November 1997.

[16] G. J. Foschini, G. D. Golden, R. A. Valenzuela, and P. W. Wolniansky, "Simplified processing for high spectral efficiency wireless communication employing multiple-element arrays," *Wireless Personal Communications*, vol. 6, pp. 311–335, March 1999.

[17] S. Loyka and F. Gagnon, "Performance analysis of the V-BLAST algorithm: An analytical approach," *IEEE Transactions on Wireless Communications*, vol. 3, pp. 1326–1337, July 2004.

[18] S. W. Kim and K. P. Kim, "Log-likelihood-ratio-based detection ordering in V-BLAST," *IEEE Transactions on Communications*, vol. 54, pp. 302–307, February 2006.

المنارة للاستشارات

www.manaraa.com

[19] S. Verdu, *Multiuser Detection, 2nd ed.* Cambridge, U.K.: Cambridge University Press, 1998.

[20] A. Benjebbour, H. Murata, and H. Yoshida, "Comparison of ordered successive receivers for space-time transmission," in *Proc. IEEE Vehicular Technology Conference '01*, vol. 4, pp. 2053–2057, October 2001.

[21] M. Xu, J. Gan, S. Zhou, and Y. Yao, "Log-likelihood detection ordering for MMSE-SIC in V-BLAST systems," in *Proc. IEEE Vehicular Technology Conference '05*, vol. 1, pp. 1–5, September 2005.

[22] C. Shen, Y. Zhu, S. Zhou, and J. Jiang, "On the performance of V-BLAST with zero-forcing successive interference cancellation," in *Proc. IEEE Globecom '04*, vol. 5, pp. 2818–2822, November 2004.

[23] J. G. Proakis, *Digital Communications, 4th ed.* Boston, MA: McGraw-Hill, 2001.

[24] Y. Jiang, X. Zheng, and J. Li, "Asymptotic performance analysis of V-BLAST," in *Proc. IEEE Globecom '05*, vol. 6, pp. 3882–3886, December 2005.

[25] W. Choi, R. Negi, and J. M. Cioffi, "Combined ML and DFE decoding for the V-BLAST system," in *Proc. IEEE ICC '00*, pp. 1243–1248, June 2000.

[26] V. Tarokh, A. Naguib, N. Seshadri, and A. R. Calderbank, "Combined array processing and space-time coding," *IEEE Transactions on Information Theory*, vol. 45, pp. 1121–1128, May 1999.

[27] S. Alamouti, "A simple transmit diversity techinique for wireless communications," *IEEE Journal on Selected Areas of Communication*, vol. 16, pp. 1451–1458, October 1998.

[28] E. G. Larsson and P. Stoica, *Space-Time Block Coding for Wireless Communications.* Cambridge, U.K.: Cambridge University Press, 2003.

المنارة للاستشارات

www.manaraa.com