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Transmission and detection for space-time block coding and v-blast systems

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ABSTRACT

TRANSMISSION AND DETECTION FOR SPACE-TIME BLOCK CODING AND V-BLAST SYSTEMS

by
Lei He

This dissertation focuses on topics of data transmission and detection of space -time block codes (STBC). The STBCs can be divided into two main categories, namely, the orthogonal space-time block codes (OSTBC) and the quasi-orthogonal space-time codes (Q-OSTBC). The space-time block coded systems from transceiver design perspective for both narrow-band and frequency selective wireless environment are studied. The dissertation also processes and studies a fast iterative detection scheme for a high-rate space-time transmission system, the V-BLAST system.

In Chapter 2, a new OSTBC scheme with full-rate and full-diversity, which can be used on QPSK transceiver systems with four transmit antennas and any number of receivers is studied. The newly proposed coding scheme is a non-linear coding. Compared with full-diversity QOSTBC, an obvious advantage of our proposed new OSTBC is that the coded signals transmitted through all four transmit antennas do not experience any constellation expansion.

In Chapter 3, a new fast coherent detection algorithm is proposed to provide maximum likelihood (ML) detection for Q-OSTBC. The new detection scheme is also very useful to analysis the diversity property of Q-OSTBC and design full diversity Q-OSTBC codes. The complexity of the new proposed detection algorithm can be independent to the modulation order and is especially suitable for high data rate transmission.

In Chapter 4, the space-time coding schemes in frequency selective channels are studied. Q-OSTC transmission and detection schemes are firstly extended for frequency selective wireless environment. A new block based quasi-orthogonal space-

time block encoding and decoding (Q-OSTBC) scheme for a wireless system with four transmit antennas is proposed in frequency selective fading channels. The proposed MLSE detection scheme effectively combats channel dispersion and frequency selectivity due to multipath, yet still provides full diversity gain. However, since the computational complexity of MLSE detection increases exponentially with the maximum delay of the frequency selective channel, a fast sub-optimal detection scheme using MMSE equalizer is also proposed, especially for channels with large delays.

The Chapter 5 focuses on the V-BLAST system, an important high-rate space-time data transmission scheme. A reduced complexity ML detection scheme for V-BLAST systems, which uses a pre-decoder guided local exhaustive search is proposed and studied. A polygon searching algorithm and an ordered successive interference cancellation (O-SIC) sphere searching algorithm are major components of the proposed multi-step ML detectors. At reasonable high SNRs, our algorithms have low complexity comparable to that of O-SIC algorithm, while they provide significant performance improvement. Another new low complexity algorithm termed ordered group-wise interference cancellation (O-GIC) is also proposed for the detection of high dimensional V-BLAST systems. The O-GIC based detection scheme is a sub-optimal detection scheme, however, it outperforms the O-SIC.

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TRANSMISSION AND DETECTION FOR SPACE-TIME BLOCK CODING AND V-BLAST SYSTEMS

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To my parents and my wife

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CHAPTER 1

INTRODUCTION

Diversity techniques are effective ways of combating channel fading and providing reliable system performance in wireless communications. Diversity techniques include time, frequency and space diversity. The *time diversity* is provided by channel coding in combination with limited interleaving using time domain processing techniques. However, while channel coding is extremely effective in fast-fading environment (high mobility), it offers very little protection under slow fading unless significant interleaving delay can be tolerated. The fact that signals transmitted over different frequencies induce different multipath structures and independent fading is exploited to provide *frequency diversity*. However, when the multipath delay spread is small, compared to the symbol period, there is no frequency diversity. To utilize the *space diversity*, the receiver/transmitter uses multiple antennas that are separated for reception/ transmission to create independent fading channels.

Traditionally, space diversity referred to receiver diversity. Multiple antennas at base stations have been exploited in various communication systems, such as microwave relay stations and cellular based stations, to obtain receive diversity and improve down-link communication. However, in many wireless systems, additional antennas may be expensive or impractical at the remote station, since the remote units are supposed to be with small volume and simple structure. The small size of the mobile receivers limits both the spatial resolution of the array (because of the small number of elements) and the diversity gain (because the elements are close to one another). In these cases multiple transmit antenna array is considered.

1.1 Transmit Diversity

One possible approach for antenna array transmit processing is transmit beam forming [1][2], which provides array gain at the subscriber unit. In these schemes, the transmitter typically operates in "colsed-loop" i.e., it use channel information that is fed to it by the receiver through the reverse link in order to shape beams in the forward link. The success of transmit beam-forming depends on the quality of the channel estimates, the feedback channel, the mapping between the two links, and the dynamics of the signal and interference. Closed-loop techniques typically suffer from reduced uplink capacity because of the extra channel information that is transmitted. It is not applicable to situation when the downlink and uplink use different frequency bands.

Most recently, transmit diversity technique, which provides diversity benefit at a receiver with multiple transmit antennas, has received much attention, especially for wireless cellular application. Transmit diversity is a technique where the information is transmitted from multiple antennas in order to mitigate the effects of signal fading. Multiple transmit antennas at the bade station will increase the downlink capacity with only minor increase in terminal complexity. Transmit diversity technique can be simpler to implement because it can operate in an open-loop, i.e., without channel knowledge at the transmitters. This mode of operation is particularly appealing when the mobile speed is too high to estimate channel information and feedback accurately. Moreover, open-loop techniques do not penalize the uplink capacity as closed-loop techniques do. These arguments suggest that multiple antenna open-loop transmit diversity is a practical way to improve the performance of current systems.

Recently, some interesting approaches for transmit diversity have been suggested. The transmit diversity scheme is a delay-transmit diversity scheme, which was proposed by Wittneben[3]. In a delay-transmit diversity, the base station transmits a delayed version of the original signal, hence creating an artificial multipath distortion [3][4][5][6]. However, due to the degrading orthogonality and increasing the interference level seen at the mobile receiver, delay transmit diversity has a limited link performance gain over non-transmit diversity [7].

Another approach in [8][9] recommended that each user can be assigned a different orthogonal code for each transmitter antenna. This scheme can provide arbitary-fold diversity and have very simple demodulation scheme, but the diversity gain and simplicity of this scheme come with the penalty of requiring more than one spreading codes per user. With M transmitter antennas, M times as many codes are needed, and with a limited number of orthogonal codes, this also means that M times fewer users can simultaneously be supported.

One open-loop approach that offers some diversity gains without requiring extra resource is orthogonal transmit diversity scheme [7][10][11][12], which was an early inclusion as an option in the IS-2000 standard (at that time called the CDMA-2000 standard. According to this technique, each user's data stream is extended by one spreading code and transmitted through each transmit antenna in a scanning way. For example, in a system with two transmit antennas, each user's data stream is split into two independent data streams, namely into its odd and even substreams, respectively. In orthogonal transmit diversity, the first antenna transmits the spread (and possibly coded) odd substream of the user's data during odd symbol periods and nothing during even symbol periods. Conversely, the second antenna transmits nothing during odd symbol periods and the spread even substream during even symbol periods. However, orthogonal transmit diversity provides unbalanced diversity to each of the user's data substreams; and usually orthogonal transmit diversity can not provide full space diversity. Without interleaving or channel coding, orthogonal transmit diversity does not improve the transmission performance.

Above mentioned transmit diversity schemes are mostly presented in the research for wide-band communication systems, DS-CDMA etc.. The transmit diversity also can be investigated in narrow band systems, and then extended for wide-band systems.

1.2 Space-Time Coding

Another very important approach for transmit diversity is space-time coding technique. Space-time coding techniques are first put forward under assumption of narrow band wireless fading environment. But these techniques can be easily applied to wide-band communication, combined with other wide-band communication techniques, such as CDMA [13], OFDM. Space-time coding can be classified into two broad categories, space-time trellis coding and space-time block coding.

*Space-time trellis coding (*STTC) proposed by Tarokh in [14] combines signal processing at the receiver with coding techniques appropriate to multiple transmit antennas. STTC operates on one input symbol at a time producing a sequence of vector symbols whose length equals to the number of transmit antennas. Like traditional trellis coding technique for the single-antenna channel, space-time trellis codes provide coding gain. Since they also provide full diversity gain, their key advantage over space-time block codes is the provision of coding gain. Specific space-time trellis codes design for 2-4 transmit antennas perform extremely well in slow-fading environments and come close to the outage capacity computed by Telatar[15] and independently by Foschini and Gans[16]. However, when the number of transmit antennas is fixed, the decoding complexity of space-time trellis codes increases exponentially with transmission rate. The disadvantage of STTC is that it is extremely difficult to design and require a computationally intensive encoder and decoder.

The other kind of space-time coding is *space-time block coding* (STBC). The first space-time block code was proposed by Alamouti [17]. It was further developed and put into a theoretical framework by Tarokh, Jafarkhani, and Calderbank in [18].

Space-time block codes operate on a block of input symbols and produce a matrix output whose columns represent time and rows represent antennas. Unlike traditional single antenna block codes for the AWGN channel, some space-time block codes may not provide error correction ability. The key feature of STBC is the provision of full diversity with extremely low encoder and decoder complexity. The space-time block codes first studied in [17] and [14] are orthogonal space-time block codes. Due to their inherent orthogonality, these STBC benefit from full diversity gain with low complexity receiver since the corresponding Maximum Likelihood decoding scheme is based on linear processing only: the detection matrix is simply the transconjugate of the equivalent channel matrix, which is unitary. When several receiving antennas can be used, the signals are recombined according to the Maximum Ratio Combining (MRC) technique. However, except for the system with two transmitters, there is no rate one complex orthogonal code design for systems with multiple transmitter (beyond 2) antennas. Among reported orthogonal space-time codes with fast ML decoding, the highest rate is 3/4 achieved in systems with 3 transmitter antennas [13]. Another rate 3/4 orthogonal code can be found in system with four transmitter antennas [19], but this code has no fast ML decoding algorithm. Trading the code orthogonality for higher coding rates, a full-rate and half-diversity code, the so called quasi-orthogonal codes, was proposed in [20]. The performance and diversity property of quasi-orthogonal codes can be improved through constellation rotation [21]. In [20] and [21], it has been pointed out that the decoding of quasi-orthogonal codes can be done by searching symbols pair by pair, but the decoding method reported is trivial and the symbol pair exhaustive searching is used for ML detection. As larger size constellation is used in modulation, the searching complexity increase exponentially. Therefore, the pair-wise exhaustive searching algorithm is still computationally intensive for applications in real time high rate data transmission.

1.3 STBC Schemes for Frequency-Selective Fading Channels

Recently, the following transmission schemes have been proposed to extend the Alamouti STBC to frequency-selective channels. For orthogonal STBC, even in some systems with frequency-selective fading, orthogonality is almost preserved by the communication channel, as long as the ensuring intersymbol interference can be reliably equalized.

1.3.1 W-CDMA-STBC

In wide band code-divsion multiple-access (W-CDMA) systems, where the signaling period is $\gg$ delay spread, and the Rake receiver separates the multipath. For each, the orthogonality of the multiantenna transmission is almost preserved. In such systems, orthogonal space-time block coding is a viable candidate for providing transmit diversity. Accordingly, the two-antenna complex modulation STBC proposed by Alamouti has been accepted as an open-loop transmit diversity scheme for third-generation wireless communication system, namely, the wideband CDMA standardized by 3GPP(3rd Generation Partnership Project)[13].

1.3.2 OFDM-STBC and FDE-STBC

OFDM-STBC [22][23], where orthogonal frequency division multiplexing (OFDM) is used to convert each frequency-selective channel into parallel independent frequency-flat subchannels using the computationally-efficient Fast Fourier Transform (FFT). The Alamouti spatio-temporal orthogonal structure is then imposed on each frequency-flat subchannel over two consecutive OFDM blocks transmitted from each antenna. Similar to OFDM-STBC, the Alamouti code is imposed in the frequency domain and over two consecutive transmission blocks, in Frequency-Domain Equalization-STBC (FDE-STBC)[24]. However, the two schemes differ in that FDE-STBC is a single-carrier transmission scheme and decision are made in the time domain.

1.3.3 TR-STBC

In time-reversal space-time block coding (TR-STBC) scheme, a clever combination of time-domain filtering, conjugation, and time reversal operation is used to convert the 2-input single-output system composed of the 2 frequency-selective channels $h_1(D)$ and $h_2(D)$ into an equivalent single-input single-output (SISO) system. Then, standard SISO equalization is applied to detect the 2 information streams.

1.4 BLAST System

Recent advances in information theory reveal an important fact that the rich-scattering multi-path wireless channel can provide enormous capacity if the multi-path propagation is properly exploited using multiple antennas in the transceiver system [16][25][22] [26]. The communications system architecture of the Vertical/Diagonal Bell Laboratories Layered Space-Time (V/D-BLAST) provides an experimental demonstration of such systems. In V-BLAST systems, independently encoded data streams are transmitted from each transmit antenna simultaneously, and detected at the receiver by nulling and successive interference cancellation (OSIC) scheme[27][28] or sphere decoding scheme[29][30][31]. In D-BLAST, each data stream switches transmit antenna after each symbol duration time slot and the completing of each data stream transmission is contribute by all transmit antennas, e.g., the t'th symbol of n'th ($n \in [0, N-1]$) data steam is transmitted by the $mod(t+n, N)$'th transmit antenna. Using V/D-BLAST system, very high data transmission rate can be achieved. However, V/D-BLAST can not be used when the number of receiver antennas is smaller than the number of transmit antennas.

If the data streams are independent to each other in BLAST system, V/D-BLAST structure can not provide transmit diversity gain. Combined with pre-channel coding techniques, V/D-BLAST are also efficient transmit diversity schemes in wireless communication.

1.5 Dissertation Overview

In this introductory chapter, the back ground have been laid for the subject materials of this dissertation. In Chapter 2, a new orthogonal space-time coding scheme with full-rate and full-diversity is proposed. The new proposed space-time coding scheme that can be used on QPSK transceiver systems with four transmit antennas and any number of receivers, is a non-linear space-time block code. The coded signals transmitted through all four transmit antennas do not experience any constellation expansion. The performance of the proposed coding scheme is studied in comparison with that of 1/2-rate full-diversity orthogonal space-time code, quasi-orthogonal code, as well as constellation rotated quasi-orthogonal code. The study shows that the proposed coding scheme offers full rate and outperforms the 1/2-rate orthogonal codes as well as full-rate quasi-orthogonal codes when SNR increases. Compared to the constellation rotated quasi-orthogonal codes (the improved QO scheme), the newly proposed code has the advantage of not expanding the signal constellation at each transmit antenna. The performance of the newly proposed code is comparable to that of the improved QO scheme.

Chapter 3 develops a new fast coherent detection algorithm, which provides maximum likelihood (ML) detection for quasi-orthogonal codes. Observed in a trans-formed signal space, the quasi-orthogonal space-time block code encoder embeds information into a group of orthogonal subspaces of rank two. A fast two-stage technique: separating information embedded in each rank two subspace by simple linear processing, and decoding the separated information independently. The newly proposed fast ML decoding algorithm decomposes the original $N \times N$ multiple-input and multiple-output (MIMO) system into $N/2$ parallel 2×2 MIMO sub-systems without information loss, and then detects the 2×2 MIMO system by sphere decoder independently. The proposed decoding algorithm has very low computational complexity.

In Chapter 4, a new block based quasi-orthogonal space-time block encoding and decoding (QO-STBC) scheme is proposed for a wireless system with four transmit antennas in frequency selective fading channels. The proposed scheme effectively combats channel dispersion and frequency selectivity due to multipath, yet still provides full diversity gain. It is also proposed a maximum likelihood (ML) solution, with a simple decoupled structure, to decoding the block based QO-STBC symbols at low complexity for channels with short delay. The proposed decoding technique can also be applied to the decoding of 3/4-rate Q-OSTBC, which is proposed in chapter 3. However, since the computational complexity of vector Viterbi Algorithm increases exponentially with the maximum delay of the frequency selective channel, the MLSE detection algorithm is not practical for the channel with long delay time. It is also developed a fast decoding algorithm for TR-QOSTBC, which highly reduces the decoding complexity, especially for channels with large value of delay. The diversity properties and performance of the MMSE detection are compared with those of the MLSE detection.

In Chapter 5, a multi-step ML decoder , which uses a pre-decoder guided local exhaustive searching algorithm with dramatically reduced complexity for ML detection in V-BLAST system is studied. A polygon searching algorithm and a O-SIC sphere searching algorithm are newly proposed for the multi-step ML decoder. The influence of pre-decoder on the decoding complexity was simulated. The complexity of the newly proposed algorithm depends on the operating signal to noise ratio (SNR) of the system. At reasonable high SNR, the complexity of newly proposed algorithm is comparable to that of ordered successive interference cancellation (O-SIC) algorithm, a commonly used decoding algorithms for V-BLAST. For V-BLAST system using a large number of transmit antennas, it is also proposed a new detection algorithm of ordered group-wise interference cancellation (O-GIC) to further reduce decoding complexity. The O-GIC based detection scheme is not a ML detection scheme, however, its performance is shown to outperform that of the O-SIC.

Chapter 6 concludes the works in this dissertation.

CHAPTER 2

A NEW FULL-RATE FULL-DIVERSITY ORTHOGONAL SPACE-TIME BLOCK CODING SCHEME

In this work, it is presented a new space-time orthogonal coding scheme with full-rate and full-diversity. The proposed space-time coding scheme is a orthogonal code with non-linear processing, which can be used on QPSK transceiver systems with four transmit antennas and any number of receivers. An additional feature is that the coded signals transmitted through all four transmit antennas do not experience any constellation expansion. The performance of the proposed coding scheme is studied in comparison with that of 1/2-rate full-diversity orthogonal space-time code, quasi-orthogonal code, as well as constellation rotated quasi-orthogonal code. The study shows that the proposed coding scheme offers full rate and outperforms the 1/2-rate orthogonal codes as well as full-rate quasi-orthogonal codes when SNR increases. Compared to the constellation rotated quasi-orthogonal codes (the improved QO scheme), the newly proposed code has the advantage of not expanding the signal constellation at each transmit antenna. The performance of the newly proposed code is comparable to that of the improved QO scheme.

2.1 System Model

Space-time block coding [17],[32], provides transmit diversity for a system with multiple transmit antennas in wireless communications. In general , a complex space-time block code is given by a $T \times N$ transmission matrix $\mathbf{C} \in \mathcal{G}$. Here, T represents the number of time slots for transmitting one block of symbols and N represents the number of transmit antennas.

2.1.1 Encoding

At first, the information bits are mapped to constellation symbols by digital modulation. Let $\mathcal{A}$ denote a signal constellation of cardinality 2^b. At each block, $b \times K$ bits are input into a digital modulator, hence mapped into K symbols of length K, $\mathbf{s} = (s_1, s_2, ..., s_K)$. Then, the block of K constellation symbols is mapped to one transmission matrix $\mathbf{C}(\mathbf{s})$. The n-th element of the t-th row of $\mathbf{C}(\mathbf{s})$, c_{tn}, represents the signal transmitted by the n-th transmit antenna at the t-th time slot. Therefore, all the transmit antennas transmit simultaneously and all the transmitted symbols have the same time duration. The coding rate is therefore defined as $R = K/T$. If $K = T$, the code is termed full rate or rate 1 code.

2.1.2 Channel Model and Data Formulation

A wireless communication system with N transmit antennas at the base station and M receiving antennas at the mobile host is considered. The wireless channel is assumed to be quasi-static so that the path gains are constant over a frame, and vary from one frame to another (block fading channel). Within each block, the path gain coefficients from transmit antenna n to receive antenna m, $h_{n,m}$'s, are modeled as a normalized samples of independent complex Gaussian random variables, $h_{n,m} \sim \mathcal{CN}(0,1)$.

At time slot t, the received signal at antenna m, $y_{t,m}$, is given by

$$y_{t,m} = \sum_{n=1}^{N} h_{n,m} c_{t,n} + v_{t,m}, \tag{2.1}$$

where the noise samples $v_{t,m}$ are spatially and temporally independent samples from a zero mean complex Gaussian random family, i.e. $v_{t,m} \sim \mathcal{CN}(0, 1/\mathrm{SNR})$. Note that the average energy of the symbols transmitted from each antenna is normalized to be $1/N$. The average power of the received signal at each receiver antenna is normalized, so that the signal to noise ratio SNR=σ^{-2} is presented as the *effective* noise variance at each receiving antenna.

This model can be recasted into an equivalent matrix form

$$\mathbf{y}_m = \sqrt{\frac{\rho}{N}}\mathbf{C}(\mathbf{s})\ \mathbf{h}_m + \mathbf{v}_m, \tag{2.2}$$

where $\mathbf{y}_m$ and $\mathbf{v}_m$ are $T\times 1$ vectors obtained by stacking $r_{t,m}$ and $v_{t,m}$ during processing time slots of dimension T, respectively. $\mathbf{h}_m = \begin{bmatrix} h_{1,m} & \cdots & h_{t,m} \end{bmatrix}^T$ is the $N\times 1$ fading channel vector , where "T" denotes the transpose operator.

If all the M receiving antenna are considered the system model is

$$\mathbf{Y} = \sqrt{\frac{\rho}{N}}\mathbf{C}(\mathbf{s})\ \mathbf{H} + \mathbf{V} \tag{2.3}$$

where $\mathbf{Y} = \{\ \mathbf{y}_1\ \ \mathbf{y}_2\ \ \cdots\ \ \mathbf{y}_m\ \}$ is the $T \times M$ receive matrix, $\mathbf{H} = \{\ \mathbf{h}_1\ \ \mathbf{h}_2\ \ \cdots\ \ \mathbf{h}_m\ \}$ is the $N \times M$ fading channel matrix, and $\mathbf{V} =\{\ \mathbf{v}_1\ \ \mathbf{v}_2\ \ \cdots\ \ \mathbf{v}_m\ \}$ is the $T \times M$ noise matrix. The entries of $\mathbf{H}$ and $\mathbf{V}$ are mutually independent, zero-mean, and circularly symmetric complex Gaussian variables of unit-variance, and ρ is the SNR per receiving antenna.

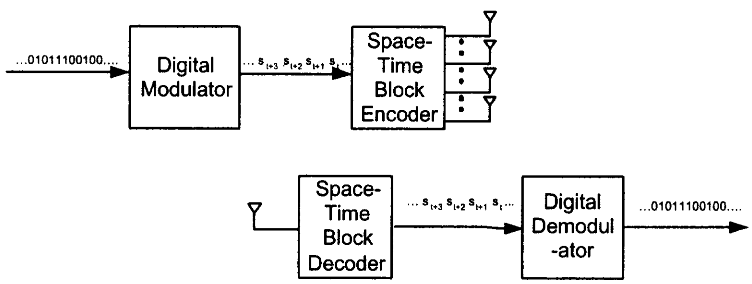

Figure 2.1 Structure of space-time block coding wireless communication system with 1 receiving anten.

2.1.3 Coherent Detection

Most work on space-time coding has assumed that perfect channel state information (CSI) is available at the receiver. It means that the receiver (but not the transmitter) knows the fading channel matrix $\mathbf{H}$. When $\mathbf{H}$ is known at the receiver, the pdf of the received data given that $\mathbf{C}(\mathbf{s}) \in \mathcal{G}$ was transmitted is,

$$p(\mathbf{Y}|\mathbf{H},\mathbf{C}(\mathbf{s})) = \frac{\exp(-\mathrm{tr}\{(\mathbf{Y}-\sqrt{\frac{\rho}{N}}\mathbf{C}(\mathbf{s})\ \mathbf{H})(\mathbf{Y}-\sqrt{\frac{\rho}{N}}\mathbf{C}(\mathbf{s})\ \mathbf{H})^H\})}{\pi^{TM}} \tag{2.4}$$

where "tr" is the trace, "H" is conjugate transpose.

Maximum-likelihood (ML) receiver reduces to the minimum Euclidean distance detector, i.e.

$$\begin{aligned} \widehat{\mathbf{C}} &= \arg\max_{\mathbf{C}\in\mathcal{C}(\mathbf{s})} p(\mathbf{Y}|\mathbf{H},\mathbf{C}(\mathbf{s})) \\ &= \arg\min_{\mathbf{C}\in\mathcal{C}(\mathbf{s})} \mathrm{tr}\{(\mathbf{Y}-\sqrt{\frac{\rho}{N}}\mathbf{C}(\mathbf{s})\ \mathbf{H})(\mathbf{Y}-\sqrt{\frac{\rho}{N}}\mathbf{C}(\mathbf{s})\ \mathbf{H})^H\} \end{aligned} \tag{2.5}$$

For unit-energy design, $\mathbf{C}^H\mathbf{C} = \mathbf{I}_{N\times N}$, Eq. (2.5) can be simplified as,

$$\widehat{\mathbf{C}} = \arg\max_{\mathbf{C}\in\mathcal{C}(\mathbf{s})} \mathcal{R}e\left\{\sum_{m=1}^{M}\mathbf{y}_m^H\mathbf{C}\ \mathbf{h}_m\right\}. \tag{2.6}$$

2.2 Orthogonal STBC and Quasi-Orthogonal STBC

In [18], the orthogonal space-time block codes are defined by

$$\mathbf{C}^H(\mathbf{s})\mathbf{C}(\mathbf{s}) = \|\mathbf{s}\|^2\,\mathbf{I}_{N\times N}.$$

The first space-time block code, which is also the first full rate full diversity orthogonal code, was proposed by Alamouti for a system with two transmit antennas [17]. It is the only existing linear orthogonal STBC design over complex constellation [18]. The scheme is defined by the following transmission matrix, the so called Alamouti matrix [17]:

$$\mathbf{C}(\mathbf{s}) = \begin{pmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{pmatrix} \tag{2.7}$$

In [14], Tarokh *et al.* developed the construction criteria, and studied trade-offs between constellation size, data rates, diversity advantage and complexity. For more than eight transmit antennas, the general design of full diversity orthogonal space-time code was presented in [18]. When the signal constellation $\mathcal{A}$ has all real symbols, such as pulse amplitude modulation (PAM), rate-1 real orthogonal designs for any fixed transmit antenna number are given by the real Hurwitz-Radon families [33]. In [14], a method to construct complex orthogonal designs of rate-1/2 for any fixed transmit antenna number is also provided based on these real rate-1 orthogonal codes; transmitting first a rate-1 orthogonal design with complex symbols, followed by transmitting the same orthogonal design with the symbols complex conjugate, produces a complex rate-1/2 block code. Examples of rate 3/4 complex orthogonal designs for 3 and 4 transmit antennas have been presented in [18][13][19][34]. The rate-7/11 and rate-3/5 generalized complex orthogonal designs for 5 and 6 transmit antennas have also been reported recently. It is also proved that the complex linear processing orthogonal space-time block code design with full-rate full-diversity is impossible for system with more than 2 transmit antennas [18][34]. It means that, if trying to design a complex orthogonal transmission matrix $\mathbf{C}_{N\times N}$ with entries chosen from $\pm s_1$, $\pm s_2$, ..., $\pm s_N$, and their conjugates $\pm s_1^*$, $\pm s_2*$, ..., $\pm s_N*$, or multiples of these by $\pm j$ where $j = \sqrt{-1}$, the design exists if and only if $N = 2$.

To improve the coding rate, a quasi-orthogonal space-time block code with full-rate, partial diversity for a system of four transmit antennas was proposed in [20]. The scheme was defined by the following transmission matrix,

$$\mathbf{C}(\mathbf{s}) = \begin{pmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{pmatrix} = \begin{pmatrix} \mathbf{C}_1 & \mathbf{C}_2 \\ -\mathbf{C}_2^* & \mathbf{C}_1^* \end{pmatrix} \tag{2.8}$$

For a transceiver system with 4 transmitters and M receivers, this kind of code achieves a diversity of $2M$, instead of full diversity $4M$, while maintaining the full-rate transmission.

The performance and diversity property of the code in (2.8) can be improved through constellation rotation. In such scheme, symbols s_3 and s_4 in (2.8) are substituted by their rotated versions, $e^{j\theta}s_3$ and $e^{j\theta}s_4$, respectively. The optimal rotation [21] for the quasi-orthogonal code using QPSK signal is found as $\theta = \pi/6$.

2.3 New Scheme of Full-Rate Full-Diversity Orthogonal Space-Time Block Coding

Although, it has been proved in [18] that orthogonal space-time code of full rate and maximum diversity of $4M$ with *linear* processing does not exist in a system with 4 transmit antennas. The possibility of constructing orthogonal space-time block code by *non-linear* processing has not been studied. Without the linear processing constraint, the full-rate full-diversity orthogonal codes may be found for complex symbols of some special constellations. In this work, it is presented such a design with QPSK symbols transmitted through a system with four transmit antennas. The goal in the design is to achieve full rate and full diversity, while remaining the orthogonality of the code. An other orthogonal full-diversity full-rate STBC for four transmit antennas was presented in [35]. However, the scheme in [35] only can be used in close-loop communication and the transmitter need to know the channel state information (CSI). The proposed scheme is designed for open-loop space-time transmission, so the transmitter need not CSI feedback.

Consider a QPSK based space-time block code for the choice of system parameters: $N = T = K = 4$. The transmission matrix can be constructed as follows

$$\mathbf{C}(\mathbf{s}) = \frac{1}{\sqrt{2}}\begin{pmatrix} \mathbf{C}_1 & \mathbf{C}_2 \\ -\mathbf{C}_1^H\mathbf{C}_2^H\mathbf{C}_1 & \mathbf{C}_1^H \end{pmatrix} \tag{2.9}$$

$$= \frac{1}{\sqrt{2}}\begin{pmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -x_1^* & x_2 & s_1^* & -s_2 \\ -x_2^* & -x_1 & s_2^* & s_1 \end{pmatrix},$$

where $x_1 = \mathcal{R}e\{s_3\} - j\mathcal{I}\text{m}\{2s_1s_2s_4^*\}$, $x_2 = s_1^{*2}s_4 + s_2^2s_4^* + s_1^*s_2s_3 - s_1^*s_2s_3^*$.

Note that the QPSK signal constellation $\mathcal{S}$ in the new scheme is defined as

$$\mathcal{S} = \frac{1}{\sqrt{2}}\left\{e^{j(k\pi/2+\pi/4)}\right\}_{k=0}^{3} \tag{2.10}$$

Since symbol $s_i \in \mathcal{S}$, $i = 1,2,3,4$, the $\mathcal{R}e\{s_3\}$ and $\mathcal{I}\text{m}\{2s_1s_2s_4^*\}$ only take values of either $\frac{1}{2}$ or $-\frac{1}{2}$. Hence, $x_1 = \mathcal{R}e\{s_3\} - j\mathcal{I}m\{2s_1s_2s_4^*\} \in \mathcal{S}$. Similarly it also can be show that $x_2 \in \mathcal{S}$, as follows. Let $s_i = e^{j(\frac{\pi}{2}k_i)}s_1$, where $k_i \in \{0,1,2,3\}$ and $k_1 = 0$, simple manipulation leads to the following form for x_2:

$$\begin{aligned} x_2 &= |s_1|^2\left(s_1^*e^{j\frac{\pi}{2}k_4} + s_1e^{j\frac{\pi}{2}(2k_2-k_4)} + s_1e^{j\frac{\pi}{2}(k_2+k_3)}\right. \\ &\quad \left.-s_1^*e^{j\frac{\pi}{2}(k_2-k_3)}\right) \\ &= \frac{1}{2}e^{j\frac{\pi}{2}k_2}(s_1^*e^{j\frac{\pi}{2}(k_4-k_2)} + s_1e^{j\frac{\pi}{2}(k_2-k_4)} + s_1e^{j\frac{\pi}{2}k_3} \\ &\quad -s_1^*e^{-j\frac{\pi}{2}k_3}) \\ &= e^{j\frac{\pi}{2}k_2}\left(Re\left\{s_1e^{j\frac{\pi}{2}(k_2-k_4)}\right\} + jIm\left\{s_1e^{j\frac{\pi}{2}k_3}\right\}\right) \end{aligned} \tag{2.11}$$

Therefore, the coding process does not expand the QPSK constellation.

Since x_1 =Re$\{s_3\} - j$Im$\{2s_1s_2s_4^*\}$, and s_3 are independent of $s_1s_2s_4^*$, to avoid the expansion of constellation, signal formed by any combination of possible real part and imaginary part of symbols from certain constellation should still belong to that constellation. Hence, it is concluded that QPSK signal is the *only possible* complex MPSK signal, that has the full-rate orthogonal design by the scheme. It is worthwhile to point out that for BPSK constellation, the new design scheme naturally leads to $s_i = \pm\sqrt{2}/2$, $x_1 = s_3$, and $x_2 = s_4$. Therefore, the transmission matrix in (2.9) simply reduces to the orthogonal code matrix with linear processing developed in [18]

Using computer simulation, the rank conditions of all the possible matrices $\mathbf{C}(\mathrm{s} - \widetilde{\mathrm{s}})$ is examined. It is found that all possible matrices of $\mathbf{C}(\mathrm{s} - \widetilde{\mathrm{s}})$ are of full rank. Hence, the newly proposed space-time block code indeed provides full transmit diversity over a quasi-static fading channel [14]. The same conclusion can also be drawn from the slopes of the BER v.s. SNR curve in Figure 2.4 and 2.5.

Since the new scheme is also a unit-energy design, to reduce decoding complexity, Equation (2.6) can be used in ML detection instead of using (2.5). In some application neither the receiver nor the transmitter has the channel state information (CSI). To solve this problem, based on orthogonal designs, Tarokh and Jafarkhani proposed a differential space-time coding (DSTC) scheme by using two transmitter antennas [36], which provides simple encoding/decoding algorithms for single differential detection. In [37], Jafarkhani presented a generalized differential coding and decoding scheme for OSTBC. Since OSTBC coding matrix may not be square-matrix, the differential OSTBC scheme is very trivial in [37], in order to obtain general results. Hughes [38] , and Hochwald and Sweldens [39] independently proposed differential unitary space-time modulation (DUSTM), in which the transmitted signal matrix at each time block is the product of the previously transmitted signal matrix and the current unitary data matrix. To adopt DUSTM techniques, the STBC coding matrix should be unitary square-matrix. Since the new proposed full-rate full-diversity STBC has an unitary square coding matrix, it can be easily applied in differential space-time coding scheme. The differential encoding can be conveniently implemented by

$$\mathbf{G}(\mathbf{s})_t = \mathbf{C}(\mathbf{s})_t \mathbf{G}(\mathbf{s})_{t-1} \tag{2.12}$$

where $\mathbf{G}(\mathbf{s})_t$ is the unitary transmission matrix at t' th block time, and $\mathbf{C}_t$ is the code matrix at t'th block time given by (2.9). The initialization of $\mathbf{G}(\mathbf{s})_{t-1}$ can be arbitary orthogonal code matrix described by (2.9). Assuming the channel vector of $\mathbf{h}$ is constant within two consecutive block durations and only one receiving antenna is used, the received data of 2 consecutive block time can be formulated as:

$$\begin{aligned} \mathbf{y}_{t-1} &= \sqrt{\frac{\rho}{N}}\mathbf{G}_{t-1}\mathbf{h} + \mathbf{v}_{t-1} \\ \mathbf{y}_t &= \sqrt{\frac{\rho}{N}}\mathbf{C}_t\mathbf{G}_{t-1}\mathbf{h} + \mathbf{v}_t \end{aligned} \tag{2.13}$$

If the noise is white Gaussian noise (AWGN), the ML estimate of $\sqrt{\frac{\rho}{N}}\mathbf{G}_{t-1}\mathbf{h}$ is $\mathbf{y}_{t-1}$. At t'th block, use $\mathbf{y}_{t-1}$ as the estimate of equivalent channel $\widetilde{\mathbf{h}}_t = \sqrt{\frac{\rho}{N}}\mathbf{G}_{t-1}\mathbf{h}$ to decode $\mathbf{C}_t$ coherently. It should be noted that the estimate of $\widetilde{\mathbf{h}}_t$ does not depends on the decoding results of $\mathbf{C}_{t-1}$, so there is no error propagation. When using $\mathbf{y}_{t-1}$ to decode $\mathbf{C}_t$ coherently, (2.13) can be rewritten into

$$\begin{aligned} \mathbf{y}_t &= \mathbf{C}_t\left(\widetilde{\mathbf{h}}_t + \mathbf{v}_{t-1}\right) + \mathbf{v}_t \\ &= \mathbf{C}_t\widetilde{\mathbf{h}}_t + \widetilde{\mathbf{v}}_t \end{aligned} \tag{2.14}$$

where $\widetilde{\mathbf{v}}_t = \mathbf{C}_t\mathbf{v}_{t-1} + \mathbf{v}_t$. Because $\mathbf{C}_t$ is unitary and $\mathbf{v}_t$ are white Gaussian noise, it is obtain $\widetilde{\mathbf{v}}_t\widetilde{\mathbf{v}}_t^H = 2\sigma^2\mathbf{I}$, which is still white Gaussian noise. From the unitary property of $\mathbf{G}_{t-1}$, the element of $\widetilde{\mathbf{h}}_t$ is still complex Gaussian distributed and statistically equivalent to $\mathbf{h}$.Comparing (2.14) to (2.8), it is found that the signal power are same but the noise power is two times in differential detection than in coherent detection. So 3dB loss occurs when differential scheme is used.

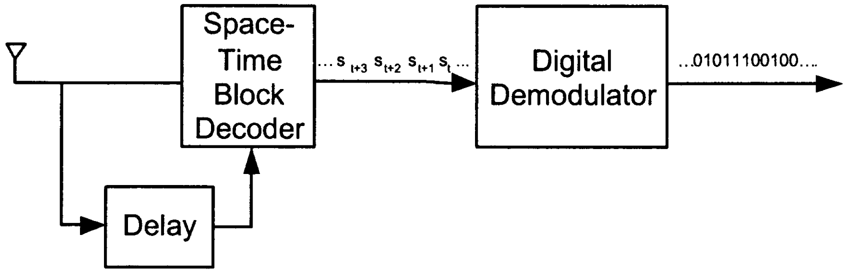

Figure 2.2 Structure of differential STBC transmitter.

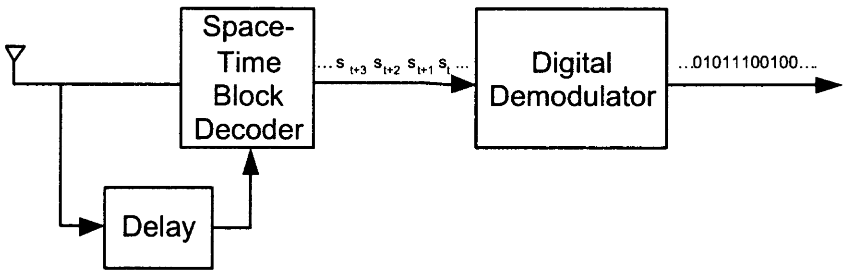

Figure 2.3 Structure of differentail STBC receiver.

2.4 Simulation Results

In this section, simulation results on performance study for the newly proposed space-time code in (2.9) are provided. It is also compared with the orthogonal code in [18] and quasi-orthogonal codes presented in [20][21], respectively. It is noticed that the comparison to 3/4 rate orthogonal code won't be fair, since it is impossible to find a signal constellation such that a 3/4 rate code will deliver date rate of 2/bits/s/Hz. Hence, the newly proposed code is compared to the 1/2-rate orthogonal, the full rate quasi-orthogonal and the improved quasi-orthogonal space-time block codes under the baseline of the same data rate.

Figure 2.4 provides simulation results for the transmission of 2 bits/s/Hz by four transmit antennas and one receiving antenna using QPSK modulated full-rate full-diversity orthogonal code, quasi-orthogonal code, constellation rotated quasi-orthogonal code, and 16 QAM modulated half rate full-diversity orthogonal

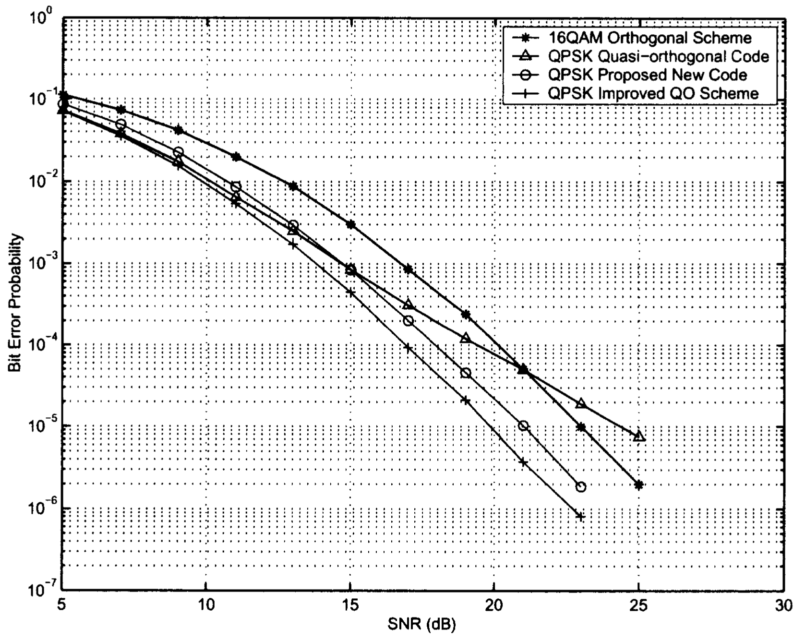

Figure 2.4 Performance of various space-time block codes at 2 bits/s/Hz. 4 transmit antennas and 1 receive antenna are used.

code. Simulation results from Figure 2.4 show that the performance of the proposed code is close to that of the quasi-orthogonal code and is better than that of the half-rate orthogonal code when signal to noise ratio (SNR) is low. As the SNR increases, codes with full diversity work better and benefit more from SNR increase than partial diversity code. The new space-time coding scheme demonstrates a performance gain about 2 dB at the bit error rate (BER) as 10^{-3}, compared to the half-rate orthogonal code. The fact that the BER-SNR curves corresponding to the full-diversity half-rate code, constellation rotated quasi-orthogonal code, and the newly proposed code all have the same slope at high SNR, indicates that these three schemes all provide full diversity. Figure 2.4 shows that the performance of the proposed code is only 1 dB worse than that of the constellation rotated quasi-orthogonal code when only one receiving antenna is used.

Figure 2.5 provides simulation results for a transceiver system with 4 transmit antennas and two receiving antennas using the proposed QPSK based full-rate (2

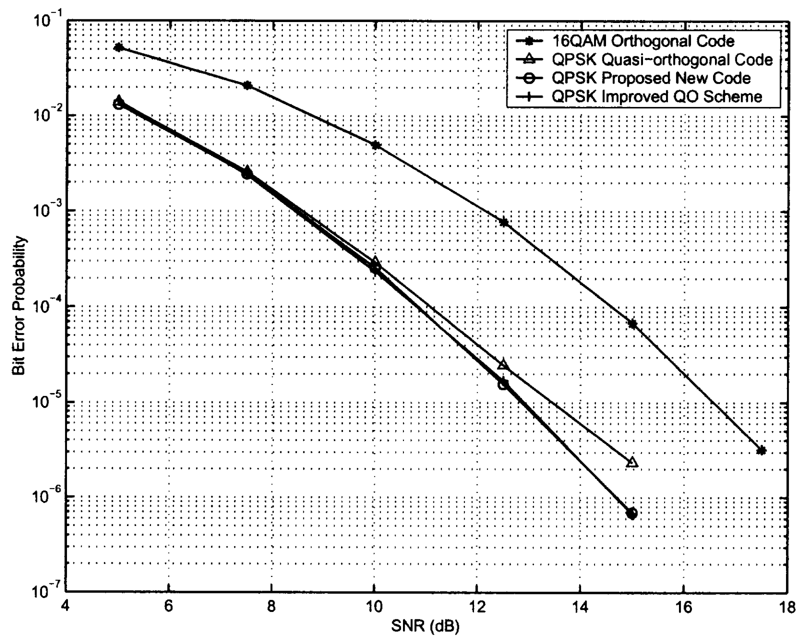

Figure 2.5 Performance of various space-time block codes at 2 bits/s/Hz. 4 transmit antennas and 2 receive antennas are used.

bits/s/Hz) full-diversity code, the full-rate quasi-orthogonal code, the constellation rotated quasi-orthogonal code, and the 16 QAM modulated 1/2 rate orthogonal code, respectively. When two receiving antennas are used, the proposed code outperforms the quasi-orthogonal code at a even lower SNR (7.5 dB), and a 4 dB performance gain is achieved at the bit error rate (BER) of 10^{-3}, compared to the half rate orthogonal code. It is interesting to notice that the performance gap between the newly proposed code and improved quasi-orthogonal code observed in one receiving antenna system disappears.

Figure 2.6 shows the performance of differential scheme of the proposed new full-rate full-diversity STBC, when CSI is not available to receiver. Compared to the coherent scheme, the differential scheme still provide the full-diversity gain and 3dB degradation is observed.

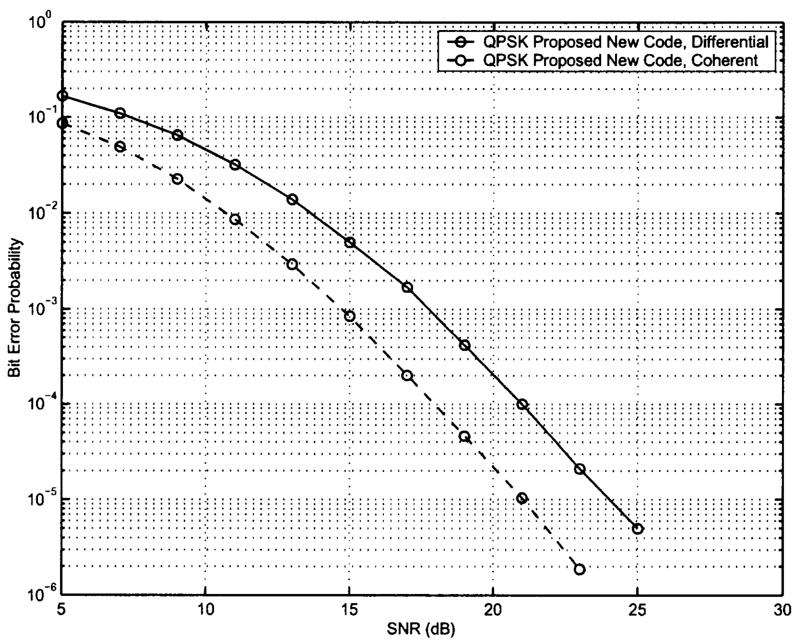

Figure 2.6 Performance of the new proposed full-rate full-diversity STBC, with coherent and differential detection, respectively.

2.5 Conclusions

This work demonstrated the existence of a full-rate full-diversity space-time orthogonal coding scheme for a QPSK system with four transmit antennas.

It should be pointed out that the receiver of the half-rate full-diversity codes can decode the symbols one by one, and that of the full-rate half-diversity quasi-orthogonal codes can decode the symbols pair by pair [20]. This means that the full-rate full-diversity potential of the proposed space-time coding scheme is achieved at the cost of increased decoding complexity. For QPSK symbols, the decoding complexity of new orthogonal code is nearly 8 times of that of the quasi-orthogonal codes. The encoding complexity of the proposed orthogonal code is only a little higher than that of the other two.

Because of the unitary property of the new proposed STBC, it is easy to extend the coherent scheme to differential scheme. In this chapter, it has been proved that a unitary coding matrix achieving full space diversity for systems with CSI at receiver also achieves full space diversity for system without CSI at receiver by means of DUSTM. A loss of about 3 dB in performance is observed in differential scheme, compared to coherent detection.

CHAPTER 3

FAST MAXIMUM LIKELIHOOD DECODING OF QUASI-ORTHOGONAL CODES

In this chapter, it is presented a new fast maximum likelihood (ML) detection algorithm for quasi-orthogonal space-time block codes (QO-STBC). Observed in a transformed signal space, the QO-STBC encoder embeds information into a group of orthogonal subspaces of rank two. It is proposed a fast technique that separates information embedded in each rank-two subspace by a simple linear processing, and decodes the separated information independently. For a multiple-input and multiple-output (MIMO) system, decoding algorithm decomposes the original $N \times N$ system into $N/2$ parallel 2×2 MIMO sub-systems without information loss followed by MRC. It then detects symbols contained in each 2×2 MIMO sub-system using a sphere decoder of reduced computational complexity.

3.1 System Model

It is assumed that there are N transmit and M receiving antennas in the system. The information bits are first mapped to constellation symbols by digital modulation. A space-time block encoder encodes a vector of input symbols of length K, $\mathbf{s} = (s_1, s_2, \cdots, s_K)$, into a $T \times N$ transmit matrix $\mathbf{C}(\mathbf{s})$, where T is the block length and the code rate is hence K/T. The received signal can be modeled as

$$\mathbf{Y} = \sqrt{\frac{\rho}{N}}\, \mathbf{C}(\mathbf{s})\ \mathbf{H} + \mathbf{V}, \tag{3.1}$$

where $\mathbf{H}$ is the $N \times M$ channel matrix; $\mathbf{Y}$ and $\mathbf{V}$ are the $T \times M$ data and noise matrices. The entries of the normalized matrices $\mathbf{H}$ and $\mathbf{V}$ are mutually independent, zero mean, and circularly symmetric complex Gaussian variables of unit variance; and ρ is The SNR per receiving antenna.

Given the channel matrix $\mathbf{H}$, the traditional ML detection algorithm estimates $\mathbf{C}$ from $\mathbf{Y}$ as follows,

$$\widehat{\mathbf{C}} = \arg\min_{\mathbf{C}\in\mathcal{C}(\mathbf{s})} \left\| \mathbf{Y} - \sqrt{\frac{\rho}{N}}\mathbf{C}\ \mathbf{H} \right\|^2 . \tag{3.2}$$

That is, the ML detector considers all possible input $\mathbf{C}$ induced by $\mathbf{s}$, and chooses the input which minimizes the Euclidean distance.

3.2 Orthogonal Space-Time Block Coding and Decoding

In this section, the coding and decoding theory of orthogonal space-time block code is reviewed. In [18], the orthogonal space-time block codes are defined by

$$\mathbf{C}^H(\mathbf{s})\mathbf{C}(\mathbf{s}) = \|\mathbf{s}\|^2 \mathbf{I}_{N\times N}.$$

The only existing full-rate complex orthogonal space-time block code, which has a decoupled ML decoding with linear complexity, is Alamouti code[18]. The coding scheme of Alamouti code for vector $\mathbf{s} = (s_1, s_2)$ is defined by the following transmission matrix, [17]:

$$\mathbf{C}(\mathbf{s}) = \begin{pmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{pmatrix} \tag{3.3}$$

At a given receiving antenna, the received data within one code block of 2 consecutive time slots are described by

$$\begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \sqrt{\frac{\rho}{N}} \begin{pmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{pmatrix} \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \tag{3.4}$$

The data in (3.4) can be re-arranged as follows,

$$\underbrace{\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix}}_{\tilde{\mathbf{y}}} = \sqrt{\frac{\rho}{N}} \underbrace{\begin{pmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{pmatrix}}_{\tilde{\mathbf{H}}} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \underbrace{\begin{bmatrix} v_1 \\ v_2^* \end{bmatrix}}_{\tilde{\mathbf{v}}} \tag{3.5}$$

Due to the orthogonality of $\tilde{\mathbf{H}}$ in (3.5), the decoupling of s_1 and s_2 can be accomplished by a simple matched filtering, i.e.

$$\begin{aligned} \underline{\mathbf{y}} &= \tilde{\mathbf{H}}^H \tilde{\mathbf{y}} \\ &= \sqrt{\frac{\rho}{N}} \tilde{\mathbf{H}}^H \tilde{\mathbf{H}} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \tilde{\mathbf{H}}^H \tilde{\mathbf{v}} \\ &= \sqrt{\frac{\rho}{N}} \left(|h_1|^2 + |h_2|^2\right) \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \underline{\tilde{\mathbf{v}}} \end{aligned}$$

Since $\tilde{\mathbf{H}}$ is a orthogonal matrix, the elements of the new noise vector $\underline{\tilde{\mathbf{v}}}$ are identical independent (iid) white Gaussian noise. The ML of s_i is simply

$$\hat{s}_i = Q(\underline{y}_i),$$

where $Q(\cdot)$ represents the slicing operation. If s_i is a symbol mapped by b bits, instead of searching 2^{2b} possible signal vectors and calculating corresponding Euclidean distance by (3.2), the decoupled ML decoding algorithm find the ML estimate by slicing the output data of a simple linear processor. Its computational complexity is much lower than that of the regular pair-based exhaustive searching algorithm.

From the fast ML decoding inherent in the orthogonal space-time block code, it can be observed that in a transformed signal space (here defined by $[y_1 \;\; y_2^*]^T$), the information or symbol s_i is embedded into two *orthogonal* subspaces spanned by $[h_1 \;\; h_2^*]^T$ and $[h_2 \;\; -h_1^*]^T$, respectively. Therefore, the information symbols of s_1

and s_2 are perfectly separated by a simple linear processing, that delivers the ML detection performance.

Extending the above decoupled fast ML decoding algorithm for system with 2 transmit antennas and M receiving antennas, a maximum ratio combine (MRC) is simply used after the linear processed data, prior to symbol decision.

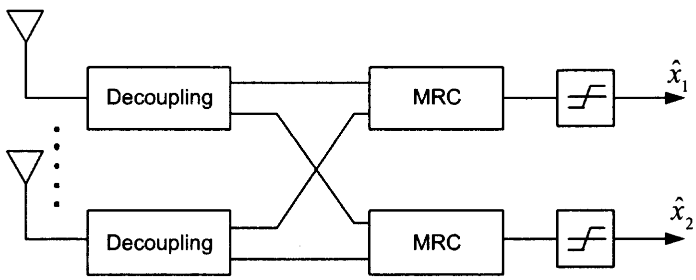

Figure 3.1 The structure of decoupled ML decoder for orthogonal space-time block code.

3.3 Full Rate Quasi-orthogonal Space-Time Code

The full-rate complex orthogonal space-time block coding with fast ML linear decoding feature, only can be realized in a system with two transmit antenna. To improve the coding rate, a *quasi-orthogonal* space-time block code of 4-transmit antennas was proposed in [20]. The scheme was defined by the following transmission matrix,

$$\mathbf{C}(\mathbf{s}) = \begin{pmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ \underbrace{s_4}_{\mathbf{c}_1} & \underbrace{-s_3}_{\mathbf{c}_2} & \underbrace{-s_2}_{\mathbf{c}_3} & \underbrace{s_1}_{\mathbf{c}_4} \end{pmatrix} = \begin{pmatrix} \mathbf{C}_1 & \mathbf{C}_2 \\ -\mathbf{C}_2^* & \mathbf{C}_1^* \end{pmatrix} \tag{3.6}$$

Since the $\mathbf{C}(\mathbf{s})$ is not an orthogonal matrix, the code represented by (3.6) is called quasi-orthogonal code. Note the fact that the subspace $\langle \mathbf{c}_1, \ \mathbf{c}_4 \rangle$ spanned by $\mathbf{c}_1$ and

$\mathbf{c}_4$ is orthogonal to the subspace $\langle \mathbf{c}_2, \ \mathbf{c}_3 \rangle$ spanned by $\mathbf{c}_2$ and $\mathbf{c}_3$, where $\mathbf{c}_i$ denotes the i-th column of matrix $\mathbf{C}(\mathbf{s})$.

In [20] and [21], it has been pointed out that the decoding of quasi-orthogonal codes can be done by searching symbols pair by pair, but the decoding method reported is trivial and the symbol pair exhaustive searching is used for ML detection. As larger size constellation is used in modulation, the searching complexity increase exponentially. Therefore, the pair-wise exhaustive searching algorithm is still computa-tionally intensive for applications in real time high rate data transmission.

In this chapter, a new fast ML decoding algorithm for quasi-orthogonal codes is presented. By a simple linear processing, the proposed decoding algorithm decomposes the original $N \times N$ MIMO system into $N/2$ parallel 2×2 MIMO sub-systems without information loss. Information contained in each 2×2 MIMO system is detected by sphere decoder independently. The computational complexity of proposed decoding algorithm is independent of the constellation size used in digital modulation.

3.3.1 Linear Decoupling of Quasi-orthogonal Space-Time Block Code

Using the similar techniques for analyzing orthogonal code, the quasi-orthogonal space-time block coding system is studied in a *transformed* signal space defined by $\widetilde{\mathbf{y}} = [y_1 \ \ y_2^* \ \ y_3^* \ \ y_4]^T$, where y_i's, $(i = 1, \cdots, 4)$ are data received at a given receiving antenna over four consecutive time slots. The equivalent system to (3.1) can be described as

$$\widetilde{\mathbf{y}} = \sqrt{\frac{\rho}{N}}\widetilde{\mathcal{H}}\mathbf{s} + \widetilde{\mathbf{v}} \ , \tag{3.7}$$

where $\mathbf{s} = [s_1 \;\; s_2 \;\; s_3 \;\; s_4]^T$

$$\begin{aligned}
\widetilde{\mathcal{H}} &= \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \\
\widetilde{\mathbf{y}} &= \begin{bmatrix} y_1 & y_2^* & y_3^* & y_4 \end{bmatrix}^T \\
\widetilde{\mathbf{v}} &= \begin{bmatrix} \eta_1 & \eta_2^* & \eta_3^* & \eta_4 \end{bmatrix}^T
\end{aligned} \tag{3.8}$$

Using $\widetilde{\mathbf{h}}_i$ to denote the i-th column of $\widetilde{\mathcal{H}}$, It is found that the subspace spanned by $\widetilde{\mathbf{h}}_1$ and $\widetilde{\mathbf{h}}_4$, is orthogonal to the subspace spanned by $\widetilde{\mathbf{h}}_2$ and $\widetilde{\mathbf{h}}_3$. Utilizing the subspace orthogonality, then reformulate (3.7) as follows,

$$\widetilde{\mathbf{y}} = \sqrt{\frac{\rho}{N}}\underline{\mathcal{H}}\;\underline{\mathbf{s}} + \widetilde{\mathbf{v}} \tag{3.9}$$

where $\underline{\mathbf{s}} = \frac{1}{2}\left[s_1 + s_4 \;\; s_2 + s_3 \;\; s_2 - s_3 \;\; s_1 - s_4\right]^T$,

and $\underline{\mathcal{H}} = \left[\widetilde{\mathbf{h}}_1 + \widetilde{\mathbf{h}}_4 \;\; \widetilde{\mathbf{h}}_2 + \widetilde{\mathbf{h}}_3 \;\; \widetilde{\mathbf{h}}_2 - \widetilde{\mathbf{h}}_3 \;\; \widetilde{\mathbf{h}}_1 - \widetilde{\mathbf{h}}_4\right]$.

Then proceed to find a simple linear processing scheme to separated the informat-ion embedded in subspaces, $\left\langle \widetilde{\mathbf{h}}_1, \;\; \widetilde{\mathbf{h}}_4 \right\rangle$ and $\left\langle \widetilde{\mathbf{h}}_2, \;\; \widetilde{\mathbf{h}}_3 \right\rangle$. In doing so, it is noticed that a unitary matrix $\underline{\mathcal{H}}_{\mathrm{U}}$ can be constructed from $\underline{\mathcal{H}}$ simply by normalizing its columns, i.e.

$$\underline{\mathcal{H}}_{\mathrm{U}} = \begin{bmatrix} \dfrac{\underline{\mathbf{h}}_1}{\|\underline{\mathbf{h}}_1\|} & \dfrac{\underline{\mathbf{h}}_2}{\|\underline{\mathbf{h}}_2\|} & \dfrac{\underline{\mathbf{h}}_3}{\|\underline{\mathbf{h}}_3\|} & \dfrac{\underline{\mathbf{h}}_4}{\|\underline{\mathbf{h}}_4\|} \end{bmatrix}$$

where $\underline{\mathbf{h}}_i$ denotes the i-th column of $\underline{\mathcal{H}}$. Mapping vector $\widetilde{\mathbf{y}}$ using this unitary matrix (decoupled match filter banks), it is obtain

$$\begin{aligned}
\underline{\mathbf{y}} &= \underline{\mathcal{H}}_{\mathrm{U}}^{H}\,\widetilde{\mathbf{y}} \\
&= \sqrt{\frac{\rho}{N}}\begin{bmatrix} |\underline{\mathbf{h}}_1| & 0 & 0 & 0 \\ 0 & |\underline{\mathbf{h}}_2| & 0 & 0 \\ 0 & 0 & |\underline{\mathbf{h}}_3| & 0 \\ 0 & 0 & 0 & |\underline{\mathbf{h}}_4| \end{bmatrix} \underline{\mathbf{s}} + \underline{\mathbf{v}}
\end{aligned} \tag{3.10}$$

where $\|\underline{\mathbf{h}}_1\|^2 = \|\underline{\mathbf{h}}_3\|^2 = 2\left(|h_1 + h_4|^2 + |h_2 - h_3|^2\right)$,
$\|\underline{\mathbf{h}}_2\|^2 = \|\underline{\mathbf{h}}_4\|^2 = 2\left(|h_1 - h_4|^2 + |h_2 + h_3|^2\right)$ and $\underline{\mathbf{v}} = \underline{\mathcal{H}}_{\mathrm{U}}\widetilde{\mathbf{v}}$. Since $\underline{\mathcal{H}}_{\mathrm{U}}$ is a unitary matrix, $\underline{\mathbf{v}}$ and $\widetilde{\mathbf{v}}$ have the same statistical characters. From (3.10), it is observed that the symbol pair (s_1, s_4) and symbol pair (s_2, s_3) are decoupled and can be ML decoded independently as follows:

$$\begin{aligned}
(\widehat{s}_1, \widehat{s}_4) &= \arg\min_{s_1,s_4\in\mathcal{S}}\left\{\left|\underline{y}_1 - \frac{1}{2}\sqrt{\frac{\rho}{N}}\,\|\underline{\mathbf{h}}_1\|\,(s_1 + s_4)\right|^2\right. \\
&\quad \left. + \left|\underline{y}_4 - \frac{1}{2}\sqrt{\frac{\rho}{N}}\,\|\underline{\mathbf{h}}_4\|\,(s_1 - s_4)\right|^2\right\} \\
(\widehat{s}_2, \widehat{s}_3) &= \arg\min_{s_2,s_3\in\mathcal{S}}\left\{\left|\underline{y}_2 - \frac{1}{2}\sqrt{\frac{\rho}{N}}\,\|\underline{\mathbf{h}}_2\|\,(s_2 + s_3)\right|^2\right. \\
&\quad \left. + \left|\underline{y}_3 - \frac{1}{2}\sqrt{\frac{\rho}{N}}\,\|\underline{\mathbf{h}}_3\|\,(s_2 - s_3)\right|^2\right\}
\end{aligned} \tag{3.11}$$

To extend (3.11) to system with M receiving antennas, it is obtained

$$\begin{aligned}
(\widehat{s}_i, \widehat{s}_j) &= \arg\min_{s_i,s_j\in\mathcal{S}}\left\{\sum_{m=1}^{M}\left|\underline{y}_{i,m} - \frac{1}{2}\sqrt{\frac{\rho}{N}}\,\|\underline{\mathbf{h}}_{i,m}\|\,(s_i + s_j)\right|^2\right. \\
&\quad \left. + \left|\underline{y}_{j,m} - \frac{1}{2}\sqrt{\frac{\rho}{N}}\,\|\underline{\mathbf{h}}_{j,m}\|\,(s_i - s_j)\right|^2\right\}
\end{aligned}$$

where (i, j) is either $(1, 4)$ or $(2, 3)$. The quantities $\underline{y}_{i,m}$, and vector $\underline{\mathbf{h}}_{i,m}$ are the data sample $\underline{y}_i$, and the channel vector $\underline{\mathbf{h}}_i$ defined above associated with the m-th receiving antenna.

3.3.2 Sphere Decoder for Output of Decoupling

When higher order modulation is involved, the complexity of symbol pairs based exhaustive searching in (3.11) increases exponentially. To further reduce decoding complexity, it is proposed a sphere decoding scheme, which is well suitable for decoding QAM modulated quasi-orthogonal coding system. The complexity of the decoding algorithm is independent of the modulation order, hence it is very useful for high data rate transmission. Sphere decoder has been applied in decoding of V-BLAST space-time system [29]. Study on sphere decoding algorithm of Fincke and Pohst shows that for a wide range of noise variances the expected complexity is polynomial, in fact often roughly cubic [29].

In the proposed decoding scheme, the data at each receiving antenna is processed by the proposed linear decoupling scheme in (3.10). The output of all decoupler from all receiving branches are then combined using the MRC technique. The results obtained after MRC can be written in a pair-wise format as follows:

$$\underbrace{\begin{bmatrix} r_i \\ r_j \end{bmatrix}}_{\mathbf{r}(i,j)} = \underbrace{\begin{bmatrix} \alpha_i & e^{j\theta}\alpha_i \\ \alpha_j & -e^{j\theta}\alpha_j \end{bmatrix}}_{\mathbf{M}(i,j)} \underbrace{\begin{bmatrix} s_i \\ s_j \end{bmatrix}}_{\mathbf{s}(i,j)} + \eta \tag{3.12}$$

where $r_i = \|\mathbf{w}_i\|^{-1} \mathbf{w}_i^T \begin{bmatrix} \underline{y}_{i,1} & \cdots & \underline{y}_{i,M} \end{bmatrix}^T$, $\alpha_i = \|\mathbf{w}_i\|$ and $\mathbf{M}$ is the matrix defined the lattice,. The MRC weighting vector of $\mathbf{w}_i = \left[\|\underline{\mathbf{h}}_{i,1}\| \quad \cdots \quad \|\underline{\mathbf{h}}_{i,M}\|\right]^T$. η is the noise vector, whose elements are iid complex white Gaussian random variables. When the constellation at transmitter 3 and 4 are not rotated, $\theta = 0$. The operation in (3.12) is similar to what used in a 2×2 V-BLAST system, which can be decoded by a sphere decoder proposed in [40][29]. The symbols contained in (3.7) can also be decoded as a 4×4 V-BLAST system by sphere decoding algorithm, but the complexity will be more than 4^3 times higher than that for decoding (3.12).

3.3.3 Diversity Discussion

Assuming the knowledge of channel state information (CSI), the probability of transmitting $\mathbf{c} = (s_1, s_4)$and deciding in favor of $\mathbf{e} = (s_1^{'}, s_4^{'})$ at the decoder can be bounded by the following [32],

$$P(\mathbf{c} \to \mathbf{e}|\mathbf{H}) \leq \exp(-d^2(\mathbf{c},\mathbf{e})/2\sigma_v^2) \tag{3.13}$$

where σ_v^2 is the variance of the noise and

$$d^2(\mathbf{c},\mathbf{e}) = \frac{1}{2}\sqrt{\frac{\rho}{N}}\sum_{m=1}^{M}\left(\left\|\underline{\mathbf{h}}_{1,m}\right\|^2\Delta_1^2 + \left\|\underline{\mathbf{h}}_{4,m}\right\|^2\Delta_4^2\right) \tag{3.14}$$

Define $\Delta_1 = \left|(s_1 + s_4) - (s_1^{'} + s_4^{'})\right|$ and

$\Delta_4 = \left|(s_1 - s_4) - (s_1^{'} - s_4^{'})\right|$. Because $\left\|\underline{\mathbf{h}}_{i,m}\right\|^2$'s are independent χ^2 distributed with 4 degree of freedom, if and only if for any $\mathbf{c}$ and $\mathbf{e}$, both Δ_1 or Δ_4 are greater than 0, the quasi-orthogonal code can provide full diversity of $4M$.

In this paper, the constellation rotation technique proposed in [21] and [41] is adopted to make both Δ_1 and Δ_4 non-zero. That means the s_3 and s_4 are substituted with $e^{j\theta}s_3$ and $e^{j\theta}s_4$ in the following code matrix (3.12). However, it should be noted that constellation rotation may not be the only way to obtain full diversity and optimize coding advantage for quasi-orthogonal code.

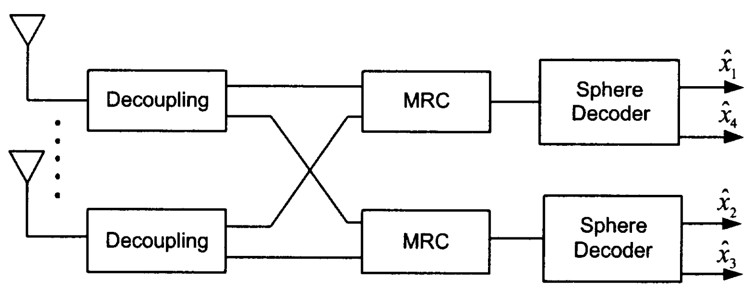

Figure 3.2 Structure of fast ML decoder for quasi-orthogonal space-time block code.

3.4 3/4 Rate Quasi-orthogonal Code for 6 Transmit Antenna System

In [20], a 3/4 rate quasi-orthogonal code is proposed for system with 8 transmit antennas. Since the code in [20] can not be decoded using the newly proposed decoupled fast ML decoding algorithm, it is proposed a new 3/4 rate orthogonal code using 6 transmit antennas, and can be decoded by above proposed fast ML decoding algorithm. The new code is represented by following transmission matrix, which is constructed by taking the 1,2,3,5,6 and 7 columns of the 3/4 rate coding matrix in [20],

$$\mathbf{C}(\mathbf{s}) = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 & s_5 & s_6 \\ -s_2^* & s_1^* & 0 & s_5^* & -s_4^* & 0 \\ s_3^* & 0 & -s_1^* & -s_6^* & 0 & s_4^* \\ 0 & -s_3^* & s_2^* & 0 & s_6^* & -s_5^* \\ -s_4 & -s_5 & -s_6 & s_1 & s_2 & s_3 \\ -s_5^* & s_4^* & 0 & -s_2^* & s_1^* & 0 \\ s_6^* & 0 & -s_4^* & s_3^* & 0 & -s_1^* \\ 0 & s_6^* & -s_5^* & 0 & s_3^* & -s_2^* \end{bmatrix} \tag{3.15}$$

Using the similar techniques and assuming M=1, denote the data vector and noise vector over one block duration of 8 consecutive time slots by $\mathbf{y} = \begin{bmatrix} y_1 & y_2 & \cdots & y_8 \end{bmatrix}^T$ and$\mathbf{v} = \begin{bmatrix} v_1 & v_2 & \cdots & v_8 \end{bmatrix}^T$, respectively.

The system can modeled as

$$\mathbf{y} = \sqrt{\frac{P}{N}}\mathbf{C}(\mathbf{s})\mathbf{h} + \mathbf{v},$$

Then the above system is equivalent to system written as follows

$$\widetilde{\mathbf{y}} = \sqrt{\frac{P}{N}}\widetilde{\mathbf{H}}\ \mathbf{s} + \widetilde{\mathbf{v}},$$

$$\widetilde{\mathbf{H}} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 & h_5 & h_6 \\ h_2^* & -h_1^* & 0 & -h_5^* & h_4^* & 0 \\ -h_3^* & 0 & h_1^* & h_6^* & 0 & -h_4^* \\ 0 & h_3^* & -h_2^* & 0 & -h_6^* & h_5^* \\ h_4 & h_5 & h_6 & -h_1 & -h_2 & -h_3 \\ h_5^* & -h_4^* & 0 & h_2^* & -h_1^* & 0 \\ -h_6^* & 0 & h_4^* & -h_3^* & 0 & h_1^* \\ 0 & -h_6^* & h_5^* & 0 & -h_3^* & h_2^* \end{bmatrix}$$

$$\widetilde{\mathbf{y}} = \begin{bmatrix} y_1 & y_2^* & y_3^* & y_4^* & y_5 & y_6^* & y_7^* & y_8^* \end{bmatrix}^T$$

$$\widetilde{\mathbf{v}} = \begin{bmatrix} v_1 & v_2^* & v_3^* & v_4^* & v_5 & v_6^* & v_7^* & v_8^* \end{bmatrix}^T$$

Omit the trivial derivation of decoupler and lattice matrix for the 3/4 rate code of (3.15) and just give out the results,

$$\begin{aligned} \underline{\mathbf{s}} &= \frac{1}{2}[-js_1 + s_4 \quad -js_2 + s_5 \quad -js_3 + s_6 \\ &\quad s_1 - js_4 \quad s_2 - js_5 \quad s_3 - js_6]^T \\ \underline{\mathcal{H}} &= [j\widetilde{\mathbf{h}}_1 + \widetilde{\mathbf{h}}_4 \quad j\widetilde{\mathbf{h}}_2 + \widetilde{\mathbf{h}}_5 \quad j\widetilde{\mathbf{h}}_3 + \widetilde{\mathbf{h}}_6 \\ &\quad \widetilde{\mathbf{h}}_1 + j\widetilde{\mathbf{h}}_4 \quad \widetilde{\mathbf{h}}_2 + j\widetilde{\mathbf{h}}_5 \quad \widetilde{\mathbf{h}}_3 + j\widetilde{\mathbf{h}}_6]^T \\ \widetilde{\mathbf{y}} &= \begin{bmatrix} y_1 & y_2^* & y_3^* & y_4^* & y_5 & y_6^* & y_7^* & y_8^* \end{bmatrix}^T \end{aligned}$$

The ML estimation statistic for symbol pair (s_i, s_{i+3}), $i = 1, 2, 3$ is

$$\begin{aligned} (\widehat{s}_i, \widehat{s}_{i+3}) &= \arg \min_{s_i, s_{i+3} \in \mathcal{S}} \left| 2\underline{y}_i - \left\| \underline{\mathbf{h}}_i \right\| (-js_i + s_{i+3}) \right|^2 \\ &\quad + \left| 2\underline{y}_{i+3} - \left\| \underline{\mathbf{h}}_{i+3} \right\| (s_i - js_{i+3}) \right|^2 \end{aligned}$$

where

$$\begin{aligned}
\|\underline{\mathbf{h}}_1\|^2 &= \|\underline{\mathbf{h}}_2\|^2 = \|\underline{\mathbf{h}}_3\|^2 \\
&= 2\left(|jh_1 + h_4|^2 + |jh_2 + h_5|^2 + |jh_3 + h_6|^2\right) \\
\|\underline{\mathbf{h}}_4\|^2 &= \|\underline{\mathbf{h}}_5\|^2 = \|\underline{\mathbf{h}}_6\|^2 \\
&= 2\left(|h_1 + jh_4|^2 + |h_2 + jh_5|^2 + |h_3 + jh_6|^2\right)
\end{aligned}$$

3.5 Simulation Results

Figure 3.3 provides performance of different space-time block coding schemes, which have fast ML decoding, at 2 bit/s/Hz data rate. The same curve slopes at high SNR indicate the same order of diversity for the corresponding encoding schemes. The constellation rotated and un-rotated quasi-orthogonal code shows *different* diversity property. The constellation rotated quasi-orthogonal code provides the same full diversity as the 1/2-rate orthogonal code does, while the constellation un-roatated quasi-orthogonal code indeed is a half-diversity space-time code and shows the same diversity gain of Alamouti code. Figure 3.4 presents the performance of several space-time coding scheme, which have fast ML decoding, at 2 bit/s/Hz data rate at 3/2 bit/s/Hz data rate. In constellation rotating schemes, the value of $\pi/6$ is used for θ optimize the code performance for QPSK signals as reported in [21] and [41].

Figure 3.5 shows the performance of full-rate and newly proposed 3/4 quasi-orthogonal codes, when 16 QAM modulation is applied and 1 receiving antenna is used.

3.6 Conclusions

Similar to the orthogonal space-time codes, the fast ML decoder of quasi-orthogonal space-time code can be constructed by a simple linear decoupling process followed by MRC and reduced dimension ML based symbol decision. In this chapter, it is also developed a systemic procedure to decouple the data of full-rate quasi-orthogonal

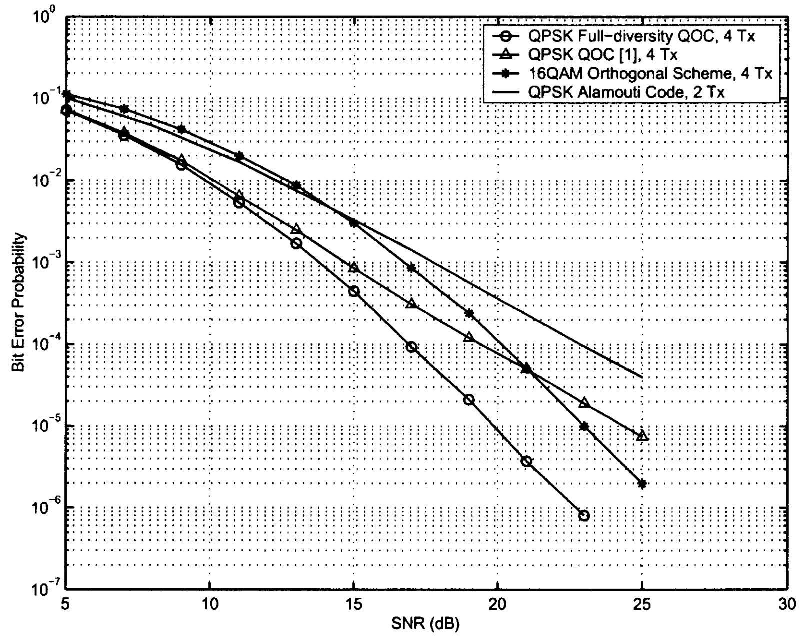

Figure 3.3 BER versus SNR for space-time block code at rate of 2 bit/s/ Hz. 4 transmit antennas and 1 receiving antenna is used. For reference (diversity), the Alamouti code with 2 transmit antennas is plotted.

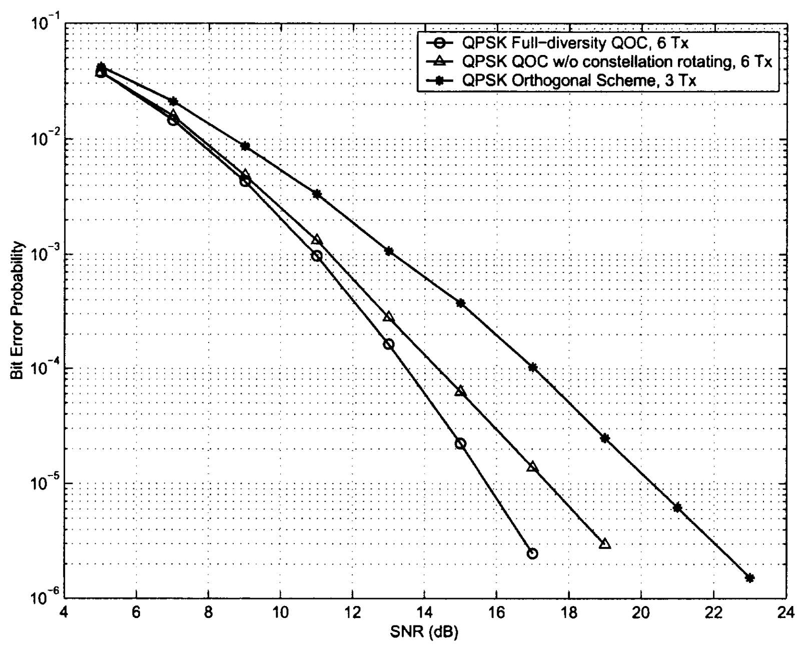

Figure 3.4 BER versus SNR for spsce-time block code at 3/2 bit/s/Hz; 1 receiving antennas is used.

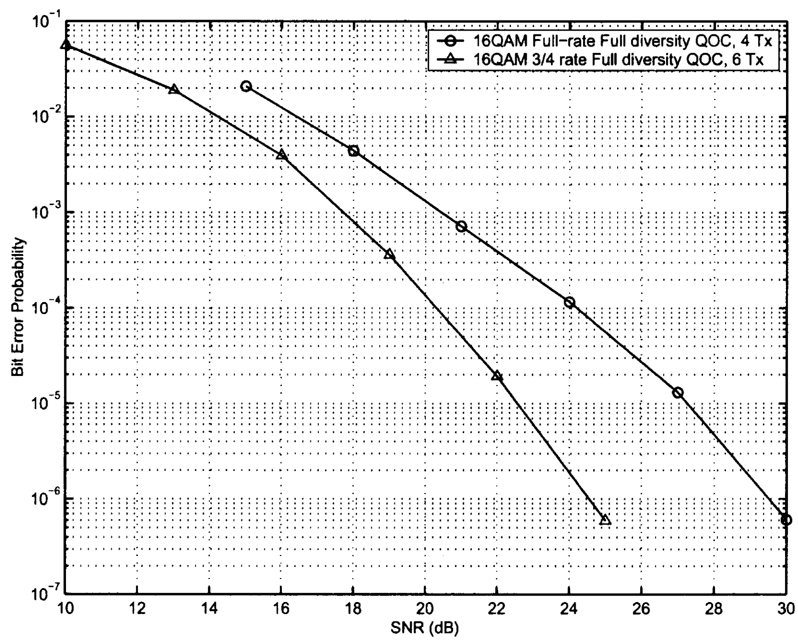

Figure 3.5 BER versus SNR for 16QAM modulated quasi-orthogonal space-time block code; 1 receiving antenna is used.

code and a newly proposed 3/4 rate quasi-orthogonal code to facilitate the fast ML detection. The similar decoupling process is performed at each receiving antenna in parallel. The output at decouplers are combined using MRC technique and fed into parallel sphere decoder. The newly proposed decoding algorithm has very low computational complexity and can use parallel computing techniques to satisfy the real-time requirement. The complexity of newly proposed decoding algorithm is independent of the constellation size of modulation, and suitable for high data rate transmission.

CHAPTER 4

Q-OSTBC CODED TRANSCEIVER SYSTEMS OVER FREQUENCY SELECTIVE WIRELESS FADING CHANNELS

In this chapter, it is proposed a time reveral QO-STBC (TR-QOSTBC) for system with 4 transmitt antennas in frequency selective channel. A MLSE decoding algorithm which provides full space and frequency diversity is also presented. In the MLSE decoding of TR-QOSTBC, a combination of time-domain filtering, and time reversal conjugate operation decouple a 4×1 MISO system over 4 independent frequency selective fading channels into 4 single-input single-output (SISO) frequency-selective sub-systems. Then these data streams are two by two jointly decoded using the vector Viterbi Algorithm (VA). However, since the computational complexity of vector Viterbi Algorithm increases exponentially with the maximum delay of the frequency selective channel, the MLSE detection algorithm is not practical for the channel with long delay time. It is also developed a fast decoding algorithm for TR-QOSTBC, which highly reduces the decoding complexity, especially for channels with large value of delay. The diversity properties and performance of the.fast decoding are compared with those of the MLSE detection.

4.1 Introduction

It should be pointed out that the works mentioned in above chapters all assumed that there is no inter-symbol interference in the received data. That is, a non-dispersive block fading narrowband channel model was assumed. This assumption is not valid for multipath wireless fading channels of time dispersive and frequency selective nature. In [42][43], a clever combination of time-domain filtering, and time reversal conjugate operation was used to convert a 2×1 multiple-input single-output (MISO) system composed of 2 frequency selective channels into 2 equivalent single-input single-output

(SISO) system with a combined frequency selective channel. After the conversion, a standard SISO equalization is applied to detect 2 independent information streams from transmitters[44][45][46].

In this chapter, such technique is successfully extended to block based QO-STBC transceiver system in combating channel dispersion and frequency selectivity. A new block based QO-STBC transceiver design scheme for wireless systems over frequency selective fading channels is proposed. In the proposed transceiver scheme, a combination of time-domain filtering, and time reversal conjugate operation is used to convert the 4×1 MISO system over 4 independent frequency selective fading channels into 2 pairs equivalent 2×1 MISO sub-systems with frequency-selective channels. Within each pair of MISO system, two independent data streams are jointly decoded using the Viterbi Algorithm (VA).

However, since the computational complexity of vector Viterbi Algorithm increases exponentially with the maximum delay of the frequency selective channel, the MLSE detection algorithm is not practical for the channel with long delay time. In this paper, it is developed a fast decoding algorithm for TR-QOSTBC, which highly reduces the decoding complexity, especially for channels with large value of delay. During the new fast decoding, the 4 decoupled data streams are MMSE equalized, then the equalized data without inter-symbol interference are two by two feed to sphere decoder and decoded as 2×2 frequency non-selective system. The noise power of equalized data is normalized before it is decoded by sphere decoder. The diversity properties and performance of the MMSE detection are compared with those of the MLSE detection.

4.2 Notation and System Model

The input/output (I/O) relation in a SISO system can be represented, in discrete time notation, as,

$$\begin{aligned} y(t) &= H(z)\,\{x(t)\} \\ &= h_0\,x(t) + h_1\,x(t-1) + \cdots + h_{L_a}x(t-L_a)\,, \end{aligned}$$

where $H(z) = \sum_{l=0}^{L_a} h_l\, z^{-l}$ is a FIR channel filter of length $(L_a + 1)$.

Similarly, for a MISO system, use a polynomial row vector to represent FIR channel filters; for a single-input-multiple-output (SIMO) system, use a polynomial column vector to represent FIR channel filters, and for a multiple-input-multiple-output (MIMO) system, use a polynomial matrix to represent the FIR filters between all possible transceiver pairs.

The conjugate reciprocal version of a SISO channel filter $H(z)$ is defined as,

$$H^*(1/z^*) \triangleq (h_0^* + h_1^* z + \cdots + h_{L_a}^* z^{L_a})\,,$$

which corresponds to a time reversal conjugate operation on the original impulse response. Similarly, the conjugate reciprocal versions of the MISO, SIMO and MIMO filters are defined according to the same time-reversal conjugate operation on their impulse responses of the corresponding matrix filters, respectively.

In this work, a MISO system with N transmitters and 1 receiver is considered. The received data sequence from such a system is,

$$r(t) = \sum_{k=1}^{N} H_k(z)\,\{\,s_k(t)\,\} + n(t) \tag{4.1}$$

where $H_k(z) = \sum_{l=0}^{L_k} h_{k,l} z^{-l}$ is the $(L_k + 1)$-ray multipath channel between the the k-th transmitter and the receiver; $s_k(t)$ is the symbol sequence transmitted from the k-th transmitter. The additive noise $n(t)$ is zero-mean complex Gaussian distributed, i.e. $n(t) \sim \mathcal{CN}(0, 1/\text{SNR})$.

It is further assumed that the maximum of L_k's equals L. The norm of each channel filter $H_k(z)$ is normalized, i.e. $\| \mathbf{h}_k \| = 1$ and $\mathbf{h}_k = [h_{k,0}\, h_{k,1} \cdots, h_{k,L_k}]^T$, $(k = 1, 2, \ldots, N)$. It is also assume that the total transmission power is equally distributed among all transmit antennas, and the received data in (4.1) is normalized so that the reception SNR is equivalently present in the noise power.

4.3 OSTBC for Frequency Selective Channels

4.3.1 Transmission Scheme

Consider a block of $2(P+1)$ digital modulated symbols to be transmitted by 2 transmit antennas and received by 1 receiving antenna over frequency selective fading channels. Using the *time-inversal* technique presented in [45][44][46][42], a block of symbols, $\{s(t)\}_{t=1}^{2(P+1)}$ is firstly decimated into 2 parallel sub-blocks, $\{s_1(t)\}_{t=0}^{P}$, $\{s_2(t)\}_{t=0}^{P}$. These 2 sub-blocks of symbols and/or their time reversal conjugate versions are to be transmitted through 2 transmitters according to the space-time scheduling outlined in the transmission code matrix in (4.6). To avoid the inter-block interference, preamble and midambles of length L are inserted into the OSTBC data streams prior to transmission. Detailed data transmission mechanism using the proposed block based OSTBC is shown in Figure 4.1.

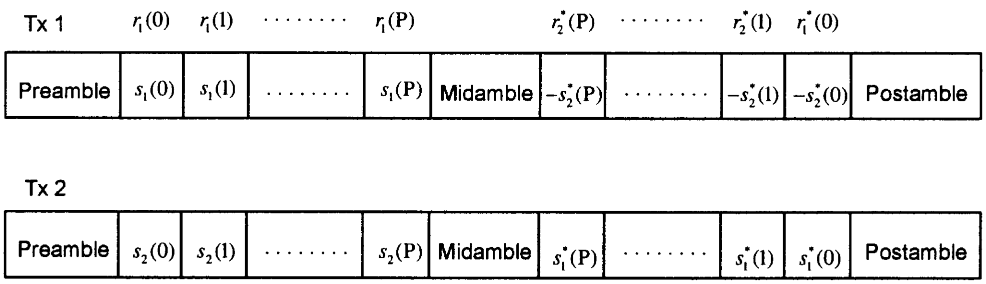

Figure 4.1 Scheme of signal transmitted from each transmit antenna.

Mathematically, such a space-time transmission scheme over frequency selective channels can be described by the following model after removing preamble/midambles and time reversal conjugating the corresponding sub-blocks of received data according to the block based OSTBC mechanism. That is, the properly arranged received data,

$$\begin{aligned} r_1(t) &= H_1(z)s_1(t) + H_2(z)s_2(t) + n_1(t) \\ r_2(t) &= H_2^*(1/z^*)s_1(t) - H_1^*(1/z^*)s_2(t) + n_2(t) \end{aligned}$$

The space-time transmission system for frequency selective channels can be described by

$$\underbrace{\begin{bmatrix} r_1(t) \\ r_2(t) \end{bmatrix}}_{\mathbf{r}(t)} = \mathbf{H}(z) \underbrace{\begin{bmatrix} s_1(t) \\ s_2(t) \end{bmatrix}}_{\mathbf{s}(t)} + \underbrace{\begin{bmatrix} n_1(t) \\ n_2(t) \end{bmatrix}}_{\mathbf{n}(t)} \tag{4.2}$$

where $\mathbf{H}(z)$ is a polynomial matrix of the form,

$$\mathbf{H}(z) = \begin{bmatrix} H_1(z) & H_2(z) \\ H_2^*(1/z^*) & -H_1^*(1/z^*) \end{bmatrix}$$

4.3.2 Optimal Detection of Time-inversal OSTBC

The polynomial channel matrix $\mathbf{H}(z)$ in (4.2) is orthogonal in the sense that

$$\begin{aligned} \mathbf{H}^H(z)\mathbf{H}(z) &= H_1^*(1/z^*)H_1(z) + H_2^*(1/z^*)H_2(z) \\ &= G(z) \end{aligned}$$

At the receiver the received data can be decoupled by the filter bank $\mathbf{H}^H(z)$. The decoupled results is given by

$$\begin{aligned} \tilde{\mathbf{r}}(t) &= \begin{bmatrix} \tilde{r}_1(t) \\ \tilde{r}_2(t) \end{bmatrix} = \mathbf{H}^H(z)\mathbf{r}(t) \\ &= G(z)\mathbf{s}(t) + \mathbf{v}(t) \end{aligned}$$

where

$$\mathbf{v}(t) = \begin{bmatrix} v_1(t) \\ v_2(t) \end{bmatrix} = \mathbf{H}^H(z)\mathbf{n}(t)$$

It can be found that the noise sequence $v_1(t)$ and $v_2(t)$ are uncorrelated as the spectrum of $\mathbf{v}(t)$ given by

$$\begin{aligned} \mathbf{R}_{\mathbf{vv}}(z) &= \sum_{m=-\infty}^{+\infty} E\left[\mathbf{v}(t)\mathbf{v}^H(t-m)\right] \\ &= \mathbf{H}^H(z)\mathbf{R}_{\mathbf{nn}}(z)\mathbf{H}(z) \\ &= \sigma_n^2\mathbf{H}^H(z)\mathbf{H}(z) \\ &= \sigma_n^2 G(z)\mathbf{I} \end{aligned}$$

has no cross terms between $v_1(t)$ and $v_2(t)$. In the third equality the fact is noticed that $\mathbf{n}(t)$ is a white vector noise sequence with $\mathbf{R}_{\mathbf{nn}}(z) = \sigma_n^2\mathbf{I}$. So symbol blocks $\{s_1(t)\}$ and $\{s_2(t)\}$, are decoupled, hence, can be ML decoded independently. The noise is spatially (inter-block-wise) *uncorrelated* and temporally (intra-block-wise) colored. Therefore, each sub-blocks of data $\tilde{\mathbf{r}}(t)$ can be whitened individually for subsequential processing.

It can be prove that $G(z)$ always can be written in the form of $G(z) = \tilde{G}^*(1/z^*)\tilde{G}(z)$, where $\tilde{G}(z)$ is a causal filter.

$$G(z) = H_1^*(1/z^*)H_1(z) + H_2^*(1/z^*)H_2(z) = \sum_{k=-L}^{+L} g_k z^{-k}$$

and it can be found that $g_k = g_{-k}^*$, it follows that $G(z) = G^*(z)$ and the $2L$ roots of $G(z)$ have the symmetry that if ρ is a root, $1/\rho^*$ is also a root. Hence, $G(z)$ can be factored and expressed as

$$G(z) = \tilde{G}^*(1/z^*)\tilde{G}(z) \tag{4.3}$$

where $\widetilde{G}(z)$ is a polynomial of degree L having the roots $\rho_1, \rho_2, \cdots, \rho_L$ and $\widetilde{G}^*(1/z^*)$ is a polynomial of degree L having the roots $1/\rho_1^*, 1/\rho_2^*, \cdots, 1/\rho_L^*$. 4.3 is called spectral decomposition.

From (4.3), the whitening filter designed for $\mathbf{v}(t)$ can be $\left(\widetilde{G}(z)\right)^{-1}$. At the back end of whitening filter it is obtained,

$$\underline{\mathbf{r}}(t) = \frac{1}{\widetilde{G}(z)}\widetilde{\mathbf{r}}(t) = \widetilde{G}^*(1/z^*)\mathbf{s}(t) + \widetilde{\mathbf{v}}(t)$$

where $\widetilde{\mathbf{v}}(t)$ is white Gaussian noise vector and the variance of each element is σ_n^2. The maximum likelihood sequence estimation (MLSE) can be performed by Viterbi algorithm.

4.3.3 Diversity Analysis

Assuming the knowledge of channel state information (CSI), the probability of transmi-tting sequence of $s(t)$ and deciding in favor of sequence of $e(t)$ at the decoder can be bounded by the following [32],

$$P\left(s(t) \rightarrow e(t)|H_1(z^{-1}), H_2(z^{-1})\right) \leq \exp(-d^2(s(t), e(t))/2\sigma_n^2)$$

where σ_n^2 is the variance of the noise and

$$d^2(\mathbf{c}, \mathbf{e}) = norm(\widetilde{G}^*(1/z^*)\left\{\Delta_1(t)\right\})^2 \tag{4.4}$$

Define $\Delta_1(t) = s(t) - e(t)$ as error sequences.

Assume that SNR is high enough so that error propagation can be ignored. The symbol errors occur independently, hence, the pairwise error probability (PEP) of the t_0-th symbol is bounded by

$$P\left(s(t_0) \rightarrow e(t_0)|\mathbf{H}(z)\right) \leq \exp(-d^2(c(t_0), e(t_0))/2\sigma_n^2)$$

where

$$
\begin{aligned}
& d^2(s(t_0), e(t_0)) \\
= \; & norm(\widetilde{G}^*(1/z^*)\{\Delta_1(t_0)\})^2 \\
= \; & \left(\sum_{n=1}^{2}\sum_{l=0}^{L} |h_{n,l}|^2\right)_{l=0}^{L} |s(t_0) - e(t_0)|^2
\end{aligned} \tag{4.5}
$$

If $h_{n,l}$'s are complex Gaussian distributed, $2(L+1)$ diversity gain is obtained.

4.4 QO-STBC for Frequency Selective Fading Channels

4.4.1 Quasi-orthogonal Space-Time Block Code

The QO-STBC code for a system with 4 transmitters was proposed by Jafarkhani in [20]. Given a block of 4 symbols to be transmitted by 4 transmitters over 4 consecutive time slots over a block flat fading channel, the QO-STBC is defined by the following transmission code matrix,

$$
\mathbf{C}(\mathbf{s}) = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ \underbrace{s_4}_{\mathbf{c}_1} & \underbrace{-s_3}_{\mathbf{c}_2} & \underbrace{-s_2}_{\mathbf{c}_3} & \underbrace{s_1}_{\mathbf{c}_4} \end{bmatrix} = \begin{bmatrix} \mathbf{C}_1 & \mathbf{C}_2 \\ -\mathbf{C}_2^* & \mathbf{C}_1^* \end{bmatrix}. \tag{4.6}
$$

Each row of the above transmission matrix denotes symbols simultaneously sent through four transmit antennas within a given transmission slot. Note that only the subspace $\langle \mathbf{c}_1, \mathbf{c}_4 \rangle$ is orthogonal to the subspace $\langle \mathbf{c}_2, \mathbf{c}_3 \rangle$, where $\mathbf{c}_i$ denotes the i-th column of the code matrix $\mathbf{C}(\mathbf{s})$. Hence, the STBC in (4.6) is called quasi-orthogonal (QO) code.

4.4.2 Block Based QO-STBC Scheme

Now consider a block of $4(P+1)$ digitally modulated symbols to be transmitted by 4 transmitters and subsequently received by 1 receiving antenna over frequency selective fading channels. For the block based transmission over a multipath fading channel, a *time reversal conjugate technique* is used to embedded the quasi-orthogonality into the STBC blocks. Essentially, first decimate a block of symbols, $\{s(t)\}_{t=1}^{4(P+1)}$, into 4 parallel sub-blocks, $\{s_1(t)\}_{t=0}^{P}$, $\{s_2(t)\}_{t=0}^{P}$, $\{s_3(t)\}_{t=0}^{P}$ and $\{s_4(t)\}_{t=0}^{P}$. These 4 sub-blocks of symbols and/or their time reversal conjugate versions are to be transmitted through 4 transmitters according to the space-time scheduling outlined in the transmission code matrix in (4.6). To avoid the inter-block interference, preamble and midambles of length L are inserted into the QO-STBC data streams prior to transmission. Detailed data transmission mechanism using the proposed block based QO-STBC is shown in Figure 4.2.

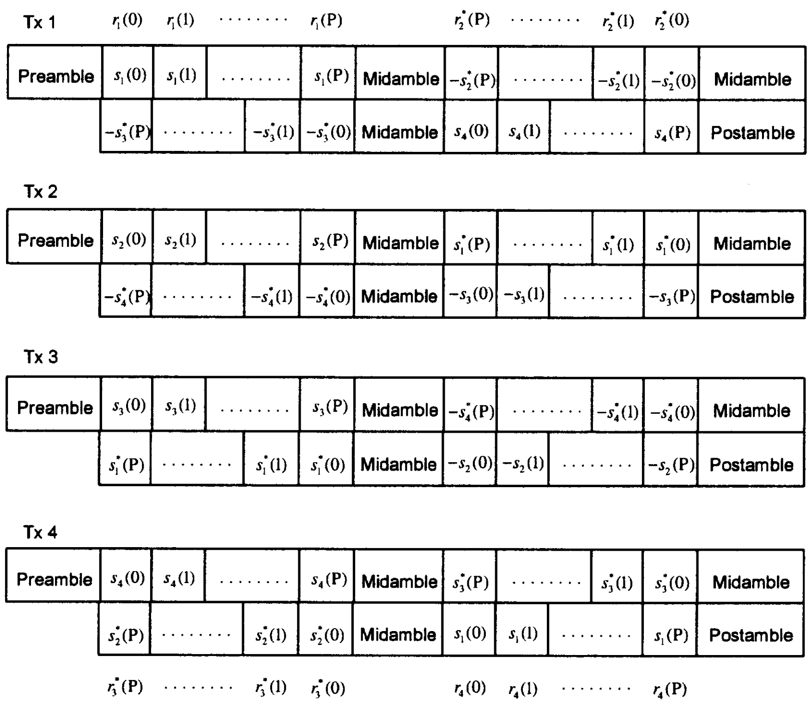

Figure 4.2 Transmission mechanism of the proposed block based QO-STBC.

Mathematically, such a space-time transmission scheme over frequency selective channels can be described by the following model after removing preamble/midambles and time reversal conjugating the corresponding sub-blocks of received data according to the block based QO-STBC mechanism. That is, the properly arranged received data,

$$\mathbf{r}(t) = \mathbf{H}(z)\,\{\mathbf{s}(t)\} + \mathbf{n}(t)\,, \tag{4.7}$$

where $\mathbf{H}(z)$ is a polynomial matrix of the form,

$$\begin{bmatrix} H_1(z) & H_2(z) & H_3(z) & H_4(z) \\ H_2^*(1/z^*) & -H_1^*(1/z^*) & H_4^*(1/z^*) & -H_3^*(1/z^*) \\ H_3^*(1/z^*) & H_4^*(1/z^*) & -H_1^*(1/z^*) & -H_2^*(1/z^*) \\ H_4(z) & -H_3(z) & -H_2(z) & H_1(z) \end{bmatrix}$$

and $\mathbf{s}(t) = [s_1(t)\ s_2(t)\ s_3(t)\ s_4(t)]^T$.

4.4.3 ML Detection of Block Based QO-STBC Symbols

Using the fact $\langle \mathbf{c}_1,\ \mathbf{c}_4 \rangle \perp \langle \mathbf{c}_2,\ \mathbf{c}_3 \rangle$, Equation (4.7) can be transformed into

$$\mathbf{r}(t) = \widetilde{\mathbf{H}}(z)\ \widetilde{\mathbf{s}}(t) + \mathbf{n}(t)$$

where

$$\widetilde{\mathbf{s}}(t) = \frac{1}{2}\begin{bmatrix} s_1(t)+s_4(t) & s_2(t)+s_3(t) & s_2(t)-s_3(t) & s_1(t)-s_4(t) \end{bmatrix}^T$$

and

$$\widetilde{\mathbf{H}}(z) = \begin{bmatrix} \mathbf{h}_1(z)+\mathbf{h}_4(z) & \mathbf{h}_2(z)+\mathbf{h}_3(z) & \mathbf{h}_2(z)-\mathbf{h}_3(z) & \mathbf{h}_1(z)-\mathbf{h}_4(z) \end{bmatrix}^T$$

and $\mathbf{h}_i(z)$ is the i-th column of the matrix $\mathbf{H}(z)$.

Note the important fact that the polynomial channel matrix $\widetilde{\mathbf{H}}(z)$ here becomes *orthogonal* in the sense that $\mathbf{G}(z) = \widetilde{\mathbf{H}}^H(z)\widetilde{\mathbf{H}}(z)$ is a diagonal matrix. That is,

$$\mathbf{G}(z) = \begin{bmatrix} G_1(z) & 0 & 0 & 0 \\ 0 & G_2(z) & 0 & 0 \\ 0 & 0 & G_1(z) & 0 \\ 0 & 0 & 0 & G_2(z) \end{bmatrix},$$

where

$$\begin{aligned} G_1(z) = & \; 2\left(H_1^*(1/z^*) + H_4^*(1/z^*)\right)\left(H_1(z) + H_4(z)\right) \\ & + 2\left(H_2^*(1/z^*) - H_3^*(1/z^*)\right)\left(H_2(z) - H_3(z)\right) \end{aligned}$$

and

$$\begin{aligned} G_2(z) = & \; 2\left(H_1^*(1/z^*) - H_4^*(1/z^*)\right)\left(H_1(z) - H_4(z)\right) \\ & + 2\left(H_2^*(1/z^*) + H_3^*(1/z^*)\right)\left(H_2(z) + H_3(z)\right) \end{aligned}$$

Therefore, the interesting decoupled results are obtained after a matrix channel matched filtering. That is,

$$\widetilde{\mathbf{r}}(t) = \widetilde{\mathbf{H}}^H(z)\{\mathbf{r}(t)\} = \mathbf{G}(z)\{\widetilde{\mathbf{s}}(t)\} + \mathbf{v}(t) \tag{4.8}$$

where the noise $\mathbf{v}(t) = \widetilde{\mathbf{H}}^H(z)\{\mathbf{n}(t)\}$, is found to be spatially (inter-block-wise) uncorrelated, since the spectrum of $\mathbf{v}(t)$ is given by

$$\begin{aligned} \mathbf{R}_{\mathbf{vv}}(z) &= \sum_m E\left[\mathbf{v}(t)\,\mathbf{v}^H(t-m)\right] z^{-m} \\ &= \mathbf{H}^H(z)\,\mathbf{R}_{\mathbf{nn}}(z)\,\mathbf{H}(z) \\ &= \sigma_n^2 \mathbf{G}(z)\,. \end{aligned}$$

From (4.8), it can be found that two pairs of symbol blocks $(\{s_1(t)\},\ \{s_4(t)\})$ and $(\{s_2(t)\},\ \{s_3(t)\})$ are decoupled, hence, can be ML decoded independently. The noise is spatially (inter-block-wise) *uncorrelated* and temporally (intra-block-wise) colored. Therefore, each sub-blocks of data $\tilde{\mathbf{r}}(t)$ can be whitened individually for subsequential processing.

The well known spectral decomposition fact can be observed in the polynomials $G_i(z) = \tilde{G}_i(z)\tilde{G}_i^*(1/z^*)$, $(i = 1, 2)$. The whitening filter for the i-th information block can be designed as $\left(\tilde{G}_i(z)\right)^{-1}$. And the whitening filter bank for decoupled information streams is simply the following diagonal polynomial matrix,

$$\left(\tilde{\mathbf{G}}(z)\right)^{-1} = \begin{bmatrix} \tilde{G}_1^{-1}(z) & 0 & 0 & 0 \\ 0 & \tilde{G}_2^{-1}(z) & 0 & 0 \\ 0 & 0 & \tilde{G}_1^{-1}(z) & 0 \\ 0 & 0 & 0 & \tilde{G}_2^{-1}(z) \end{bmatrix}.$$

At the output of whitening filter it is obtained,

$$\begin{aligned} \underline{\mathbf{r}}(t) &= \left(\tilde{\mathbf{G}}(z)\right)^{-1}\{\tilde{\mathbf{r}}(t)\} \\ &= \tilde{\mathbf{G}}^*(1/z^*)\{\tilde{\mathbf{s}}(t)\} + \tilde{\mathbf{v}}(t), \end{aligned}$$

where $\tilde{\mathbf{v}}(t)$ is WGN vector and the variance of each element is σ_n^2.

The ML sequence estimation (MLSE) can then be performed on the above data set using the Viterbi algorithm.

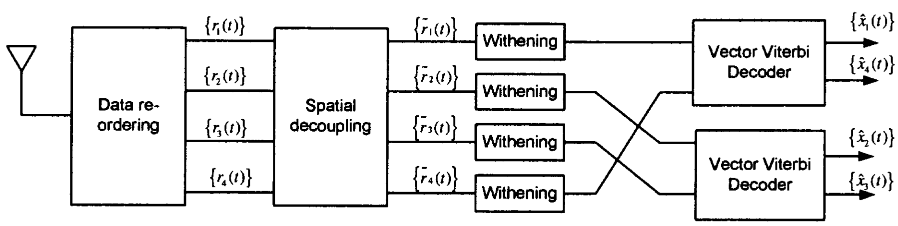

Figure 4.3 The structure of MLSE decoder for TR-QOSTBC.

Diversity Discussion Given the knowledge of channel state information (CSI), the probability of transmitting sequence of $\mathbf{c} = (s_1(t),\ s_4(t))$ and deciding in favor of sequence of $\mathbf{e} = \left(s_1^{'}(t),\ s_4^{'}(t)\right)$ at the decoder can be bounded by the following [32],

$$P\left(\mathbf{c} \to \mathbf{e}|\mathbf{H}(z)\right) \leq \exp(-d^2(\mathbf{c}, \mathbf{e})/2\sigma_n^2)\,,$$

where σ_n^2 is the variance of the noise; and

$$d^2(\mathbf{c}, \mathbf{e}) = norm(\widetilde{G}_1^*(1/z^*)\left\{\Delta_1(t)\right\})^2 + norm(\widetilde{G}_4^*(1/z^*)\left\{\Delta_4(t)\right\})^2 \tag{4.9}$$

Define $\Delta_1(t) = (s_1(t) + s_4(t)) - (s_1'(t) + s_4'(t))$ and $\Delta_4(t) = (s_1(t) - s_4(t)) - (s_1'(t) - s_4'(t))$ as error sequences.

Assume that SNR is high enough so that error propagation can be ignored. The symbol errors occur independently, hence, the pairwise error probability (PEP) of the t_0-th symbol is bounded by

$$P\left(\mathbf{c}(t_0) \to \mathbf{e}(t_0)|\mathbf{H}(z)\right) \leq \exp(-d^2(\mathbf{c}(t_0), \mathbf{e}(t_0))/2\sigma_n^2)$$

where

$$\begin{aligned} & d^2(\mathbf{c}(t_0), \mathbf{e}(t_0)) \qquad\qquad (4.10)\\ = \ & norm(\widetilde{G}_1^*(1/z^*)\left\{\Delta_1(t_0)\right\})^2 + norm(\widetilde{G}_4^*(1/z^*)\left\{\Delta_4(t_0)\right\})^2 \\ = \ & \sum_{l=0}^{L}\left[\left(|h_{1,l} + h_{4,l}|^2 + |h_{2,l} - h_{3,l}|^2\right)|\Delta_1(t_0)|^2 + \left(|h_{1,l} - h_{4,l}|^2 + |h_{2,l} + h_{3,l}|^2\right)|\Delta_4(t_0)|^2\right] \end{aligned}$$

If $h_{k,l}$'s are complex Gaussian distributed and both $\Delta_1(t_0)$ and $\Delta_4(t_0)$ are non-zero, an order of $4(L+1)$ diversity gain is obtained.

4.5 Fast decoding scheme for TR-Q-OSTBC

Since the computational complexity of vector Viterbi Algorithm increases exponentially with the maximum delay of the frequency selective channel, the MLSE detection algorithm is not practical for the channel with long delay time. Observing Eq.(4.8), it is found that the system has been decoupled into four SISO frequency selective system which can be easily equalized. The decoupled data stream of $\widetilde{\mathbf{r}}(t)$ is equalized by means of Wiener filter $\mathbf{w}_o$ or MMSE equalizer.

$$\begin{aligned} r_{e,i}(t) &= \mathbf{w}_{o,i} \star \underline{r}_i(t) \\ &= \widetilde{s}_i(t) + \widetilde{v}_{e,i}(t)\,, \end{aligned} \tag{4.11}$$

where $\star$ respresents the convolution operation. The MMSE equalized result $\mathbf{r}_e(t)$ has no inter-symbol interference (ISI) and the residue noise $\widetilde{\mathbf{v}}_e(t)$ is with Gaussian noise whose variance can be easily calculated [47].

From (4.11), the jointly ML detection of $\big(s_{1/4}(t)\ s_{2/3}(t)\big)$ can be implemented by exhaustive searching. When higher order modulation is involved, the complexity of symbol pairs based exhaustive searching in (4.11) increases exponentially. To further reduce decoding complexity, it is propose a sphere decoding scheme, which is well suitable for decoding QAM modulated quasi-orthogonal coding system. The complexity of the decoding algorithm is independent of the modulation order, hence it is very useful for high data rate transmission. Sphere decoder has been applied in decoding of V-BLAST space-time system [29]. Study on sphere decoding algorithm of Fincke and Pohst shows that for a wide range of noise variances the expected complexity is polynomial, in fact often roughly cubic [29].

In the proposed decoding scheme, the data at receiving antenna is processed by the proposed linear decoupling scheme in (4.8). The output of all decoupler is equalized by Wiener filter. The results obtained after equalization can be written in a pair-wise format as follows:

$$\underbrace{\begin{bmatrix} \frac{1}{\sigma_{\widetilde{v}_{e,i}}} r_{e,i}(t) \\ \frac{1}{\sigma_{\widetilde{v}_{e,j}}} r_{e,j}(t) \end{bmatrix}}_{\mathbf{r}(i,j)} = \underbrace{\begin{bmatrix} \frac{1}{2\sigma_{\widetilde{v}_{e,i}}} & e^{j\theta}\frac{1}{2\sigma_{\widetilde{v}_{e,i}}} \\ \frac{1}{2\sigma_{\widetilde{v}_{e,j}}} & -e^{j\theta}\frac{1}{2\sigma_{\widetilde{v}_{e,j}}} \end{bmatrix}}_{\mathbf{M}(i,j)} \underbrace{\begin{bmatrix} s_i \\ s_j \end{bmatrix}}_{\mathbf{s}(i,j)} + \eta \tag{4.12}$$

where $\mathbf{M}$ is the matrix defined the lattice and η is the noise vector, whose elements are iid complex white Gaussian random variables. When the constellation at transmitter 3 and 4 are not rotated, $\theta = 0$. The operation in (4.12) is similar to what used in a 2×2 V-BLAST system, which can be decoded by a sphere decoder proposed in [40][29].

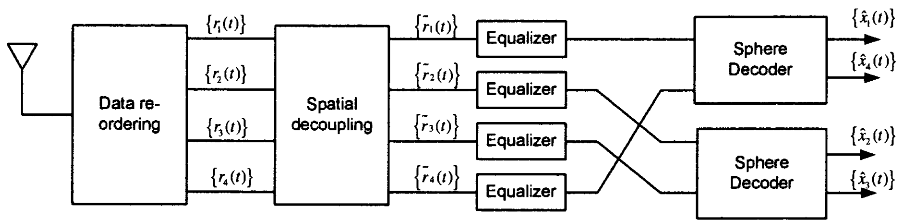

Figure 4.4 Structure of the fast decoding of TR-QOSTBC.

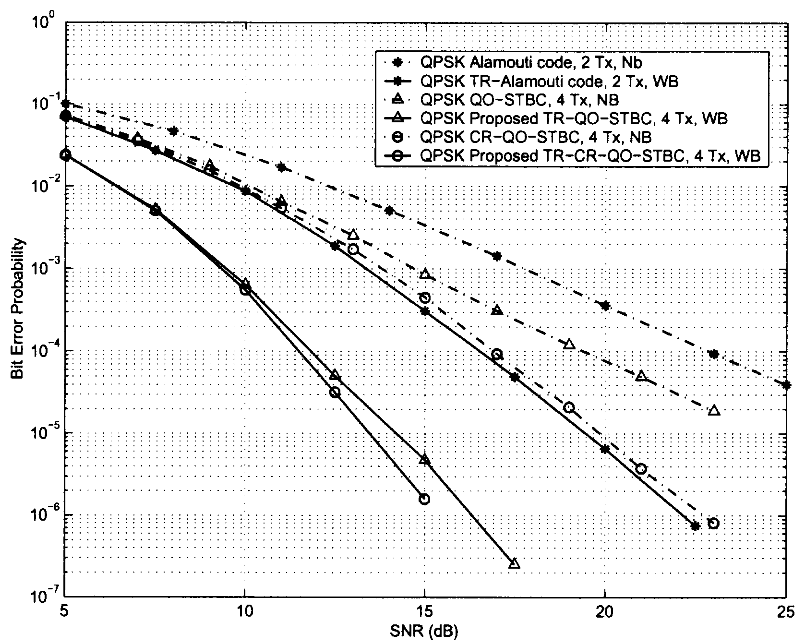

Figure 4.5 BER versus SNR performance of time-inversal space-time block coding schemes in frequency selective channels compared with performance of space-time block coding schemes in non-frequency selective channels.

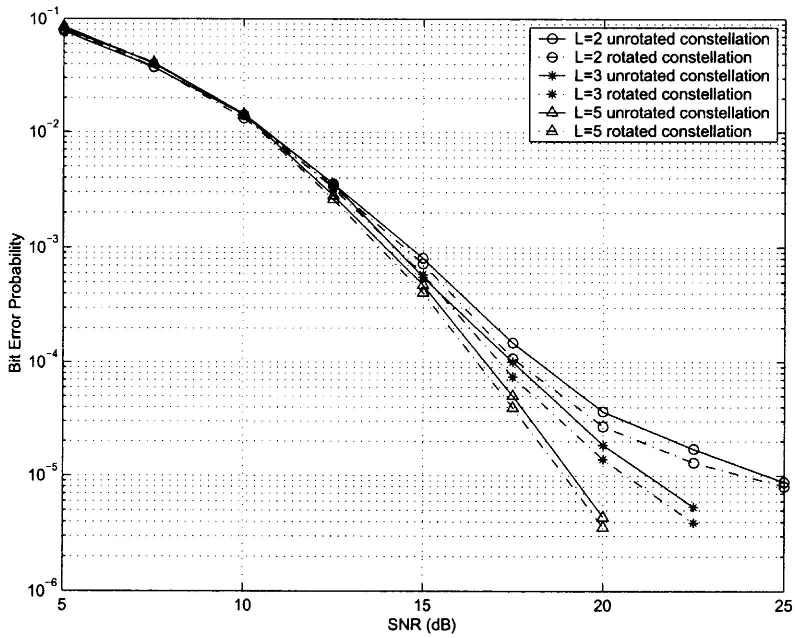

Figure 4.6 BER of fast decoding versus SNR performance of time-inversal space-time block coding schemes in frequency selective channels.

4.6 Simulation Results

In the computer simulated experiments, a block containing $4(P+1) = 120$ QPSK modulated data symbols is transmitted. The system under consideration has multiple transmitters and one receiver. A QO-STBC system with $N = 4$ transmitters is considered in a frequency selective fading channel. As a performance reference, the Alamouti O-STBC system with $N = 2$ transmitters in a frequency *non-selective* fading channel is also considered. To simulate the BER of MLSE decoding, the block faded frequency selective channels are modeled as two tap FIR filters with a delay of one symbol duration time; and the value of each tap are i.i.d. complex Gaussian distributed. In the simulation of fast decoding algorithm, the MMSE equalizers use $4(L-1)+1$ taps and the SNR is known at receiver side.

Figure 4.5 shows the performance of MLSE or ML decoded STBC transceiver systems using the constellation rotated and non-rotated QO-STBC (4×1 systems) in frequency selective environment, as well as the O-STBC (the Alamouti code, 2×1

system) in frequency non-selective environment. The same curve slopes at high SNRs indicate the same order of diversity for the corresponding encoding schemes. As expected, the constellation rotated QO-STBC shows better diversity property than the un-rotated quasi-orthogonal code. Figure 4 proves that block based time reversal conjugate space-time block coding technique can not only be used for *orthogonal* STBC, such as Alamouti code [42][43], but also be used for *quasi-orthogonal* scheme to obtain diversity gain from frequency selective channels. Compared with the narrowband scheme, the time reversal conjugate constellation rotated and non-rotated QO-STBC achieve SNR gains about 8 dB and 5 dB, respectively.

The performance of the newly proposed fast decoding algorithm is presented by Figure 4.6 It shows that the new algorithm does not provided full diversity gain and error floor is observed. As the delay time of frequency selective channel increase, the new decoding algorithm also can provide higher degree of diversity gain.

4.7 Conclusions

Applying the time reversal conjugate technique in combination with the Q-OSTBC scheduling, it is proposed a block based Q-OSTBC scheme to effective combat channel dispersion and frequency selective fading, and obtained diversity gain. The ML detector of the time-reversal Q-OSTBC can be constructed by a simple linear filter bank for space domain decoupling, a whitening filter bank and a vector MLSE detector. The coding rate of the time reversal Q-OSTBC is $L_B/4L$, where L_B is length of the data block. The encoding complexity is very low. The decoding complexity is determined by the computational complexity of vector MLSE detector, which increases exponentially with the channel delay time. When channel delay time is short, the ML detection can be accomplished at low complexity. As channel delay time is long, the decoding complexity can be reduced by using some sub-optimum detector instead of the optimum MLSE detector.

CHAPTER 5

REDUCED COMPLEXITY MAXIMUM LIKELIHOOD DETECTION FOR V-BLAST SYSTEMS

Proposed and studied in this chapter is a reduced complexity maximum likelihood (ML) detection scheme, which uses a pre-decoder guided local exhaustive search, for V-BLAST systems. A polygon searching algorithm and an ordered successive interference cancellation (O-SIC) sphere searching algorithm are major components of the proposed multi-step ML detectors. The effects of pre-decoder's performance on the total decoding complexity was studied by simulation. The complexity of the proposed algorithm depends on the operating signal to noise ratio (SNR) of the system. At reasonable high SNRs, its complexity is comparable to that of O-SIC algorithm, a commonly used decoding algorithm for V-BLAST. For V-BLAST systems using a large number of transmit antennas, it is also proposed a new detection algorithm termed ordered group-wise interference cancellation (O-GIC) to further reduce decoding complexity. The O-GIC based detection scheme is not a ML detection scheme, however, its performance is shown to outperform that of the O-SIC.

5.1 Detection of V-BLAST System

Recent advances in information theory reveal an important fact that the rich-scattering multi-path wireless channel can provide enormous capacity if the multi-path propagation is properly exploited using transceiver systems with multiple antennas [16][25]. The communications system architecture of the V-BLAST provides an experimental demon-stration of such a system. Such a system typically involves an exhaustive searching over all possible signal vector to obtain the ML decision on transmitted discrete symbols, given the channel state information (CSI). However, the computational complexity of the global exhaustive search increases exponentially with the number of transmit antennas and the size of signal modulation constellation. Due to its computational complexity, global exhaustive search algorithm has limited potential in many real-time applications. In order to reduce computational complexity, the ordered successive interference cancellation (O-SIC) algorithm was proposed as a detection scheme [28][27]. However, there exists a wide gap between the performance obtained by O-SIC and that of the ML detection.

To reduce the complexity of ML detection for MIMO system, a local search based multi-step detection scheme has been presented in [48]. In the first step, an initial data estimate is provided by a suboptimal detector. In the second step, a localized ML search is performed over the combination of reduced constellation for each transmit antenna, composed by the neighboring original constellation points surrounding the initial estimate. Hence, a total number of 4^N (N is the number of transmit antennas) possible signal points will be exammed to find the ML estimate using detector proposed in [48]. It has been observed that it is possible to search in a smaller region when a better initial estimate is obtained at higher system SNR. The sphere decoding algorithm [49][50], which searches through the points of the lattice found inside a sphere of given radius d centered at the received points, is also a local searching algorithm. Recently, sphere decoder was applied to decode the V-BLAST system [51] and obtain obvious performance improvement over O-SIC decoder. For sphere decoding algorithm, the radius d should be chosen carefully. If the d is chosen too tight,a lattice inside the sphere can not be found; if the d is chosen too loose, unnecessary computation will increase.

In this chapter, the multi-step ML detector consisting of a pre-decoder followed by a local exhaustive search decoder is proposed and studied. A new polygon searching algorithm for the back end decoder is firstly proposed. The complexity of the polygon searching detector depends on the performance of the pre-decoder, as well as the system SNR. A higher SNR, and/or a better performance of the pre-decoder will result in lower complexity on the subsequent local search detector. At reasonable high SNR, the complexity of proposed detector can be very close to that of pre-decoder. When sphere-decoder is used as the back end decoder, it is also proposed a modified O-SIC algorithm for the pre-decoder to improve the computational efficiency. The distance between initial estimate of pre-decoder and received data point is a intuitively reasonable choice of d for back end sphere decoder. When V-BLAST system use a large number of transmit antennas, the complexity of multi-step ML detection may still be high. Hence, it is further developed a novel ordered group interference cancellation (O-GIC) detection algorithm for such applications. Detection using O-GIC is not ML scheme, however, the O-GIC provides much better performance than the O-SIC.

5.2 System Description

A general V-BLAST architecture with N transmitting and M receiving antennas ($M \geq N$) is considered. A data stream is divided into N sub-streams and sent to N transmit antennas. Transmitters operate co-channel at symbol rate $1/T$ symbols/sec., with synchronized symbol timing. At the receiver side, each receiving antenna picks up a combination of signals coming from all N transmitters [52]. the quasi-static block fading wireless channel model is adopted, so that the path gains are constant over a frame, and vary from frame to frame.

At time t, the signal y_m received at antenna m can be written as

$$y_m = \sum_{n=1}^{N} h_{n,m}\, x_n + v_m, \;\; m = 1, 2, \dots, M \tag{5.1}$$

Considering all the data received at M receiving antennas, the received data can be modeled as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{v}, \tag{5.2}$$

where $\mathbf{y}$, $\mathbf{x}$ and $\mathbf{v}$ are $M \times 1$ vectors obtained by stacking y_m, x_n and v_m, where $m = 1, 2, \ldots, M$, $n = 1, 2, \ldots, N$. $\mathbf{H} = [\mathbf{h}_1, \mathbf{h}_2, \cdots \mathbf{h}_N]$ is the $M \times N$ channel matrix with $\mathbf{h}_n = \begin{bmatrix} h_{n,1} & h_{n,2} & \ldots & h_{n,M} \end{bmatrix}^{\mathsf{T}}$, where T denotes the transpose operator. The entries of $\mathbf{H}$ are mutually independent, zero-mean, and circularly symmetric complex Gaussian variables of unit-variance. The noise samples v_m are spatially and temporally independent samples from a zero mean complex Gaussian family, i.e. $v_m \sim \mathcal{CN}(0, 1/\text{SNR})$. Note the fact that equal average power is assigned among all transmit antennas, and the total power is normalized to 1, the signal to noise ratio SNR at reception is present in the *effective* noise variance at each receiving antenna.

Consistently with the V-BLAST concept, the signals transmitted from each transmit antenna are independent. Hence, the covariance matrix of $\mathbf{x}$ can be expressed as follows,

$$E[\mathbf{x}\,\mathbf{x}^H] = \frac{1}{N}\mathbf{I}_{N\times N}\,. \tag{5.3}$$

5.3 Multi-step ML Detectors

Given the channel matrix $\mathbf{H}$, the ML solution for detecting $\mathbf{x}$ from $\mathbf{y}$ is

$$\widehat{\mathbf{x}}_{\text{ML}} = \arg\min_{\mathbf{x}\in\mathcal{C}} \|\mathbf{y} - \mathbf{H}\mathbf{x}\|^2 \tag{5.4}$$

where $\mathcal{C}$ is the discrete symbol constellation set. The global searching ML detection scheme considers all possible input vector $\mathbf{x}$ within $\mathcal{C}$, and choose the candidate that minimizes the Euclidean distance between the received data vector $\mathbf{y}$ and the hypothetic signal $\mathbf{Hx}$. It is easy to see that the computation complexity increases exponentially with the number of transmit antennas, N, and the constellation size of modulation at each transmit antenna.

In order to reduce the computational complexity, yet still achieving the ML detection performance, it is propose to use a multi-step ML detector instead of global exhaustive searching detector. The structure of the multi-step ML detector is shown in Figure 5.1. In the first step, an initial estimate of data, $\hat{\mathbf{x}}_p$, is generated by a suboptimal detector, such as ZF or O-SIC detector. Then the initial estimate is used by the back end decoder to shrink the searching region from a global area to a localized small area. The key point is that if all signal points whose distance to received data point is smaller than d, the distance between the initial estimate data points and received data points, are all searched, the ML detection of the system can always be found.

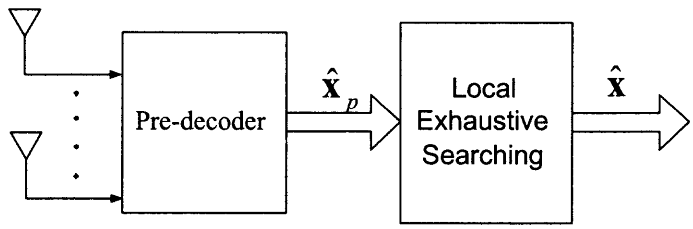

Figure 5.1 Structure of the multi-step ML detector.

5.3.1 Polygon local Searching Detector

Since the ML solution $\hat{\mathbf{x}}_{\text{ML}}$ is located within a sphere of radius d centered at the received data point $\mathbf{y}$. The sphere region can be described as

$$\|\mathbf{y} - \mathbf{H}\,\hat{\mathbf{x}}_{\text{ML}}\|^2 \leq \|\mathbf{y} - \mathbf{H}\,\hat{\mathbf{x}}_p\|^2 = d^2. \tag{5.5}$$

where $\hat{\mathbf{x}}_p$ is the pre-estimation of $\mathbf{x}$.

By defining a region that includes this sphere specified by (5.5), the ML solution always can be found by means of the exhaustive search over all the signal points within the defined region. The worst case is that it is needed to search all possible signal points, the global searching. The best case is that the searching area is just the sphere area.

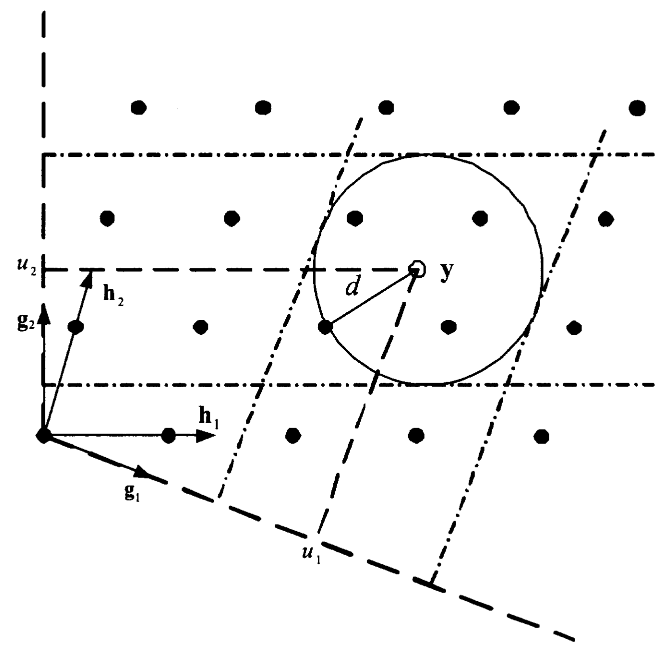

Figure 5.2 Geometrical representation of the local-searching algorithm.

In this subsection, it is defined a polygon searching area that contains the sphere determined by pre-decoder. In the space spanned by the column vector of the inverse or pseudo inverse of **H**, the center of the sphere, **u**, is determined by

$$\mathbf{u} = \mathbf{G}^H \mathbf{y}, \tag{5.6}$$

where $\mathbf{G}^H = (\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H$.

It is assumed that $\mathbf{s} = \mathbf{Hx}$ is a signal point in the received data space. Then projections of the noise free data **s** and the noisy data **y** into the space spanned by a given vector $\mathbf{g}_i$, the columns of matrix **G**, are simply $\mathbf{P}_{\mathbf{g}_i}\mathbf{x} = \frac{x_i}{||\mathbf{g}_i||^2}\mathbf{g}_i$ and $\mathbf{P}_{\mathbf{g}_i}\mathbf{y} = \frac{u_i}{||\mathbf{g}_i||^2}\mathbf{g}_i$, respectively. If the point of **Hx** is located within the sphere defined by pre-decoding, then it is obtained

$$\left|\frac{u_i - x_i}{||\mathbf{g}_i||}\right|^2 \leq d^2 \tag{5.7}$$

The region surrounded by the projection of the sphere boundary on each vector of $\mathbf{v}_i$, $i = 1, 2, \cdots N$, forms a polygon containing the sphere. Searching within the polygon, the ML solution can always be found.

Making looser bound approximation on each real-dimension of $\mathbf{v}_i$, It is obtained

$$\begin{aligned} -d &\leq Re\left\{\frac{u_i - x_i}{\|\mathbf{g}_i\|}\right\} \leq d \\ -d &\leq Im\left\{\frac{u_i - x_i}{\|\mathbf{g}_i\|}\right\} \leq d \end{aligned} \tag{5.8}$$

The candidate symbols for each transmit antenna can be determined independently by slicing. If there is only one candidate symbol for the i'-th transmit antenna, the signal component from i'-th transmit antenna can be subtracted directly to reduce the searching dimension. Figure 5.2 shows the geometrical representation of the polygon searching algorithm. In Figure 5.2, all vectors in this figure represents complex vectors. The proposed searching region in (5.8) is a polygon that includes the sphere involving the ML solution.

5.3.2 O-SIC Sphere Searching Detector

Both polygon and sphere local searching need to calculate the matrix $\mathbf{G}$, the inverse or pseudo-inverse of channel matrix $\mathbf{H}$. Moore- Penrose pseudo-inverse [53] of matrix is used in the original O-SIC algorithm. In sphere local searching algorithm, Cholesky's factorization needs to be compute to get up-triangular equivalent channel matrix. However, in a multi-step ML decoder composed of a O-SIC pre-decoder and a sphere local searching decoder, it not necessary to compute the inverse, Moore- Penrose pseudo-inverse and Cholesky's factorization separately. Hence the complexity can be reduce by modifying the O-SIC algorithm and sphere searching algorithm, such that the later step computation can utilize the earlier computational results.

At first, the channel inverse matrix $\mathbf{G}$ defined in (5.6) is calculated. Then the O-SIC algorithm is modified as follows to utilize the results of $\mathbf{G}$,

$$
\begin{aligned}
&\text{normalization:} && \mathbf{g}_k = \mathbf{g}_k / \|\mathbf{g}_k\|, \ k = 1, 2, \ldots, N \\
&\text{recursion:} && \text{for } i = 1, 2, \ldots, N \\
& && \text{strongest link search:} \\
& && k_i = \arg \max_{j \notin \{k_1, k_2, \ldots, k_{i-1}\}} \left\|\mathbf{g}_k^H \mathbf{h}_k\right\| \\
& && \mathbf{w}_i = \mathbf{g}_{k_i} \\
& && \text{filter-bank refining} \\
& && \mathbf{g}_j = \frac{\mathbf{P}_{\mathbf{w}_i}^{\perp} \mathbf{g}_j}{\left\|\mathbf{P}_{\mathbf{w}_i}^{\perp} \mathbf{g}_j\right\|}_{j \notin \{k_1, k_2, \ldots, k_i\}} \\
& && \text{decision statistic calculation} \\
& && z_{k_i} = \frac{\mathbf{w}_i^H \mathbf{y}}{\left|\mathbf{w}_i^H \mathbf{h}_{k_i}\right|} \\
& && \widehat{x}_{k_i} = Q(z_{k_i}) \\
& && \mathbf{y} = \mathbf{y} - \mathbf{h}_{k_i} \widehat{x}_{k_i} \\
& && \text{end of recursion}
\end{aligned}
\tag{5.9}
$$

where $\mathbf{g}_j$ is the j'-th column of $\mathbf{G}$, and $\mathbf{w}_i$ is the i'-th column of $\mathbf{W}$. It is easy to see that due to the nulling,

$$
\mathbf{w}_i^H \mathbf{w}_j = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases} \tag{5.10}
$$

and the spaces spanned by the column vectors of $\mathbf{G}$ and $\mathbf{W}$ are same. Since $\mathbf{G}^H$ is the inverse or pseudo-inverse of $\mathbf{H}$, $\mathbf{G}$ and $\mathbf{H}$ have same eigen vectors. The space spanned by column vectors of $\mathbf{G}$ and $\mathbf{H}$ are the same space. So the signal space spanned by $\mathbf{W}$ is equivalent to the space spanned by the column vectors of $\mathbf{H}$. Considering (5.10),

$$
\arg \min_{\mathbf{x} \in \mathcal{C}} \|\mathbf{y} - \mathbf{H}\mathbf{x}\|^2 = \arg \min_{\mathbf{x} \in \mathcal{C}} \left\|\mathbf{W}^H (\mathbf{y} - \mathbf{H}\mathbf{x})\right\|^2 \tag{5.11}
$$

Using (5.11), the received data vector and hypothetic signal vector in the signal space are spanned by the column vectors of $\mathbf{H}$, using orthogonal-normal column vectors of matrix $\mathbf{W}$. From Eq.(5.9), it can be shown that $\mathbf{w}_i^H \mathbf{h}_{k_j} = 0$ (if $i < j$), therefore, $\mathbf{R} = \mathbf{W}^H \widetilde{\mathbf{H}}$ is a lower triangular matrix, where $\widetilde{\mathbf{H}} = [\mathbf{h}_{k_1}\ \mathbf{h}_{k_2}\ \cdots\ \mathbf{h}_{k_N}]$. Sphere searching idea can be conveniently applied.

Define new vectors $\widetilde{\mathbf{y}} = \mathbf{W}^H \mathbf{y}$ and $\widetilde{\mathbf{x}} = \mathbf{W}^H \widetilde{\mathbf{H}} [x_{k_1}\ x_{k_2}\ \cdots\ x_{k_N}]$. If searching within the sphere of radius $d = \|\mathbf{y} - \mathbf{H}\mathbf{x}_p\|$ centered at the received received point, the candidate signal for k_i'th layer or the k_i'th transmit antenna is given by

$$\left| \widetilde{y}_i - r_{i,i} x_{k_i} - \sum_{l=1}^{i-1} r_{i,l} \widetilde{x}_{k_l} \right|^2 \leq d_{k_i}^2 = d^2 - \sum_{l=1}^{i-1} \left| \widetilde{y}_l - \sum_{j=1}^{l} r_{l,j} \widetilde{x}_{k_j} \right|^2$$

where $d_{k_i}^2$ can be calculated recursively. it can be observed that the search region for each layer depends on the estimation of previous layer and it is of smaller size than the searching region of above layers.

5.4 Ordered Group Wise Interference Cancellation

Although the proposed multi-step ML decoder efficiently reduces the computational complexity of ML detection for MIMO systems, compared to the global search based ML decoder. The detection complexity of the proposed scheme may still be high, when large number of transmit antennas are used and the system SNR is not high enough.

The idea of group detection was previously suggested for CDMA multi-user detection [54]. In [48], group detection was also used to balance the trade-off between performance and decoding complexity of V-BLAST system, but no detail grouping algorithm was presented. In this paper, it is proposed a new detection algorithm termed *ordered* group wise interference cancellation (O-GIC), which can be treated as an extended O-SIC for the V-BLAST system using large transceiver arrays. In the proposed O-GIC, the cancellation units are signals from a group of transmit antennas, instead of signals from just one transmit antenna as in O-SIC scheme. In each group, the ML detection is implemented by means of ML detector.

For simplicity, a 8×8 V-BLAST system is considered as an example. At first, the method mentioned in O-SIC-local searching algorithm is used to form the matrix $\mathbf{W}$ matrix. Then the matrix is partitioned into two sub-matrices, $\mathbf{W}_1 = [\mathbf{w}_1, \mathbf{w}_2, \mathbf{w}_3, \mathbf{w}_4]$ and $\mathbf{W}_2 = [\mathbf{w}_5, \mathbf{w}_6, \mathbf{w}_7, \mathbf{w}_8]$ using the grouping idea. Hence the group 1 sub-system is composed of the transmit antennas numbered k_1, k_2, k_3 and k_4. Using subspace projection, data for the sub-system of the group 1 is formulated as follows,

$$\mathbf{r}_1 = \mathbf{W}_1^H \mathbf{y} = \mathbf{H}_{e1}\mathbf{x}_1 + \mathbf{v}_1 \tag{5.12}$$

where $\mathbf{H}_{e1} = \mathbf{W}_1^H \mathbf{H}_1$, $\mathbf{H}_1 = [\mathbf{h}_{k_1}, \mathbf{h}_{k_2}, \mathbf{h}_{k_3}, \mathbf{h}_{k_4}]$, $\mathbf{x}_1 = [x_{k_1}, x_{k_2}, x_{k_3}, x_{k_4}]^T$, and $\mathbf{v}_1 = \mathbf{W}_1^H \mathbf{v}$.

Now the above mentioned O-SIC-local searching algorithm can be used to decode symbols contained in Eq. (5.12).

Before preceding to decode symbols contained in group 2, the signal component from group 1 antennas is cancelled (assuming no estimation error in group 1), hence, formulate the data for sub-system of group 2 as follows,

$$\begin{aligned} \mathbf{y}_2 &= \mathbf{y} - \mathbf{H}_1\hat{\mathbf{x}}_1 = \mathbf{H}_2\mathbf{x}_2 + \mathbf{v} \\ \mathbf{r}_2 &= \mathbf{W}_2^H \mathbf{y}_2 \end{aligned} \tag{5.13}$$

where $\mathbf{H}_2 = [\mathbf{h}_{k_5}, \mathbf{h}_{k_6}, \mathbf{h}_{k_7}, \mathbf{h}_{k_8}]$, and $\mathbf{x}_2 = [x_{k_5}, x_{k_6}, x_{k_7}, x_{k_8}]^T$.

5.5 Simulation Results and Analysis

Figure 5.3, 5.4 and 5.5 show that ML detection provides considerable performance advantage over sub-optimal detection, such as ZF and O-SIC detection. Since the ML detection is achieved, the performance of proposed multi-step detector should provide same performance, no matter which kind of pre-decoder is used. Using QPSK modulation, compared to O-SIC decoding, the proposed ML detection schemes obtain 2dB and about 5dB gain at the block error rate (BLER) as 0.1, in V-BLAST systems with transceiver array dimension of 2×2 and 4×4, respectively. When SNR increases, the performance gains are even more remarkable. It indicates that ML decoders provide more space diversity gain than ZF or O-SIC decoder.

In Figure 5.6, 5.7 and 5.8, the study on computational complexity of the proposed multi-step ML decoder using polygon searching algorithm is shown. The searching complexity can be parameterized by the signal points within the polygon, except the initial estimate point. O-SIC-local searching has lower complexity than ZF-local searching at a given SNR, because smaller d is provided by O-SIC pre-decoder than by ZF pre-decoder. When reasonable pre-decoding performance is obtain, the searching complexity can be very low. In a 2×2 system using QPSK modulation, at pre-decoding BLER of 0.1 (at about 20dB SNR for O-SIC, and 23.5dB SNR for ZF), the average polygon searching points per detection is only about 0.25 points. It means that averagely during 3 detections among every 4 detections, it is found that there are no other signal points within the polygon area, except the initial estimate, which is just the ML solution. When the modulation scheme is 16 QAM, at pre-decoding BLER as 0.1 (at 27.5 dB SNR for O-SIC) , the average searching points per detection is about 2.25. In a 4×4 system using QPSK modulation, at pre-decoding BLER of 0.1 (at about 20dB SNR for O-SIC), the average searching points per detection is only about 5 points, while the global searching need test 256 points.

Compare the performance of ML decoder using sphere searching with O-SIC decoder, in Figure 5.9. Figure 5.10 gives out the average decoding time of O-SIC sphere decoder normalized by the decoding time of traditional sphere decoder. Figure 5.10 shows that OSIC-sphere decoder have obvious advantage over traditional sphere decoder, when the number of transmit antenna in the system is big.

For a 8×8 V-BLAST system, the performance of the proposed scheme O-GIC is provided by Figure 5.11, in comparison with performance of O-SIC. At BLER as 0.1

and 0.01 more than 5 dB and 10dB gain are achieved by means of O-GIC technique, respectively.

5.6 Conclusions

The ML detection has great advantage on performance compared to ZF and O-SIC detection methods, in V-BLAST system. The proposed pre-decoder guided local-searching algorithm is an efficient decoding technique to achieve ML performance, while reducing the computational complexity. The complexity of local searching algorithm not only depends on the number of array elements and the order of modulation used, but also depends on the reception SNR. As SNR increases, the complexity of the proposed solution decreases rapidly. At reasonable high SNR, a local searching detector can provide ML detection performance with very low complexity. The proposed O-SIC sphere decoder has obvious advantage on computation than traditional sphere decoder, when the dimension of the system is large.

For large size V-BLAST system working at low SNR, it is proposed a O-GIC algorithm using group-wise ordered decoding and interference cancellation to further reduced the computation. Local searching ML detection method can then be implemented in each group. Hence, the decoding complexity is controlled, and large performance gain can still be achieved compared to O-SIC detection.

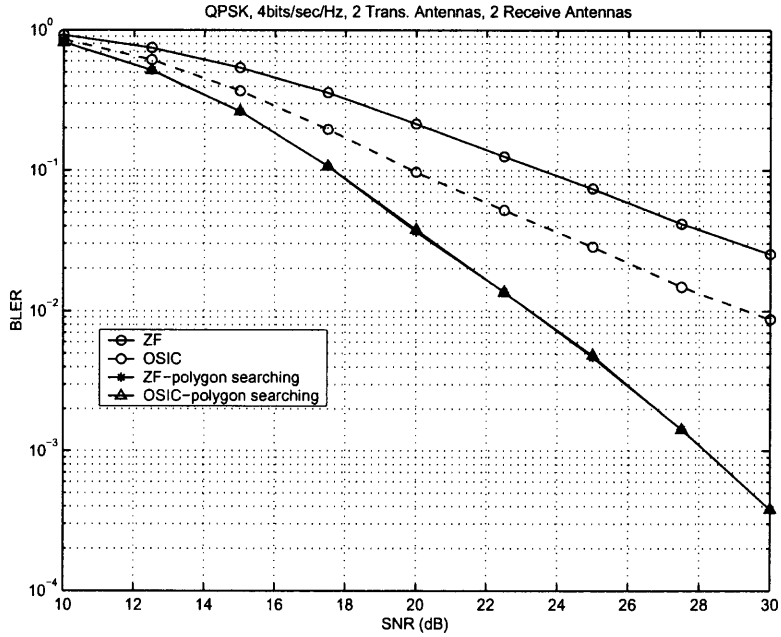

Figure 5.3 V-BLAST architecture, N=M=2, average block error rate of the QPSK modulation, 4 bits/s/Hz.

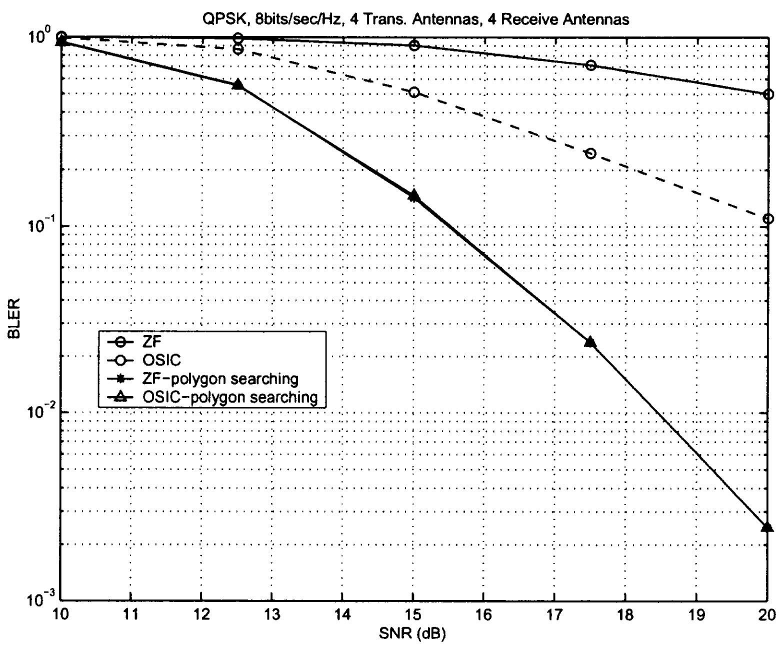

Figure 5.4 V-BLAST architecture, N=M=4, average block error rate of the QPSK modulation, 8 bits/s/Hz.

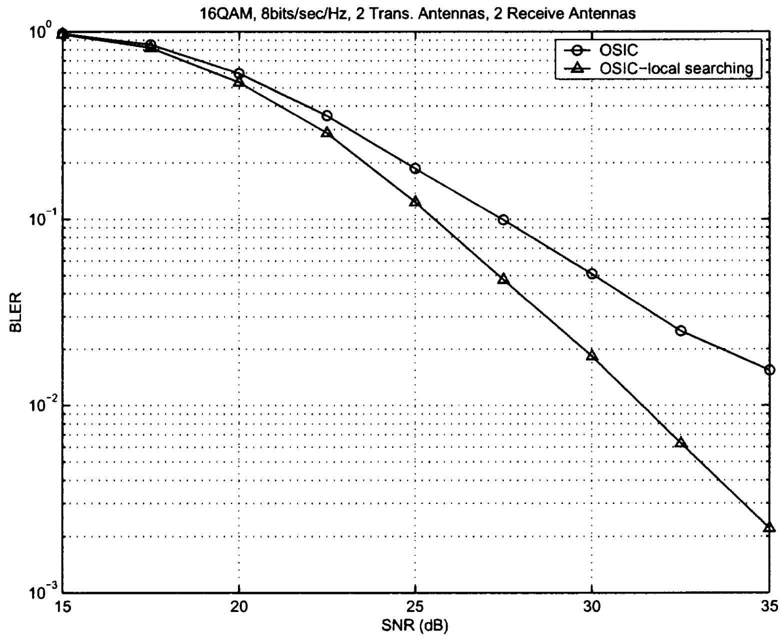

Figure 5.5 V-BLAST architecture, N=M=2, average block error rate of the 16QAM modulation, 8 bits/s/Hz.

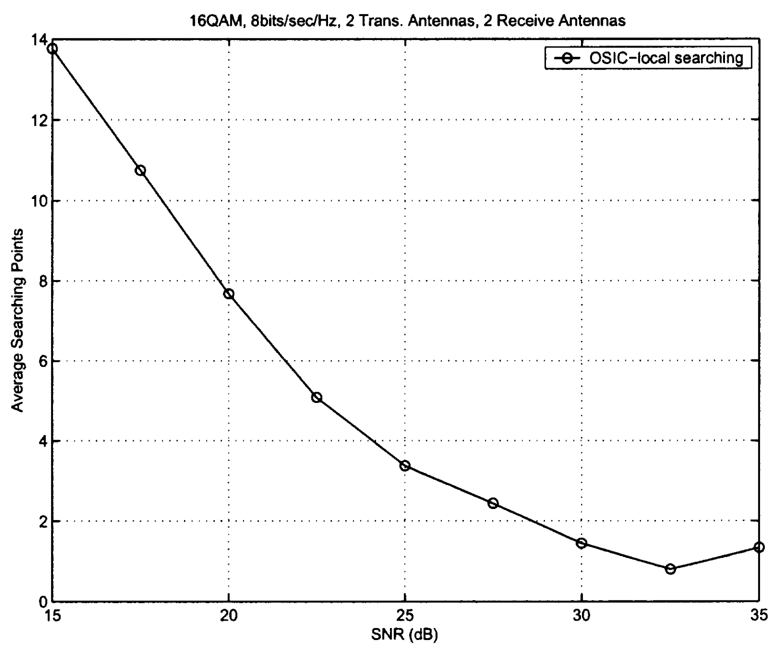

Figure 5.6 V-BLAST architecture, N=M=2, average searching points per detection using polygon local-searching algorithm, 16QAM modulation, 8 bits/s/Hz.

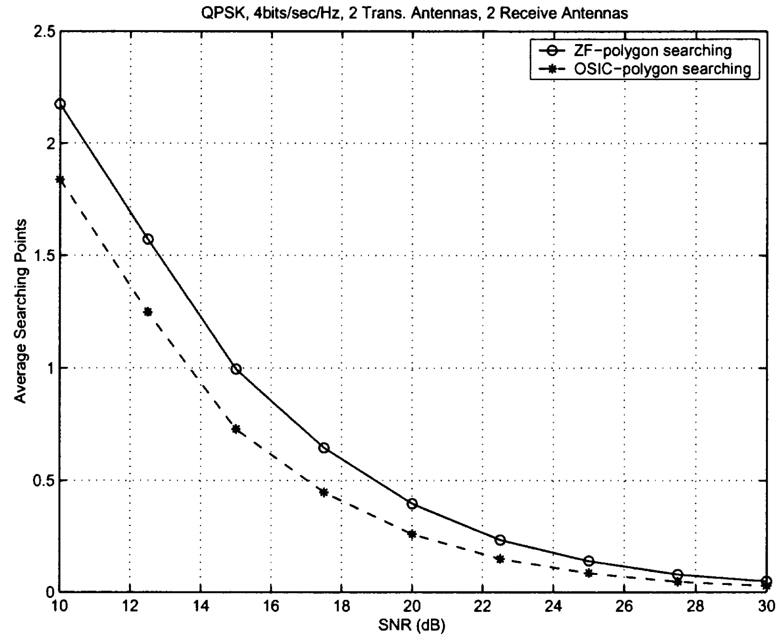

Figure 5.7 V-BLAST architecture, N=M=2, average searching points per detection using polygon local-searching algorithm, QPSK modulation, 4 bits/s/Hz.

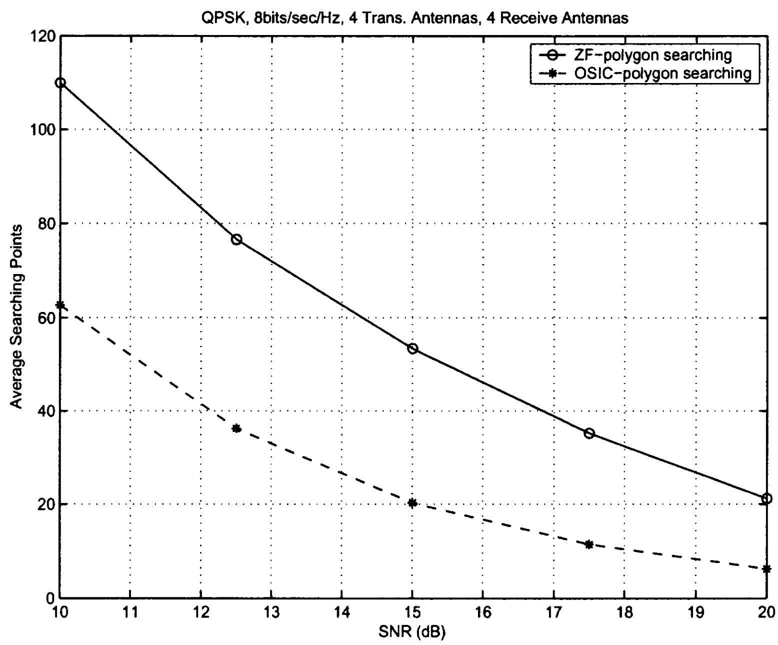

Figure 5.8 V-BLAST architecture, N=M=4, average searching points per detection using polygon local-searching algorithm, QPSK modulation, 8 bits/s/Hz.

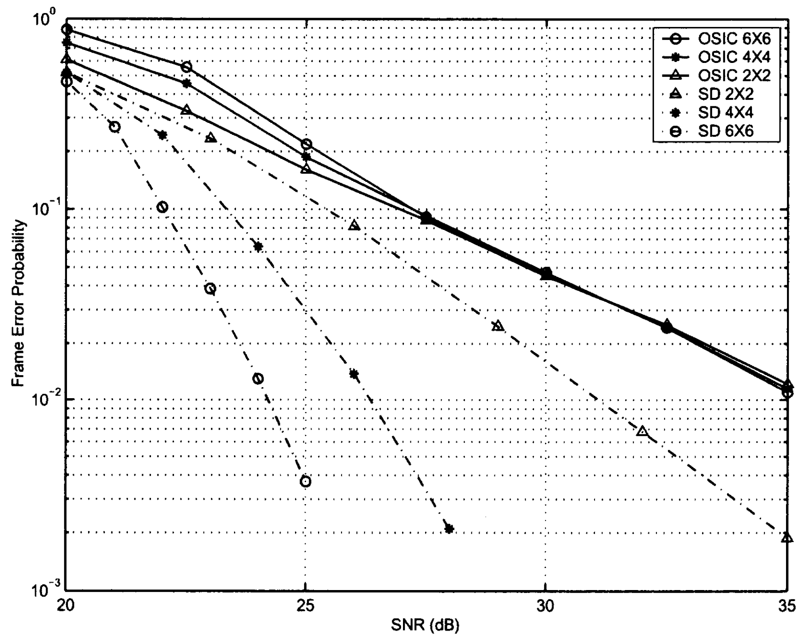

Figure 5.9 Average block error rate of 16QAM modulation. O-SIC sphere decoder is used for ML decoding.

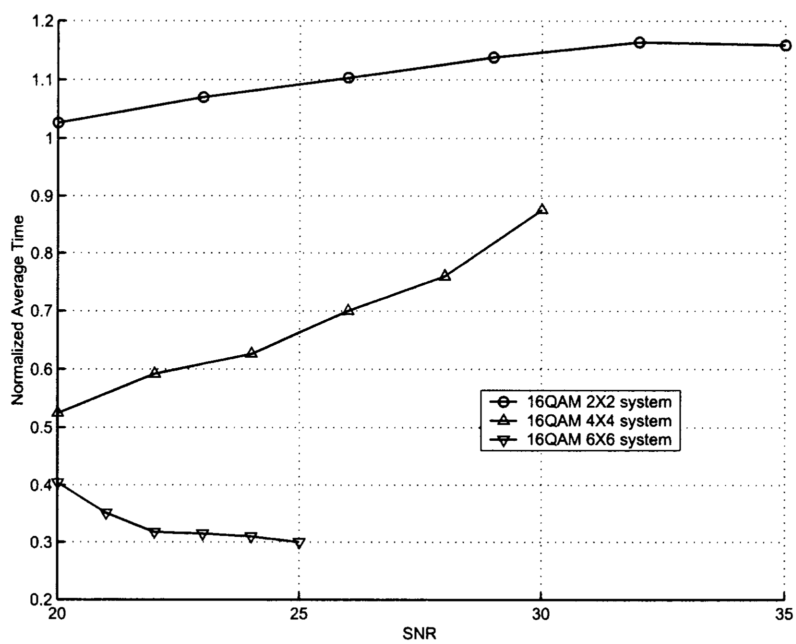

Figure 5.10 Average decoding time of O-SIC sphere decoder normalized by the decoding time of traditional sphere decoder. 16 QAM modulation was used.

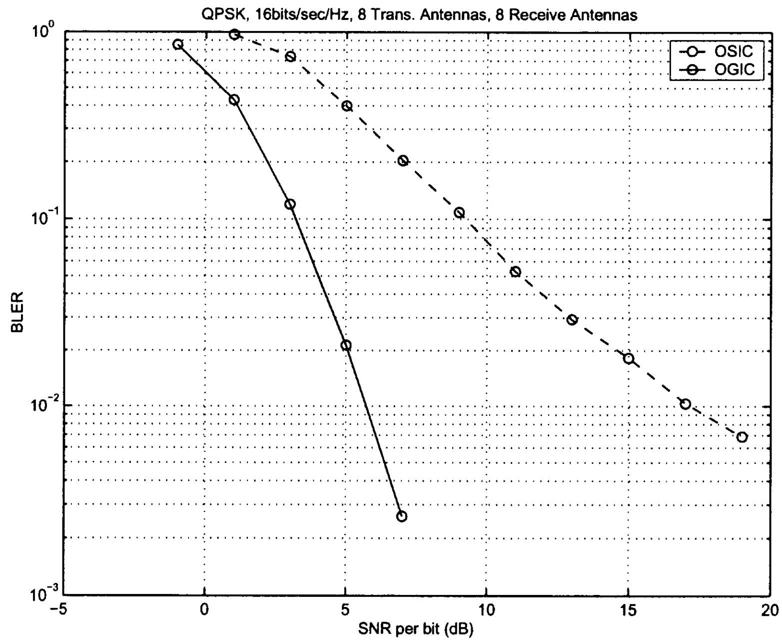

Figure 5.11 V-BLAST architecture, N=M=8, average block error rate of the QPSK modulation, 16 bits/s/Hz.

CHAPTER 6

CONCLUSIONS

6.1 A New Full-Rate Full-Diversity Orthogonal Space-Time Block Coding Scheme

Alamouti code which is used for wireless communication systems with two transmit antennas, is the only full-rate full-diversity linear OSTBC. To design full-rate full-diversity OSTBC for communication system with more than two transmit antennas, introducing non-linear coding may be an applicable method which has not been investigated. The work in Chapter 2 has demonstrated the existence of a full-rate full-diversity space-time orthogonal coding scheme for a QPSK system with four transmit antennas.

The new code shows comparable performance to that of full-diversity Q-OSTBC. Compared with Q-OSTBC, the obvious advantage of the new code lies in that the coded signals transmitted through all four transmit antennas do not experience any constellation expansion during data transmission. The study also shows that the proposed new code outperforms the half rate OSTBC and constellation unrotated Q-OSTBC when SNR increases. However, it should be pointed out that the receiver of the half-rate full-diversity codes can decode the symbols one by one, while that of the full-rate half-diversity quasi-orthogonal codes can decode the symbols pair by pair. This means that the full-rate full-diversity potential of the proposed space-time coding scheme is achieved at the cost of increased decoding complexity. For QPSK symbols, the decoding complexity of new orthogonal code is nearly 8 times of that of the quasi-orthogonal codes. The encoding complexity of the proposed orthogonal code is only a little higher than those of the OSTBC and Q-OSTBC.

Because of the unitary property of the new proposed STBC, it is easy to extend the coherent scheme to differential scheme by means of differential unitary space-time modulation (DUSTM). Without the CSI at both transmitter and receiver sides, the differential coded space-time system still obtains the full diversity gain, while a loss of about 3 dB in performance is observed in differential scheme, compared to coherent detection. [55] [56]

6.2 Fast Maximum Likelihood Decoding of Quasi-orthogonal Codes

A Q-OSTBC for a wireless communication system with 4 transmit antennas is originally presented to provide partial diversity gain and full rate transmission with reasonable decoding complexity. The later researches show that the diversity property can be improved by rotating the constellations of some signals in coding matrix. However, the original detection statistics is too tedious to find the diversity property of coding matrix by observing the decision statistics. As multiple receiving antennas are used, it is also difficult to using simple MRC technique to implement the joint ML detection

Studies show that similar to the orthogonal space-time codes, the fast ML decoder of quasi-orthogonal space-time code can be constructed by a simple linear decoupling process followed by MRC and reduced dimension ML based symbol decision. In this part, it is developed a systematic procedure to decouple the data of full-rate quasi-orthogonal code and obtain a new detection statistic for Q-OSTBC. The new detection statistic provides convenience to analysis the diversity properties of Q-OSTBCs and design full diversity Q-OSTBCs. A new 3/4 rate quasi-orthogonal code is also presented to facilitate the fast ML detection.

In the decoding scheme, the simple linear decoupling process is performed at each receiving antenna in parallel. The output at decouplers are combined using MRC technique and fed into parallel sphere decoder. The newly proposed decoding algorithm has very low computational complexity and can use parallel computing techniques to satisfy the real-time requirement. The complexity of newly proposed decoding algorithm is independent of the constellation size of modulation, so it is suitable for high data rate transmission.

Based on the study, Q-OSTBCs have many attractive features including

- By some simple symbol operation, such as constellation rotation, Q-OSTBC achieves full spatial diversity at high transmission rate (full rate for 4 transmit antennas and 3/4 rate for 6 transmit antennas) for any (real or complex signal constellation.
- Q-OSTBC does not require channel state information at transmitter.
- Q-OSTBC decoupling at receiver involves only simple linear processing operation (due to the group orthogonal spatio-temporal structure imposed by the code).

The main disadvantage is that in order to obtain the full spatial diversity, the constellation rotation operation expands the signal constellation at each transmit antenna.[57] [58]

6.3 Quasi-orthogonal Space-Time Block Coded Transceiver Systems Over Frequency Selective Wireless Fading Channels

The fast ML detection scheme for Q-OSTBC in Chapter 3 assumes a flat fading channel. In Chapter 4, the Q-OSTBC is extended to frequency-selective channels.

Applying the time reversal conjugate technique in combination with the Q-OSTBC scheduling, it is proposed a block based Q-OSTBC scheme to effective combat channel dispersion and frequency selective fading, and obtained diversity gain. The ML detector of the time-reversal Q-OSTBC can be constructed by a simple linear filter bank for space domain decoupling, a whitening filter bank and a vector MLSE detector. The coding rate of the time reversal Q-OSTBC is $L_B/4L$, where

L_B is length of the data block. The encoding complexity is very low. The decoding complexity is determined by the computational complexity of vector MLSE detector, which increases exponentially with the channel delay time. When channel delay time is short, the ML detection can be accomplished at low complexity. As channel delay time is long, the decoding complexity can be reduced by using some sub-optimum detector instead of the optimum MLSE detector.

The newly proposed MLSE receiver structure successfully exploits the spatial diversity offered by the multiple transmit antennas and the temporal diversity offered by the channel frequency selectivity.

However, since the computational complexity of vector Viterbi Algorithm in MLSE detection increases exponentially with the maximum delay of the frequency selective channel, the MLSE detection algorithm is not practical for the channel with long delay time. In chapter 4, it is also developed a fast decoding algorithm for TR-QOSTBC, which highly reduces the decoding complexity, especially for channels with large value of delay. During the new fast decoding, the 4 decoupled data streams are MMSE equalized, then the equalized data without inter-symbol interference are two by two feed to sphere decoder and decoded as 2×2 frequency non-selective system. The noise power of equalized data is normalized before it is decoded by sphere decoder. The sub-optimal fast decoding algorithm does not provide full space and frequency diversity,but good performance of BER and diverdsity gain is sill achevied.[59] [60]

6.4 Reduced Complexity Maximum Likelihood Detection for V-BLAST Systems

How to achieve the ML detection of MIMO system by controlled computational complexity is a very important topic in MIMO communication system. The ML detection has great advantage on performance compared to ZF and O-SIC detection methods, in V-BLAST system. The proposed pre-decoder guided local-searching algorithm is an efficient decoding technique to achieve ML performance, while reducing the computational complexity. The complexity of local searching algorithm not only depends on the number of array elements and the order of modulation used, but also depends on the reception SNR. As SNR increases, the more accurate pre-estimate efficiently reduces the local searching volume, so the complexity of the proposed solution decreases rapidly. At reasonable high SNR, a local searching detector can provide ML detection performance with very low complexity.

A new reduced complexity ordered sphere decoder (O-SD) is also proposed for maximum likelihood symbol decoding in a V-BLAST system, an important high-rate space-time data transmission scheme. Unlike the traditional sphere decoder using a zero-forcing (ZF) pre-decoder, the proposed O-SD uses an ordered successive interference canceller (O-SIC) as the pre-decoder. The proposed O-SIC sphere decoder has obvious advantage on computation compared to traditional sphere decoder, when the dimension of the system is large.

For large size V-BLAST systems working at low SNR, a O-GIC algorithm using group-wise ordered decoding and interference cancellation is proposed to further reduced the computation. Local searching ML detection method can then be implemented in each group. Hence, the decoding complexity is controlled, and large performance gain can still be achieved compared to O-SIC detection.[40] [61]

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