Distributed Space Time Block Coding and Application in Cooperative Cognitive Relay Networks

by

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 \bigodot by Walid Qaja, 2015

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I dedicate this thesis to my late brother, my parents, my brothers and sisters.

ABSTRACT

The design and analysis of various distributed space time block coding schemes for cooperative relay networks is considered in this thesis. Rayleigh frequency flat and selective fading channels are assumed to model the links in the networks, and interference suppression techniques together with an orthogonal frequency division multiplexing (OFDM) type transmission approach are employed to mitigate synchronization errors at the destination node induced by the different delays through the relay nodes.

Closed-loop space time block coding is first considered in the context of decode-and-forward (regenerative) networks. In particular, quasi orthogonal and extended orthogonal coding techniques are employed for transmission from four relay nodes and parallel interference cancellation detection is exploited to mitigate synchronization errors. Availability of a direct link between the source and destination nodes is studied. Outer coding is then added to gain further improvement in end-toend performance and amplify-and-forward (non regenerative) type networks together with distributed space time coding are considered to reduce relay node complexity. A novel detection scheme is then proposed for decode-and-forward and amplify-and-forward networks with closedloop extended orthogonal coding and closed-loop quasi-orthogonal coding which reduce the computational complexity of the parallel interference cancellation. The near-optimum detector is presented for relay nodes with single or dual antennas. End-to-end bit error rate simulations confirm the potential of the approach and its ability to mitigate synchronization errors.

A relay selection approach is also formulated which maximizes spatial diversity gain and attains robustness to timing errors. Furthermore, in this thesis, robust schemes for cooperative relays based on the modified distributed quasi-orthogonal space-time block coding (M-D-QO-STBC) and the extended-orthogonal space-time block coding (EO-STBC) schemes employing an OFDM data structure with cyclic prefix insertion at the source to combat the effects of time asynchronism. As such, this technique can effectively cope with the effects of timing errors. Finally, closed-loop extended orthogonal space-frequency-block coding (CL-EO-SFBC) is proposed for use within cognitive wireless relay networks. It is shown that the coding scheme together with the CL-EO-SFBC technique do not only guarantee a seamless and continuous transmission for the cognitive users without causing any interference to the primary users, but also improve the diversity gain as well as the array gain of the system. Moreover, the outage probability performance of a proposed cooperative cognitive relay network is evaluated. A closed form expression for the outage probability for cooperation over frequency selective fading channels is derived for both perfect and imperfect spectrum acquisition. The results showed the advantage in outage probability performance of the cooperative cognitive systems when the spectrum acquisition is perfect. In contrast, when the spectrum acquisition is imperfect the outage performance degrades significantly.

STATEMENT OF ORIGINALITY

The contributions of this thesis are concerned with cooperative relay based wireless communications systems with particular emphasis on distributed STBCs under imperfect synchronization. The contributions are supported by six published conference papers, together with one published IET journal in the communications area. These contributions can be summarized as follows:

In Chapter 3, investigates the cooperative strategy for distributed spacetime block coding (STBC) design which is based on a linear dispersion code and exploits outer convolutive coding for asynchronous cooperative relay networks. Furthermore, to overcome the lack of synchronization and achieve high performance, a parallel interference cancelation (PIC) detection scheme with distributed closed-loop extended orthogonal STBC (CL-EO-STBC) design with outer coding for four relay nodes and distributed closed-loop quasi orthogonal STBC (CL-QO-STBC) design with outer coding for two dual-antenna relay nodes are considered. Finally, the pairwise error probability (PEP) is analyzed to reveal the available cooperative diversity which can be utilized with distributed CL-EO-STBC and distributed CL-QO-STBC within cooperative relay networks. The results have been published in:

- W.M. Qaja, A.M. Elazreg, and J.A. Chambers, "Distributed space time transmission with two relay selection and parallel interference cancellation detection to mitigate asynchronism," *IEEE Conference EMS 2012, pp. 220-225, November 2012.*
- A.M. Elazreg, W.M. Qaja, J. A. Chambers, "PIC Detector with Coded Distributed Closed-Loop EO-STBC under Imperfect Synchronization", *IEEE conference SoftCOM 2012, pp. 1-5, September 2012.*

In Chapter 4, a novel detection scheme for decode-and-forward (DF) and amplify-and-forward (AF) asynchronous cooperative relay networks is proposed utilizing distributed closed-loop extended orthogonal space time block coding (CL-EO-STBC). These techniques are both designed to effectively remove the interference at the destination node induced by different time delays from the antennas of each relay node and achieve full cooperative diversity gain with unity data transmission rate between the relay nodes and the destination node. Moreover, a simple max-min relay selection scheme is proposed for cooperative relay networks to enhance the system performance. The best two relays are selected based on the overall path gain and the smallest timing error, then the CL-EO-STBC in a distributed manner is applied over the selected two relay nodes. The performance of the proposed schemes is studied for frequency flat fading channels and they are shown to be very effective to mitigate inter-symbol-interference (ISI) at the destination node with low detection complexity as compared to parallel interference cancellation (PIC) detection and to achieve full cooperative diversity

with unity data transmission rate between the relay nodes and the destination node. The results have been published in:

- W.M. Qaja, A.M. Elazreg, and J.A. Chambers, "Near-optimum Detection Scheme with Relay Selection Technique for Asynchronous Cooperative Relay Networks", *IET Communications, Volume 8,* no. 8, pp. 1347-1354, 22 May 2014
- W.M. Qaja, A.M. Elazreg, and J.A. Chambers, "Near-optimum detection for use in closed-loop distributed space time coding with asynchronous transmission and selection of two dual-antenna relays", Wireless Conference (EW), Proceedings of the 2013 19th European, pp. 1-6, April 2013

In Chapter 5, a novel robust scheme for two dual-antenna relay nodes to employ in cooperative relay networks without the requirement of exact synchronization between relay nodes is proposed. The design exploits distributed modified quasi orthogonal space-time block coding (M-QO-STBC) type of transmission that can achieve full cooperative diversity and code gain distance, and distributed extended orthogonal spacetime block coding type of cooperative transmission (D-EO-STBC). Orthogonal frequency division multiplexing (OFDM) is implemented at the source node. A cyclic prefix (CP) is added at the source and relay nodes to combat the effects of random delays at the relay nodes. The relays operate in a simple amplify-and-forward (AF) mode. In this chapter, a narrowband system, where the two-hop channels are assumed to be flat Rayleigh fading channels is initially considered with the M-QO-STBC transmission technique. Next, a new low complexity one-bit feedback scheme based on the selection of the cyclic phase rotation scheme is proposed for broadband systems, where the two-hop channels are assumed to be frequency-selective Rayleigh fading channels, and the D-EO-STBC transmission technique is employed. This approach attains unity rate over each hop in the network and full cooperative spatial diversity. Simulation results are included to confirm that these schemes provide better performance as compared with previous schemes. The results have been published in:

- M. A. Manna, W. M. Qaja, and J. A Chambers, "OFDM-based Modified Quasi-Orthogonal Space-Time Scheme for Use in Asynchronous Cooperative Networks with Relay Selection", 20th International Conference on Telecommunications (ICT), pp. 1-4, 2013
- W. M. Qaja, A.M. Elazreg, and J. A Chambers, "Closed-Loop EO-STBC Based on Selection of Cyclic Rotation for Asynchronous Cooperative Relay Networks over Frequency Selective Wireless Fading Channels", *European Modelling Symposium (EMS2013)*, pp. 231-236, 2013

In chapter 6, closed-loop extended orthogonal space-frequency-block coding (CL-EO-SFBC) is proposed for use within cognitive wireless relay networks. To exploit the available spectrum opportunities within the EO-SFBC, a spectrum indicator matrix is used in the development of the relay transmission scheme. The approach has the flexibility to dynamically adapt the code matrix to the number of available relays. The performance of the scheme is evaluated in terms of end-to-end frame error rate. Furthermore, the probability density function of the multi-path links is modeled in the time domain with an Erlang distribution function. The analytical expressions for the probability density function and cumulative density function of the end-to-end signal-tonoise ratio are obtained for one and two cognitive relay nodes and multi-path channel lengths of two and three. Moreover, expressions for outage probability are determined for a perfect and imperfect spectrum acquisition. The results have been published in:

• W.M. Qaja, and J.A. Chambers, "Closed-loop extended orthogonal space-time coding for cognitive wireless relay networks", 18th European Wireless Conference, pp. 1-4, 2012

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LIST OF ABBREVIATIONS

MIMO	Multi-Input Multi-Output
VAA	Virtual Antenna Array
\mathbf{CL}	Closed-Loop
OL	Open-Loop
STBC	Space Time Block Code
A-STBC	Alamouti Space Time Block Coding
O-STBC	Orthogonal Space Time Block Coding
D-STBC	Distributed Space Time Block Coding
QO-STBC	Quasi Orthogonal Space Time Block Coding
EO-STBC	Extended Orthogonal-STBC
CL-STBC	Closed-Loop STBC
CL-QO-STBC	Closed-Loop QO-STBC
CL-EO-STBC	Closed-Loop EO-STBC
M-QO-STBC	Modified QO-STBC
OFDM	Orthogonal Frequency Division Multiplexing

PIC	Parallel Interference Cancellation
ISI	Intersymbol Interference
PEP	Pairwise Error Probability
IET	The Institution of Engineering and Technology
IEEE	Institute of Electrical and Electronics Engineers
СР	Cyclic Prefix
EW	Europe Wireless
ICT	International Conference on Telecommunications
EMS	European Modelling Symposium
RS	Relay Selection
AF	Amplify and Forward
DF	Decode and Forward
BER	Bit Error Rate
FER	Frame Error Rate
CWER	Codeword Error Rate
FFT	Fast Fourier Transform
IFFT	Inverse Fast Fourier Transform
DFT	Discrete Fourier Transform
IDFT	Inverse Discrete Fourier Transform
PS	Perfect Synchronization

ML	Maximum Likelihood
BPSK	Binary Phase Shift Keying
QPSK	Quadrature Phase Shift Keying
$4\mathrm{G}$	Fourth Generation
Wi-Fi	Wireless Fidelity
WiMAX	Worldwide Interoperability for Microwave Access
LTE	Long Term Evolution
SNR	Signal-to-Noise Ratio
CSI	Channel State Information
dB	Decibel
\mathbf{CR}	Cognitive Radio
PU	Primary User
\mathbf{SU}	Secondary User
DT	Direct Transmission
CGD	Code Gain Distance
PAPR	Peak-to-Average Power Ratio
\mathbf{RF}	Radio Frequency
LD	Linear Dispersion
LS	Least Squares
P/S	Parallel to Serial

S/P	Serial to Parallel
PhD	Doctor of Philosophy
SIC	Successive Interference Cancelation

MATHEMATICAL NOTATIONS

m	Subcarrier index
q	Number of the PIC iteration
М	Size of constellation
Ν	Duration of one OFDM symbol
Р	Total transmitted power
\mathbf{S}	Code matrix of STBC
R	Number of relay nodes
Т	Number of symbol transmissions
N_T	Number of transmit antennas
N_R	Number of receiver antennas
M_R	Number of antennas on each relay node
E_s	Average power of the source
E_b	Energy per bit

N_o	Noise spectral density
D_g	Diversity gain
C_g	Coding gain
G_p	Guard period
N_S	Number of transmission symbols
R_s	The set of best relay selection node
N_P	Number of transmission periods
I_{Bits}	Input bits of convolutional encoder
O_{Bits}	Output bits of convolutional encoder
P_1	Transmitted power at the source node
P_2	Transmitted power at the relay nodes
l_{cp}	Length of cyclic prefix
$ au_{max}$	Maximum of possible relative timing error
$R_{\mathbf{B}}$	The rank of matrix \mathbf{B}
$N_T \times N_R$	Spatial diversity
Δ	Grammian matrix
λ	Total channel gain
λ_c	Conventional channel gain
λ_f	Feedback performance gain
L	Angle of a complex number

σ_s^2	Variance of transmitted signal
σ_n^2	variance of the AWGN noise
σ_r^2	variance of the received signal
CN(0,1)	Circularly symmetric complex Gaussian distribution
	with zero mean and unit variance
\otimes	The Hadamard product
j	$\sqrt{-1}$
$ au_k$	Timing misalignments between relay nodes
β_k	Reflect the impact of timing misalignments
$h_k(-l)$	Coefficient to reflect ISI from previous symbols
$g_k(-l)$	Coefficient to reflect ISI from previous symbols
log_2	Base-2 logarithm
I_m	$m \times m$ Identity matrix
$0_{m \times n}$	$m \times n$ Matrix with all zero entries
$k \in R$	k is an element of R
$\mathbf{E}(.)$	The statistical expectation operator
$\mathbf{det}(.)$	The determinant operator
$\xi(.)$	Time reversal of the signals
$P_e(.)$	Average probability of error

.	Absolute value of a complex number
.	The Euclidean norm
$ \cdot _F$	The Frobenius norm
$\min\{.\}$	Select the minimum value
$\max\{.\}$	Select the maximum value
$(.)^T$	Transpose operator
$(.)^{*}$	Complex conjugate operator
$(.)^H$	Hermitian transpose operator
$\Re\{.\}$	Real part of a complex number
$\arg(a_1,\cdots)$	(a_n) Argument of a_1, \cdots, a_n

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INTRODUCTION

Cooperative relay communications has recently gained much attention in academic and industrial advanced wireless research centers across the globe due to its potential to enable efficient solutions for challenging problems in wireless communications. In fact, this technique, through distributed transmission, can achieve the same diversity gain benefits as conventional point-to-point multiple-input multiple-output (MIMO) systems without requiring multiple-antenna terminals at a single terminal.

In this chapter, conventional MIMO systems will firstly be presented. Then, the design of space-time codes and their enhancement for cooperative relay networks will be briefly discussed. The most common and efficient coding techniques to exploit spatial diversity within MIMO systems are space-time codes (STCs) which are also the origin for distributed STCs. A concise background for STCs will be given and the milestone works in this field will be described. This will be followed by discussion of the concept of maximum likelihood (ML) decoding and pairwise error probability (PEP). The definition of uncoded and coded transmission will be introduced thereafter. Then, the principle of parallel interference cancelation (PIC) will be presented, followed by the principle of orthogonal frequency division multiplexing (OFDM) type transmission and its advantages and disadvantages. A review of synchronous cooperative relay networks and cooperative operation will be next provided, then, the literature relevant to asynchronous cooperative relay networks will be presented. Cooperative relays can also be exploited in cognitive radio, therefore, a quick introduction to this field and cooperative cognitive networks in particular will be included.

This background material is essential for the reader to appreciate the research which relates to this thesis.

The aims, objective and structure of the thesis are defined and presented.

1.1 Conventional MIMO Systems

In the past few decades, wireless communication technologies have witnessed an exponential growth and have become part of everyday wireless applications. An important technology for wireless networks is multiple antenna systems which are also known as MIMO systems (Figure 1.1) due to their ability to increase the system reliability and capacity without requiring additional bandwidth or transmit power [10]. The



Figure 1.1. MIMO wireless communication system diagram

benefits of using multiple antennas resulted in MIMO wireless technology being exploited in many wireless communication standards such as
the Wi-Fi (IEEE 802.11) standard, the WiMAX (IEEE 802.16) standard, and are a major focus for 4^{th} generation (4G) and long-term evolution (LTE) cellular systems [11].

The three main advantages of a point-to-point MIMO system are multiplexing gain, diversity gain and array gain [10] and [11]. A MIMO system can offer linear increase in the capacity or transmission data rates proportional to the number of transmit-receive antennas pairs or the minimum number of transmit or receive antennas when the channels between all antennas are uncorrelated [12], [13] and [14], i.e. multiplexing gain. On the other hand, MIMO systems have the ability to obtain high diversity gain. Diversity gain is equal to the number of independent channels in the multiple antennas system [10]. The diversity gain indicates how fast the probability of error decreases with an increase in the signal strength [15] as shown mathematically in the next section. Finally, MIMO systems can achieve an array gain which means the average increase in signal to noise ratio at the receiver [10]. The average increase in signal power is proportional to the number of receive antennas.

However, the requirement of multiple-antenna terminals increases the system complexity and the separation between the antennas increases the terminal size. Also, MIMO systems suffer from the effect of path loss and shadowing, where the path loss is referred to the signal attenuation between the source and destination nodes due to propagation distance, while the shadowing is the signal fading due to objects obstructing the propagation path between the source and destination nodes [16]. These different problems limit MIMO systems functionality and applicability which challenge researchers to look for another inno-

vative technology, and hence cooperative communications has emerged as a new paradigm that can offer effective solutions for the aforementioned problems.

1.2 Cooperative Relay Systems

Unlike traditional point-to-point MIMO systems, a cooperative relay system (Figure 1.2) allows different nodes in a wireless network to share their antennas based on cooperation protocols [17] and [18]. Such cooperative nodes can be regarded as a distributed antenna array (i.e. virtual MIMO) where each node becomes part of this virtual array. This new communication paradigm has become a powerful technique



Figure 1.2. Basic structure of a parallel cooperative relay network with two phases for the cooperative transmission process.

that can achieve the same gain benefits of point-to-point MIMO systems whilst avoiding some of their drawbacks. In fact, it promises significant improvements in the system reliability and capacity and in service coverage without additional bandwidth or transmit power [19]. It has recently been adopted for various new wireless systems such as 3GPP LTE-Advanced [20]. Also, it has been considered in other wireless system standards such as WiMAX standards (IEEE 802.16j and IEEE 802.16m) [21] and Wi-Fi standards (IEEE 802.11s and IEEE 802.11n) [22] and [23]. This new technology has been underpinned by Laneman's contributions in this area in 2000 and beyond [17] and [24] which specifically introduced different relaying protocols and proved that significant system performance and outage gains can potentially be achieved.

The main feature of a cooperative relay system is its ability to obtain high spatial diversity gain. In fact, a cooperative relay system achieves a new form of spatial diversity which is known as cooperative diversity. This feature provides the ability to overcome the detrimental effects of severe fading in the wireless channel [17]. Thus by having several intermediate relays between the source and destination that forward copies of the same information in parallel via independent channels with or without the information received from the direct path, will result in diversity gain. This gain comes from the fact that as the number of independent paths carrying the same information between the source and destination increases, the probability of all of them being in fade decreases [24], [25]. Therefore, diversity gain can be computed as the number of independent channels in the cooperative relay system, which depends on the number of the relay nodes and the environment [25]. So, in a frequency flat channel, the maximum diversity gain equals G_d $= N_s \ge N_r \ge N_d$ where N_s , N_r and N_d are the number of single-antenna source nodes, relay nodes and destination nodes, respectively.

Increased diversity gain leads to improvements in the system performance. The diversity gain indicates how fast asymptotically the probability of error decreases with an increase in the signal strength typically measured by signal-to-noise ratio (SNR) [10], [15]. The diversity gain or diversity order, G_d , in terms of error probability is given by [16]

$$G_d = -\lim_{SNR \to \infty} \frac{\log(P_e(SNR))}{\log(SNR)}$$

where P_e denotes the average probability of error at average SNR over the randomness of the channel, the noise, and the data transmission. The diversity gain G_d can be defined as the slope of error probability curve in term of received SNR in log-log scale.

Recently, the area of cooperative communications has sparked much attention among researchers [24], [26], [18] and [25]. If the transmission from all relay nodes arrives at the destination node at the same time, then the term synchronous cooperative relay networks is used. Otherwise, the expression asynchronous cooperative relay networks is used. Among the several types of cooperative relay networks, this thesis considers asynchronous two-stage relay networks with the assumption that all nodes are equipped with a single or two half-duplex antenna at the relay nodes.

1.2.1 Design of Space-Time Codes and their Enhancement for Cooperative Relay Networks

In recent years it has been found that the framework of distributed STBC also plays an important role in coding for cooperative relay networks. Distributed STBC refers to a cooperative strategy where the conventional STBC for co-located antennas is implemented between the relay nodes in a distributed manner. More specifically, when the relay nodes receive the signals from the source node, they are linearly processed and then broadcast to the destination node in the form of a space time codeword [27] and [28]. Distributed STBC in a cooperative relay network allows the relay nodes to maximize cooperative diversity gain without the availability of channel state information (CSI) at the relay nodes, similar to the conventional STBC in point-to-point MIMO systems, where transmit diversity is exploited without the need of CSI at the transmit antennas. As a consequence, it is well known that the transmission reliability of the source signals over cooperative relay networks utilizing distributed STBC can be significantly improved. Current distributed STBCs include distributed orthogonal STBC (O-STBC) and distributed quasi orthogonal STBC (QO-STBC) [28]. Distributed O-STBC with full transmission rate design and complex elements in its transmission matrix is impossible for more than two relay nodes. The only example of full data transmission rate, full cooperative diversity, complex STBC using orthogonal design is distributed Alamouti STBC (A-STBC) [29]. Consequently, distributed QO-STBC was proposed in which the constraint of orthogonality is relaxed to achieve full data transmission rate. In general distributed QO-STBC does not achieve full cooperative diversity provided in proportion to the number of transmitting relay nodes and it has pair-wise decoding complexity as compared to distributed O-STBC [28]. Therefore, a number of closed-loop (CL STBC) techniques were proposed to provide full data transmission rate and full diversity gain with simple-wise decoding complexity in a point-to-point MIMO system [30], [31] and [32]. Basic definitions of closed-loop and open-loop systems are given in the following section.

1.2.2 Closed-Loop Versus Open-Loop System

When the relay nodes do not have any knowledge about the channel coefficients and the destination node is capable of estimating the channel coefficients accurately through some pre-defined training data sequence and then use the CSI for decoding operations, the system is called an open-loop system as shown in Figure 1.3 (a). However, if the relay node can have access to the channel information improvement in the error rate performance may result and this should be exploited provided complexity and system overhead is minimized. Therefore, in some communication systems the relay nodes are assumed to obtain knowledge of the channel condition through a feedback link from the destination node to the relay node, in this case it is called a closed-loop system as depicted in Figure 1.3 (b). In a closed-loop system, the gain achieved



Figure 1.3. Basic structure of (a) open-loop cooperative relay network and (b) closed-loop cooperative relay network with feedback link from the destination node to the relay node.

by additional processing at the relay node and the destination node can be termed as array gain. Similar to diversity gain, the array gain results in an increase in average SNR. In most practical applications, the amount of feedback information required from the destination node to the relay node should be kept as small as possible due to limited available feedback bandwidth. Therefore, the feedback information should be quantized into levels, and these levels can be fed back to the relay node to improve the system performance [30]. In particular, the feedback can be made available to all relay nodes through a separate feedback channel which is practically achievable in bi-directional control channels present in many communication systems. Thus, in this thesis, distributed CL STBC that achieves full data transmission rate and full cooperative diversity gain with linear decoding for asynchronous twostage relay networks with and without outer coding is investigated and also a new closed-loop extended orthogonal STBC (EO-STBC) scheme with one-bit feedback information is proposed to enhance the system performance of cooperative relay networks as compared to previous closed-loop methods [30] and [31] in asynchronous distributed manner.

1.3 Impairments on Cooperative Relay Channels

1.3.1 Wireless Fading Channels

One of the most challenging phenomena in wireless communication channels is the signal attenuation caused by the fading nature of the channels. The transmitted signal in a wireless network usually reaches the receiver node via multiple paths and these paths change with time due to the mobility of the user nodes and/or reflectors in the environment [33]. The changing strength of each path and the changing interference between these paths result in fading. Fading can be characterized on large and small time scales. Large fading is a result of movement distance large enough to cause total variation in the overall path between the source node and the receiver nodes. However, the short term fluctuation in the signal amplitude effect by movement of the user over distances of a small number of wavelengths is called small-scale fading. Small-scale fading itself can be categorized as frequency flat or frequency selective fading [34]. In frequency flat fading, the channel has a constant gain and linear phase response over a bandwidth which is greater than the bandwidth of the transmitted signal. Therefore, all frequencies of the transmitted signal experience the same channel condition. Nevertheless, in frequency selective fading, the channel possesses a constant gain and linear phase over a bandwidth that is smaller than the signal bandwidth, inter-symbol-interference (ISI) exists and the received signal at the received node is distorted [7].

1.3.2 Interference Between Cooperative Relay Nodes

The use of multiple relaying nodes provides high cooperative diversity gain thereby improving robustness against channel impairments, in order to maximize the cooperative diversity gain of cooperative relay networks in proportion to the number of transmitting relay nodes [19], the relay nodes are scheduled to transmit simultaneously with perfect synchronization among the cooperative relay nodes at the symbol level, which means that the timing, carrier frequency and propagation delay relay nodes are identical as shown in Figure 1.4 (a) [26]. However, perfect synchronization in cooperative relay networks does not exist in practice. To achieve exact synchronization among the relay nodes positioned at different locations, which are probably subject to movement, each having its own oscillator, is difficult or impossible to achieve [35].



Figure 1.4. Basic structure of the received signals at the destination node from the relay nodes, (a) shows the received signals which arrive at the destination node at the same time instants (perfect synchronization) (b) shows the received signals arrive at the destination node with different time delay τ_{1R} (imperfect synchronization).

As shown in Figure 1.4 (b), the major synchronization issue is the time delay τ_{1R} of signals, when they arrive at the destination node in the next hop, where R is the number of relay nodes in the network. Propagation delays may be unknown to them, while transmission time instants may be different at the destination node. This lack of synchronization results in ISI between the received signals from the relay nodes at the destination node. As such the channel becomes dispersive even under a flat fading environment; and will lead to substantial performance degradation. Mitigating this problem is the main research focus of this thesis.

1.4 Cognitive Radio

The growth in wireless communications initiated by the variety of wireless applications and systems being used nowadays has introduced an important challenge for the adaption of new spectrum allocation and utilization methods. The scarcity of licensed spectrum bands, controlled by governmental policies, makes such challenges even more difficult and complicated. However, the utilization of radio spectrum has been found to be not optimal and remains unutilized most of the times. Hence, Mitola [36] proposed cognitive radio (CR) to optimize the de-

been found to be not optimal and remains unutilized most of the times. Hence, Mitola [36] proposed cognitive radio (CR) to optimize the demand for the radio spectrum and utilize the unutilized licensed spectrum. CR arose as a promising solution to spectral crowding problem by allowing the unlicensed (secondary) user to opportunistically utilize the frequency bands that are not heavily occupied by the licensed (primary) user. Figure 1.5 below illustrates the basic CR behavior to train itself about the radio environment. The main two challenges to the success of cognitive radio include the primary user (PU) detection and the transmission opportunity exploitation [37] [38]. The detection of PU (spectrum sensing) leads to the detection of spectrum holes that can be defined as the unoccupied spectrum bands that are licensed to the PU. The cognitive users have the ability to exploit the opportunity of transmission (spectrum sharing) to improve their performance without causing any interference to the PU. Moreover, unlicensed users should posses the ability of measuring, sensing, learning, and the awareness of the surrounding radio environment for the reliable detection of the spectrum in order to check for the presence and activity of PU's [37] as well as adapting itself for the exploitation of unused spectrum. This concept has recently been developed to be fully aware of the surrounding environment and the primary users, and thereby further improve the efficiency of spectrum utilization [39]. Recently, several IEEE 802 standards for wireless systems have considered cognitive radio systems such as IEEE 802.22 standard [40] and IEEE 802.18 standard [41]. Cognitive radio classification can be performed according to the utiliza-



Figure 1.5. Key concepts in cognitive radio spectrum sensing

tion of spectrum band, the spectrum availability, or the interference as shown in Figure 1.6. The classification of CR based on the utilization of spectrum band can be divided into ideal CR and spectrum sensing CR [1]. Ideal CR is supposed to be aware of the operating parameters of all radios in its environment and can intelligently decide and use any unutilized spectrum band. Spectrum sensing CR, on the other hand, is supposed to notice and observe the spectrum bands of the PU before transmission [2]. Moreover, depending on the availability of spectrum bands, CR can be divided into licensed or unlicensed band CR [1] [2]. In the first category, i.e. licensed band CR, the cognitive system is able to utilize the spectrum band assigned to the licensed users, whereas, the unlicensed band CR is allowed only to utilize the unlicensed part of the spectrum bands that are available for secondary users only. Therefore, there is no need for the CR to sense the entire spectrum before the secondary user (SU) use the channel. An example of unlicensed band CR is IEEE.802.19 [42].

In CR, a secondary user (SU) can only borrow spectrum if it does not generate interference to the PU. Generally, there are two common approaches to avoid interference namely spectrum overlay (interference avoidance), and spectrum underlay (interference control) [1–6]. In the



Figure 1.6. Classification of cognitive radio [1–6]

underlay approach or interference control approach (Figure 1.7(a)), SU utilizes the radio spectrum at the same time with the PU provided that its transmission power is below the noise floor of the PU in order to avoid any interference to the PU. The advantage of this approach is the ability of the SU to transmit at any time without performing the sensing operation for detecting spectrum holes. However, the power control techniques related to this approach are likely to have a large degree of complexity. In the overlay approach (Figure 1.7(b)), the SU can access the detected spectrum holes only and utilize the unoccupied spectrum for transmission causing no interference to the PU. This approach enjoys interference-free communication between the primary and secondary users. However, this approach requires the continuous sensing of spectrum holes which is a difficult task due to the range of potential modulation schemes as well as the problem of hidden terminals, not to mention the fast time required for sensing and detecting these unoccupied spectrum bands [1]. The work introduced in this



Figure 1.7. Approaches for avoiding harmful interference

thesis related to this subject assumes an overlay approach during the transmission of the cognitive system, whereas, the sensing and detection operation is outside of the scope of the research.

The inherited fading phenomena of wireless channels limits the service reliability and coverage of the wireless communication systems. A potential solution to this problem is by using cooperative relaying networks which is a technique exploited by a relay network to cope with the challenges of cognitive radio [43] [44] [45]. Cognitive technology can be deployed over the intermediate wireless relay nodes in order to achieve reliable communication and coordinate the spectrum sharing among primary and secondary users. Moreover, cooperative cognitive relaying is about fully utilizing the dynamic spectrum through a number of cognitive relay nodes in order to achieve reliable communication and seamless transmission without causing any interference to primary users when they are present.

At the end of this thesis, the combination between a cooperative relay network and cognitive radio where the intermediate relay nodes are equipped with cognitive radios will be presented. This network will be referred to as a cooperative cognitive relay network considering spectrum sharing and transmission robustness.

1.5 Challenges and Motivations of the proposed research work

Cooperative multiple-input multiple-output technology allows a wireless network to coordinate among distributed antennas and achieve considerable performance gains similar to those provided by conventional MIMO systems. It promises significant improvements in spectral efficiency and network coverage and is a major candidate technology in various standard proposals for the fourth and fifth generation wireless communication systems. In fact, conventional point-to-point MIMO systems suffer from four main challenging problems namely, system complexity due to the requirement of multiple antennas; large terminal size due to the requirement of positioning the multiple antennas apart from each other otherwise the problem of correlated channels will adversely affect the system performance; signal attenuation due to path loss; and lastly severe fading due to shadowing effects. This thesis addresses these aforementioned challenging problems by exploiting different single-antenna, as well as dual-antenna, terminals within a wireless network to cooperatively form a virtual antenna array (VAA) that can achieve the same diversity gain as in a conventional MIMO systems in proportion to the number of transmitting relay nodes together with full data transmission rate over each stage of transmission and low decoding and detection complexity either at the relay nodes or the destination node. Furthermore, the main scope of this thesis is to achieve the objective of eliminating the issue of imperfect synchronization among the cooperating relay nodes over frequency flat/selective fading channels.

The introduced work in this thesis has been inspired by the contributions of [32], [31], [30] and [46] on STBCs as applied to MIMO systems; [47], [27] and [28] on distributed STBCs for synchronous cooperative relay networks; [48] and [9] on distributed STBCs for asynchronous cooperative relay networks operating in a frequency flat environment. However, the assumption of frequency flat channels limits the applications of the aforesaid approaches to narrowband communication systems only, therefore, in this thesis, a one-bit feedback per subcarrier is proposed for distributed EO-STBC over frequency selective fading channels in wireless relay networks inspired by the work in [49]. On the basis of this foundation work the aims and objectives of this thesis are listed in the next section.

1.5.1 Aims and objectives

The aims of this thesis are:

- 1. To exploit the advantages of the closed-loop EO-STBC in [30] and [31], and closed-loop QO-STBC in a point-to-point MIMO system within asynchronous cooperative relay networks in an AF strategy, thereby providing a framework in which the advantages of theses codes are more likely to be practically realized.
- 2. To extend the framework in [28] and [50] to distributed CL EO-STBC and CL QO-STBC with outer coding for asynchronous cooperative relay networks and provide solutions to overcome the issues therein.

- 3. To provide a new framework for distributed CL EO-STBC with outer coding for more than two relay nodes under imperfect synchronization without a direct transmission (DT) connection between the source node and the destination node, and proposes a solution to overcome the drawback in [48] and [9], where the first used the near-optimum detector, which can achieve near-Alamouti simplicity, however cannot be extended to the case of more than two relay nodes, whereas, [9] relied mainly on the existence of DT link between source and destination, which is an impractical wireless mobile scenario.
- To apply OFDM type of transmission on modified quasi-orthogonal STBC (M-QO-STBC) in cooperative relay networks to mitigate timing errors for frequency flat channels.
- 5. To extend the work in [49] for more than two cooperative transmitting antennas using distributed OFDM type of transmission with closed loop EO-STBC in cooperative relay networks to mitigate timing errors over frequency selective channels utilizing onebit feedback.
- 6. To extend the EO-SFBC to cognitive relay networks and then apply codeword error rate and outage probability as a performance measurement for the proposed system.

At the end of the study the objectives are to have

 Demonstrated that the closed loop EO-STBC, closed loop QO-STBC and M-QO-STBC in cooperative relay systems, in various scenarios, can achieve full cooperating diversity.

- 2. Performed different techniques for eliminating the interference component at the destination occurring due to time misalignment between the cooperating relay nodes in a cooperative relaying communication system transmitting over both frequency flat and selective fading channels, taking into consideration as much as possible the simplicity at both relay and destination nodes.
- 3. Performed outage probability analysis for a cooperative cognitive network based on the overlay approach with perfect and imperfect spectrum acquisition.
- Published the research findings in international conferences and journals.

The outline of this thesis is summarized in the following section.

1.6 Outline of Thesis

This thesis is organized as follows: an introduction and discussion of conventional MIMO systems, cooperative relay networks, distributed STBCs and wireless communication channel impairments and cognitive radio are presented in Chapter 1.

In Chapter 2, a literature survey is provided together with the necessary theoretical background for distributed space-time coding schemes, details of orthogonal and quasi-orthogonal block codes, followed by the extended STBCs and a review of uncoded and coding gain. Then the PIC detection scheme and the OFDM type transmission are included. Finally, some background material and a review of previous work in synchronous and asynchronous cooperative relay networks are given. The core research is presented in Chapters 3, 4, 5 and 6. Chapter 3 focuses on the development of distributed CL EO-STBC for four relay nodes and distributed CL QO-STBC for two dual-antenna relay nodes with PIC detection in an AF asynchronous cooperative relay network with a DT link between the source node and the destination node.

Chapter 4, proposes a near-optimum detection scheme for the DF asynchronous cooperative relay networks utilizing distributed CL EO-STBC using two-bit feedback information without the existence of a DT link and with outer convolutive coding.

In Chapter 5, a closed-loop scheme for distributed EO-STBC using one-bit feedback per sub-carrier is proposed based on selection cyclic phase rotation for asynchronous wireless relay networks over frequency selective fading channels to achieve full data transmission rate with full cooperative diversity and reduce the amount of feedback information required from the destination node. Moreover, the timing error between the relay nodes is concatenated with the OFDM type transmission. Moreover, a M-QO-STBC scheme with full diversity and code gain distance (CGD) for use in asynchronous relay networks is presented employing an OFDM scheme with cyclic prefix (CP) at the source to mitigate the effect of random delays between relays and destination.

In Chapter 6, a closed-loop extended orthogonal space-frequency-block coding (CL-EO-SFBC) is proposed for use within cognitive wireless relay networks. The approach has the flexibility to dynamically adapt the code matrix to the number of available relays. Furthermore, the probability density function of the multi-path links is modeled in the time domain with an Erlang distribution function. The analytical expressions for the probability density function and cumulative density function of the end-to-end signal-to-noise ratio are obtained for one and two cognitive relay nodes and multi-path channel lengths of two and three. Moreover, expressions for outage probability are determined for a perfect and imperfect spectrum acquisition. Finally, the theoretical results are compared with simulations to confirm the validity of the analysis.

Lastly, in Chapter 7, the conclusions of the thesis are provided and also suggesting some future possible research directions.

LITERATURE SURVEY AND BACKGROUND ON COOPERATIVE WIRELESS NETWORKS

2.1 Introduction

Cooperative communication is a rapidly growing area of research, and it is likely to be a key enabling technology for efficient spectrum use in the future. Recently, researchers have been looking for methods to exploit spatial diversity provided by antennas of different users to combat fading and improve the reliability of transmission [51] [52]. This improvement is called cooperative diversity since it is achieved by having different users in the network cooperating in some way. One or more relay nodes are generally used in this paradigm to forward signals transmitted from the source node to the destination node. There are two main cooperative methods in a cooperative communication system: decode-and-forward (DF) (regenerative relaying protocol) and amplify-and-forward (AF) (transparent relaying protocol) [53]. In the DF relaying method, relay nodes decode the source information and then re-encode and re-transmit it to the destination, whereas in the AF method, relay nodes only amplify and retransmit received signals with noise, to the destination. For this reason the AF type of relaying schemes have the advantage of simple implementation and low complexity in practical scenarios as compared to the DF scheme. Cooperative communication via distributed space-time block codes (D-STBCs) has recently attracted much attention as an efficient technology that can provide considerable gains in fading wireless environments. The term distributed is used to indicate that conventional space-time block coding schemes can be applied to relay networks via distributed processing at the different relay nodes thus forming a virtual antenna array [53] to achieve cooperative diversity. In this chapter, a brief overview of cooperative wireless network concepts relevant to this thesis are presented. The chapter begins with an introduction to distributed space-time coding schemes and overviews two important codes, namely orthogonal and quasi-orthogonal codes. The discussion on the concept of maximum likelihood (ML) decoding and pairwise error probability (PEP) is considered next. Then the definition of uncoded and coded transmission is introduced, and then the principle of parallel interference cancelation (PIC) is presented. This is followed by a description of the principle of the orthogonal frequency division multiplexing (OFDM) type transmission and its advantages and disadvantages. Finally, a review of synchronous cooperative relay networks and cooperative operation is provided.

2.2 Overview of Distributed Space-Time Coding Schemes

In a general wireless relay network, different relays receive different noisy copies of the same information symbols. The relays process these received signals and forward them to the destination. The distributed processing at the different relay nodes thus forms a virtual antenna array [53]. Therefore, conventional space-time block coding schemes can be applied to relay networks to achieve cooperative diversity. In this section, the focus is the design of distributed space-time block codes based on an AF type relay protocol. There is much literature on AF type space-time block codes, i.e. [54], [55], [56] and [27]. Next, the fundamental designs proposed in [27] are considered in detail.

2.2.1 Distributed Transmission Technology

A wireless communication relay network is represented in Figure 2.1, it consists of one transmitter node S with one antenna (only one of the antennas in the diagram for the source is being used), one destination node D with one antenna and R relay nodes. Each relay node (R) has a half-duplex antenna for reception and transmission. It is assumed that the communication channels are quasi-static independent Rayleigh flat fading and the receiver has perfect channel information h_{sri} and h_{rid} , where h_{sri} and h_{rid} denote respectively the channels from the transmitter to the i^{th} relay and from the i^{th} relay to the receiver. It is assumed also that there is no direct link between the source and the destination as path loss or shadowing is expected to render it unusable. It is assumed that the transmitter sends the signal vector $\mathbf{s} = [s_1, ..., s_M]^T$, which is normalized so that $E[\mathbf{s}^H \mathbf{s}] = 1$ where M is the length of the time slot, $(.)^T$, $(.)^H$ and E[.] denote the transpose, Hermitian trans-



Figure 2.1. The block diagram of a two-hop wireless communication relay network over which distributed linear space time codes can be transmitted.

pose and the expectation of a random variable, respectively [28]. The transmission operation has two steps, in step one the transmitter sends signals $\sqrt{P_1Ms}$ to each relay where P_1 is the average power used at the transmitter for every transmission, whereas in step two, the i^{th} relay sends a signal vector to the receiver. The noise terms at the i^{th} relay within the vectors v_i and at the receiver w_i are independent complex Gaussian random variables with zero-mean and unit-variance. The received signal vector at the relays is given by

$$\mathbf{r}_i = \sqrt{P_1 M} h_{sr_i} \mathbf{s} + \mathbf{v}_i \qquad \text{for} \qquad i \in 0, \dots, R \tag{2.2.1}$$

The i^{th} relay transmits the signal vector \mathbf{t}_i which corresponds to the received signal vector \mathbf{r}_i multiplied by a scaled unitary matrix. The

transmitted signal vector from the i^{th} relay node can be generated from

$$\mathbf{t}_{i} = \sqrt{\frac{P_{2}}{P_{1}+1}} (A_{i}\mathbf{r}(i) + B_{i}\mathbf{r}^{*}(i))$$

= $\sqrt{\frac{P_{1}P_{2}T_{s}}{P_{1}+1}} (h_{sr_{i}}A_{i}\mathbf{s} + h_{sr_{i}}^{*}B_{k}\mathbf{s}^{*})$
+ $\sqrt{\frac{P_{2}}{P_{1}+1}} (A_{i}\mathbf{v}_{i} + B_{i}\mathbf{v}_{i}^{*})$ (2.2.2)

where A_i and B_i are MXM complex matrices, which depend on the distributed space time code, the 1 in the denominator scaling terms is the unity noise power, (.)* denotes the complex conjugate and P_2 is the average transmission power at every relay node. The received signal vector \mathbf{y} at the receiver, assuming perfect synchronization between all the relays and the destination node, is given by

$$\mathbf{y} = \sum_{i=1}^{N_r} h_{r_i d} \mathbf{t}_i + \mathbf{w}_i \tag{2.2.3}$$

where R is the number of relay nodes. The special cases that either $A_i = 0_M$, B_i is unitary or $B_i = 0_M$ and A_i is unitary are considered, where 0_M represents the MXM zero matrix. $A_i = 0_M$ means that the i^{th} relay column of the code matrix only contains the conjugates $s_1^*, ..., s_M^*$. and $B_i = 0_M$ means that the i^{th} relay column contains only the information symbols $s_1, ..., s_M$. Thus the following variables are defined as [28]

$$\hat{A}_{i} = A_{i}, \quad \hat{h}_{sr_{i}} = h_{sr_{i}}, \quad \hat{\mathbf{v}}_{i} = \mathbf{v}_{i}, \quad \mathbf{s}^{(i)} = \mathbf{s}, \qquad if \quad B_{i} = 0_{M}$$
$$\hat{A}_{i} = B_{i}, \quad \hat{h}_{sr_{i}} = h_{sr_{i}}^{*}, \quad \hat{\mathbf{v}}_{i} = \mathbf{v}_{i}^{*}, \quad \mathbf{s}^{(i)} = \mathbf{s}^{*}, \qquad if \quad A_{i} = 0_{M}$$

From (2.2.2)

$$\mathbf{t}_i = \sqrt{\frac{P_1 P_2 M}{P_1 + 1}} \hat{h}_{sr_i} \hat{A}_i \mathbf{s}^{(i)} + \sqrt{\frac{P_2}{P_1 + 1}} \hat{A}_i \hat{\mathbf{v}}_i$$

The signal vector at the receiver can be calculated from equations (2.2.1) and (2.2.3) to be

$$\mathbf{y} = \sqrt{\frac{P_1 P_2 M}{P_1 + 1}} \mathbf{S} \mathbf{h} + \mathbf{w}'_d \tag{2.2.4}$$

where

$$\mathbf{S} = [\hat{A}_1 \mathbf{s}^{(1)} \dots \hat{A}_R \mathbf{s}^{(R)}], \quad \mathbf{h} = [\hat{h}_{sr_1} h_{r_1 d} \dots \hat{h}_{sr_n} h_{r_n d}]^T \quad (2.2.5)$$

and

$$\mathbf{w}_{d}' = \sqrt{\frac{P_{2}}{P_{1}+1}} \sum_{i=1}^{R} h_{sr_{i}} \hat{A}_{i} \hat{\mathbf{v}}_{i} + \mathbf{w}$$
(2.2.6)

Therefore, without decoding, the relays generate a space-time codeword **S** distributively at the receiver. The vector **h** is the equivalent channel and \mathbf{w}'_d is the equivalent noise vector. The optimum power allocation [27] is when the transmitter uses half the total power and the relays share the other half. If the total power is P and the number of relays is R, the average powers used at the source and relays are

$$P_1 = \frac{P}{2}$$
 and $P_2 = \frac{P}{2R}$

If the channel vector h is known at the receiver, the maximum-likelihood (ML) decoding is

$$\hat{\mathbf{s}} = \arg\min_{\mathbf{s}} \left\| \mathbf{y} - \sqrt{\frac{P_1 P_2 M}{P_1 + 1}} \mathbf{Sh} \right\|^2$$
(2.2.7)

where $\|.\|$ denotes the Euclidean norm, and *arg min* represents finding the smallest Euclidean norm from all possible **S** formed as in (2.2.5) from the source signal vectors **s** defined by the chosen source constellation. The distributed space-time codes (D-STBCs) implemented at the cooperative relay nodes play an important role in wireless communication systems and are very attractive for next generation wireless communication systems. D-STBCs refer to coding across space by using multiple cooperative relaying antennas and receiver antennas and across time by using multiple symbol transmission periods. Their aim is to exploit the spatial diversity available to the system with linear processing at the receiver. There are several types of STBCs which can be distributed over cooperating relay nodes, these include Alamouti STBC (A-STBC) [29], orthogonal STBC (O-STBC) [57], quasi orthogonal STBC (QO-STBC) [58] and extended orthogonal STBC (EO-STBC) [31].

2.2.2 Alamouti Space Time Block Coding

The Alamouti code is probably the most well-known STBC and the simplest transmit diversity scheme [29]. It is the first STBC scheme that can provide full diversity order of $2N_R$ and full data transmission D_R rate¹ for complex constellations by transmitting signals across two transmit antennas and N_R receiver antennas with different symbol transmission periods. The block diagram of an A-STBC operation with a single receiver antenna is presented in Figure 2.2. STBCs are represented by a code matrix, which defines what is to be broadcasted

¹Data transmission rate D_R is defined as the ratio between the number of transmitted symbols that can be sent in one codeword and the number of symbol transmission periods used in transmitting the codeword.



Figure 2.2. Basic structure of Alamouti's STBC for two transmit antennas and one receive antenna, depicting transmission over two symbol periods.

from the transmit antennas during the transmission of a block. The code matrix is dimension of $N_T \times T$, where N_T represents the number of transmitter antenna and T the number of symbol transmission periods.

As shown in Figure 2.2 for a two transmit antennas and one receive antenna scheme with a frequency flat Rayleigh fading channel, the code matrix for Alamouti's code is given by

$$\mathbf{S} = \begin{bmatrix} s_1 & s_2 \\ \\ -s_2^* & s_1^* \end{bmatrix}$$
(2.2.8)

where (.)* denotes complex conjugate. It is readily apparent that the data transmission rate D_R is unity². The columns of the matrix in (2.2.8) represent the number of transmitter antennas and the rows represent the number of symbol transmission periods. The orthogonality of the columns of the matrix **S** can be easily verified by calculating the

²Because two symbols are transmitted during two symbol transmission periods.

inner product of the columns \mathbf{s}_1 and \mathbf{s}_2 of (2.2.8) and the inner product is given by $\langle \mathbf{s}_1, \mathbf{s}_2 \rangle = s_1 s_2^* - s_2^* s_1 = 0$ (2.2.9)

From (2.2.9) the inner product between columns of matrix is zero, as this is the basis for it to be orthogonal. During the first symbol transmission period, the transmitter sends s_1 from the first antenna and s_2 from the second antenna, during the second symbol transmission period, it transmits $-s_2^*$ and s_1^* from the first and second antenna, respectively. Exploiting the basic feature of code matrix **S**, the orthogonal columns of code matrix **S** as in (2.2.9) provide a linear decoding scheme and give rise to a simplified ML decoding scheme [11]. Therefore, the symbols in the same symbol transmission period must be synchronized and transmitted together so that they can be detected independently at the receiver. As shown in Figure 2.2 in the case of one receive antenna, the received signals at this antenna, at two symbol transmission periods can be represented with a 2 × 1 vector **y** as follows

$$\mathbf{y} = \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}$$
(2.2.10)

where y_1 and y_2 are the received signals at two symbol transmission periods, h_1 and h_2 are the fading channel coefficients from transmitter antenna one and two which are assumed to remain constant across these two symbol transmission periods, and n_1 , n_2 are independent, zero-mean circularly symmetric, additive Gaussian noise terms across the channel at two symbol transmission periods, respectively. Therefore (2.2.10) can be written in matrix form as

$$\mathbf{y} = \mathbf{S}\mathbf{h} + \mathbf{n} \tag{2.2.11}$$

where **y** represents a 2×1 column vector, **h** is a 2×1 column vector, **S** is a 2×2 vector, and **n** represent a 2×1 column vector. Alternately, without loss of generality, by conjugating the second row in (2.2.10), then the received signals can be equivalently written in the following form

$$\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2^* \end{bmatrix}$$
(2.2.12)

also (2.2.12) can be written in matrix form as

$$\tilde{\mathbf{y}} = \mathbf{H}\mathbf{s} + \tilde{\mathbf{n}} \tag{2.2.13}$$

where $\tilde{\mathbf{y}}$ represents a 2 × 1 column vector, **H** is a 2 × 2 matrix of the transmission path vector, **s** is a 2 × 1 column vector, and $\tilde{\mathbf{n}}$ represents a 2 × 1 column vector. Assuming perfect channel knowledge at the receiver, Alamouti's combiner can be performed by multiplying both sides of (2.2.13) by \mathbf{H}^{H} , where (.)^{*H*} denotes Hermitian conjugate, it is a scaled unitary matrix i.e., $\mathbf{H}^{H}\mathbf{H} = \lambda \mathbf{I}_{2}$, where \mathbf{I}_{2} is the 2 × 2 identity matrix and λ is the gain of the channel with $\lambda = |h_{1}|^{2} + |h_{2}|^{2}$, which is due to the fact that the transmitted symbol block has an orthogonal structure as shown in (2.2.9). Therefore, the combiner combines the received signals as follows

$$\tilde{\mathbf{s}} = \mathbf{H}^H \tilde{\mathbf{y}} \tag{2.2.14}$$

Thus

$$\tilde{s}_{1} = \lambda s_{1} + h_{1}^{*} \tilde{n}_{1} + h_{2} \tilde{n}_{2}^{*}$$
$$\tilde{s}_{2} = \lambda s_{2} + h_{2}^{*} \tilde{n}_{1} + h_{1} \tilde{n}_{2}^{*}$$
(2.2.15)

From (2.2.15), it can be noted that the decision for \tilde{s}_1 depends on s_1 and the decision for \tilde{s}_2 depends on s_2 . As shown in Figure 2.2, these combined signals are then sent to the ML decision rule used at the receiver to choose which symbol was actually transmitted by applying least squares (LS) detection as follows

$$\hat{s}_k = \arg \min_{s_k \in S} |\tilde{s}_k - \lambda s_k|^2 \quad \text{for} \quad k \in 1,2 \quad (2.2.16)$$

where S is the alphabet containing M symbols of phase shift keying (PSK) and |.| is a magnitude operator. It can be observed that the ML detection is a very simple decoding scheme which includes decoupling of signals transmitted from different transmitter antennas via linear processing at the receiver side. Alamouti further extended this scheme to the case of two transmit antennas and N_R receive antennas and showed that the new scheme provided a maximal diversity order of $2N_R$ [11]. The Alamouti scheme is a very popular scheme in practical application in several wireless communication systems such as WiFi, WiMax, 3G LTE and 4G [59] because it provides two important properties, simple decoding combined with maximal spatial diversity gain advantage.

2.2.3 Orthogonal Space Time Block Coding

Alamouti's code in (2.2.8) is only designed for two transmit antennas to achieve both full diversity order and full data transmission rate (unity) for complex constellations with simple decoding algorithm at the receiver. In [57] a general design of O-STBC for more than two transmit antennas is proposed, which may achieve full diversity order that is identical to the number of transmit antennas however with a data transmission rate less than unity. All O-STBC's share the common unitary-type property as follows

$$\mathbf{S}^{H}\mathbf{S} = (\sum_{k=1}^{n_{s}} |s_{k}|^{2}) \mathbf{I}_{N_{T}}$$
(2.2.17)

where n_s is the number of symbols to be transmitted within a codeword. Some other examples of O-STBC's are given below [60], [57] and [15]

$$\mathbf{S} = \begin{bmatrix} s_1 & s_2 & s_3 \\ 0 & s_1^* & -s_2^* \\ -s_1^* & 0 & -s_2^* \\ s_2^* & -s_3^* & 0 \end{bmatrix}$$
(2.2.18)

and

$$\mathbf{S} = \begin{bmatrix} s_1 & s_2 & s_3 & 0 \\ -s_2^* & s_1^* & 0 & -s_3 \\ -s_3^* & 0 & s_1^* & s_2 \\ 0 & s_3^* & -s_2^* & s_1 \end{bmatrix}$$
(2.2.19)

where the rows of the code matrix **S** represent the number of symbol transmission periods and the columns of the code matrix **S** denote the number of transmitter antennas. Therefore, the code matrix in (2.2.18) and (2.2.19) correspond to transmission of three symbols s_1, s_2, s_3 over four time symbol transmission periods, achieving full diversity order of four and data transmission rate equal to $D_R = 3/4$ for three and four transmitter antennas, respectively. The zeros in both transmission code matrices **S** represent that there is no transmitted signal from that particular antenna during that symbol transmission period. Their performance is presented and simulated in Chapter 3 for asynchronous cooperative relay networks. Furthermore, there are O-STBCs that can achieve a data transmission rate equal to $D_R = 1/2$ for any given number of transmitter antennas. For example the code matrix in (2.2.20) which denotes transmission of four symbols over eight time symbol periods from three transmit antennas, achieving full diversity order of three and 1/2 data transmission rate [57];

$$\mathbf{S} = \begin{bmatrix} s_1 & s_2 & s_3 \\ -s_3 & s_1 & -s_4 \\ -s_3 & s_4 & s_1 \\ -s_4 & -s_3 & s_2 \\ s_1^* & s_2^* & s_3^* \\ -s_2^* & s_1^* & -s_4^* \\ -s_3^* & s_4^* & s_1^* \\ -s_4^* & s_3^* & s_2^* \end{bmatrix}$$
(2.2.20)

In summary, O-STBCs can provide full diversity order and have a simple decoding algorithm. However, they suffer from low data transmission rate when used with more than two transmit antennas and complex constellation, which result in bandwidth expansion. It is desirable to construct STBCs from complex orthogonal design that have higher data transmission rate when there are more than two transmit antennas. This class of codes is called QO-STBCs [58], [15], [61] and [62], which are considered in the following section.

2.2.4 Quasi Orthogonal Space Time Block Coding

QO-STBCs with full data transmission rate for four transmit antennas have been proposed in [58] and [61] to overcome the shortcoming of O-STBCs as mentioned in the previous section. Although the code matrices presented in these publication are different, it has been shown that their properties and performance are identical. Therefore, only the code matrix proposed in [58] is considered in Chapter 3 for cooperative relay networks under imperfect synchronization, which provides more insight into the behavior of QO-STBC. In this scheme, the A-STBC defined in (2.2.8) is extended to construct the code matrix by using two Alamouti codes \mathbf{S}_k (hence with four data symbols), where $k \in 1, 3$,

$$\mathbf{S}_{k} = \begin{bmatrix} s_{k} & s_{k+1} \\ -s_{k+1}^{*} & s_{k}^{*} \end{bmatrix}$$
(2.2.21)

Therefore, the QO-STBC matrix of [58] is represented as

$$\mathbf{S} = \begin{bmatrix} \mathbf{S}_{1} & \mathbf{S}_{3} \\ -\mathbf{S}_{3}^{*} & \mathbf{S}_{1}^{*} \end{bmatrix} = \begin{bmatrix} s_{1} & s_{2} & s_{3} & s_{4} \\ -s_{2}^{*} & s_{1}^{*} & -s_{4}^{*} & -s_{3}^{*} \\ -s_{2}^{*} & s_{1}^{*} & -s_{4}^{*} & -s_{3}^{*} \\ -s_{3}^{*} & -s_{4}^{*} & s_{1}^{*} & s_{2}^{*} \\ s_{4} & -s_{3} & -s_{2} & s_{1} \end{bmatrix}$$
(2.2.22)

It can be clearly observed that from the codeword in (2.2.22), the data transmission rate D_R is unity³, nevertheless the diversity order is reduced by half $(2N_R)^4$ due to coupling between the symbols in the codeword (2.2.22) and the relaxation of the orthogonality of QO-STBC increases the ML decoding complexity and in fact, the ML decoding of QO-STBC is in general complex symbol pair-wise decoding⁵. To show

 $^{^3\}mathrm{Because}$ four data symbols are transmitted over four symbol transmission periods.

⁴It has been proven in [57] that the maximum diversity order of $4N_R$ for unity data transmission rate is impossible in the case of QO-STBC.

⁵The decoder of QO-STBC works with pairs of transmitted symbols instead of single symbols.

this, the codeword in (2.2.22) is considered to be transmitted through four transmit and one receive antennas in frequency flat quasi-static fading environment with Rayleigh distributions. After taking the complex conjugate of the symbols in the second and third rows of the matrix (2.2.22), the received signal can be expressed as follows

$$\begin{bmatrix} y_1 \\ y_2^* \\ y_3^* \\ y_4 \end{bmatrix} = \underbrace{\begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ -h_2^* & h_1^* & -h_4^* & -h_3^* \\ -h_3^* & -h_4^* & h_1^* & h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix}}_{\mathbf{H}} \begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2^* \\ n_3^* \\ n_4 \end{bmatrix} \quad (2.2.23)$$

Therefore, (2.2.23) can be expressed in vector form as follows

$$\tilde{\mathbf{y}} = \mathbf{H}\mathbf{s} + \tilde{\mathbf{n}} \tag{2.2.24}$$

where $\tilde{\mathbf{y}}$ represents a 4 × 1 column vector of the received signal, **H** is a 4 × 4 matrix of the transmission path vector, and $\tilde{\mathbf{n}}$ represents a 4 × 1 column vector containing the zero-mean circularly symmetric complex valued Gaussian noise components. After applying the matrix \mathbf{H}^{H} to perform matched filtering, i.e. perfect CSI is assumed to be available at the receiver, then the estimates of the transmitted symbols can be expressed as in (2.2.14). Therefore (2.2.14) can be represented in matrix form as follows

$$\begin{bmatrix} \tilde{s}_1\\ \tilde{s}_2\\ \tilde{s}_3\\ \tilde{s}_4 \end{bmatrix} = \underbrace{\begin{bmatrix} \gamma & 0 & 0 & \alpha\\ 0 & \gamma & -\alpha & 0\\ 0 & -\alpha & \gamma & 0\\ \alpha & 0 & 0 & \gamma \end{bmatrix}}_{\Delta} \begin{bmatrix} s_1\\ s_2\\ s_3\\ s_4 \end{bmatrix} + \mathbf{H}^H \begin{bmatrix} \tilde{n}_1\\ \tilde{n}_2^*\\ \tilde{n}_3^*\\ \tilde{n}_4 \end{bmatrix}$$
(2.2.25)

where $\Delta = \mathbf{H}^H \mathbf{H}$ is a 4 × 4 matrix with entries $\gamma = \sum_{k=1}^4 |h_k|^2$ and $\alpha = \Re(h_1^*h_4 - h_2^*h_3)$, where $\Re\{.\}$ denotes the real part operator. As mentioned in the previous sections, for O-STBC, all the off-diagonal terms of Δ will be zeros as in Alamouti's scheme [29]. However, for the QO-STBC it can be seen that due to the term α (some non-zeros off diagonal terms appear), there is a form of coupling between estimated symbols reducing the diversity gain of the code, which increases the complexity of ML decoding to a pair-wise operation. Several interesting methods have been proposed to increase the performance of the open-loop QO-STBC by minimizing or removing this coupling factor. One method is proposed in [32] to achieve full diversity order and full data transmission rate for four transmit antennas. A feedback method is used to orthogonalize the QO-STBC and is achieved by rotating the transmitted symbols from the third and fourth antenna with particular angles ϕ and θ while the other two antennas are kept unchanged. Therefore, full diversity order and full data transmission rate are achieved by eliminating the off-diagonal elements $\alpha = 0$, however at the expense of feedback overhead as follows

$$\alpha = \Re\{h_1^* h_4 e^{j\theta} - h_2^* h_3 e^{j\phi}\}$$
$$= |\kappa| \cos(\theta + \lambda \kappa) - |\kappa| \cos(\phi + \lambda \kappa) \qquad (2.2.26)$$

where $\kappa = h_1^* h_4$ and $\dot{\kappa} = h_2^* h_3$. Here, |.| and \angle denote respectively, the absolute value and the angle operator. In order to orthogonalize the QO-STBC, it is sufficient to set the phase angle value θ to

$$\theta = \cos^{-1}\left(\frac{|\hat{\kappa}|}{|\kappa|}\cos(\phi + \angle \hat{\kappa})\right) - \angle \kappa \tag{2.2.27}$$

provided that ϕ is in the range $\phi \in [0, 2\pi]$ if $|\dot{\kappa}| < \kappa$, or otherwise, $\phi \in [\pi - \mu - \angle \hat{\kappa}, \mu - \hat{\kappa}] \cup [-\mu - \hat{\kappa}, \pi + \mu - \hat{\kappa}],$ where μ is defined by $\mu = \arccos(\kappa/\kappa)$ [32]. However, in a practical application this may not be possible due to the very limited feedback bandwidth. It was further shown in the same work that identical performance of the system can also be obtained by rotating the third and fourth antennas by a common phaser to keep the feedback from the receiver to transmit antenna as small as possible. It is shown in [63], [64] and [65] that QO-STBC achieves full data transmission rate by using constellation rotation and partial feedback, however all these methods either have increased decoding complexity or reduced diversity order. In the context of this thesis only the single phase approach in [32] is presented in Chapter 3 for asynchronous cooperative relay networks, as it requires minimum feedback information while minimizing the decoding complexity of ML decoding, thereby making it symbol-wise decoding with full data transmission rate and full cooperative diversity order.

2.2.5 Extended Alamouti Space Time Block Coding

As was discussed in the last section due to data transmission rate limitations in O-STBCs, Alamouti's codeword matrix in (2.2.8) can also be used to build a new class of codes named EO-STBC, which was introduced in [31]. Slightly different from the QO-STBC, this code achieves full data transmission rate of unity⁶ with symbol-wise decoding by transmitting two symbols over two symbol transmission periods through four transmitting antennas. Therefore, the codeword matrix

 $^{^{6}\}mathrm{Because}$ two data symbols are transmitted over two symbol transmission periods.
of EO-STBC is represented as

$$\mathbf{S} = \begin{bmatrix} s_1 & s_1 & s_2 & s_2 \\ -s_2^* & -s_2^* & s_1^* & s_1^* \end{bmatrix}$$
(2.2.28)

Assuming that the codeword matrix in (2.2.28) is transmitted over a four transmit and one antenna channel with each path experiencing independent flat fading with a Rayleigh distribution, the received signal at the receiver side over two symbol transmission periods after taking the complex conjugates of the symbols in the second symbol transmission period can be modelled as

$$\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \begin{bmatrix} h_1 + h_2 & h_3 + h_4 \\ h_3^* + h_4^* & -h_1^* - h_2^* \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2^* \end{bmatrix}$$
(2.2.29)

Therefore, (2.2.29) can be rewritten in vector form as follows

$$\tilde{\mathbf{y}} = \mathbf{H}\mathbf{s} + \tilde{\mathbf{n}} \tag{2.2.30}$$

where $\tilde{\mathbf{y}}$ denotes a 2 × 1 column vector of the received signal, **H** is a 2 × 2 matrix of the channel gain vector representing the paths between the transmit antennas and receive antenna, and $\tilde{\mathbf{n}}$ represents a 2 × 1 column vector containing the additive Gaussian noise term at each receive antenna. Applying the matrix \mathbf{H}^{H} as in (2.2.14) to perform matched filtering at the receiver, the Grammian matrix can be obtained as follows

$$\Delta = \mathbf{H}^{H} \mathbf{H} = \begin{bmatrix} \lambda_{c} + \lambda_{f} & 0\\ 0 & \lambda_{c} + \lambda_{f} \end{bmatrix}$$
(2.2.31)

where $\lambda_c = \sum_{k=1}^4 |h_k|^2$ is the conventional channel gain for four transmit antennas and $\lambda_f = 2\Re(h_1h_2^* + h_3h_4^*)$ can be interpreted as the channel dependent interference parameter. From (2.2.31) it can be noted that the Grammian matrix Δ of the EO-STBC is orthogonal (unlike that of QO-STBCs), which indicates that the codeword in (2.2.28) can be decoded using a simpler receiver as follows

$$\begin{bmatrix} \tilde{s_1} \\ \tilde{s_2} \end{bmatrix} = \Delta \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \mathbf{H}^H \begin{bmatrix} \tilde{n}_1 \\ \tilde{n}_2^* \end{bmatrix}$$
(2.2.32)

Then, the ML decision is used at the receiver to estimate which symbol was actually transmitted by applying LS detection as in (2.2.16). However, although the decoding complexity is low, λ_f in (2.2.31) may be negative, which leads to some diversity loss [30] and [31]. In order to overcome this issue and maximize the diversity order achieved by the codeword in (2.2.28), several feedback methods are utilized to ensure the channel dependent interference λ_f in (2.2.31) is positive during the whole transmission to achieve full data transmission rate and full diversity order with the array gain. In [30] and [31] a closed-loop EO-STBC scheme for four transmit and one receive antenna is proposed, where phase rotation is applied to rotate the phases of the symbols for certain transmit antennas based on CSI feedback from the receive antenna. The analysis of these two closed-loop methods is presented in Appendix A. In the context of this thesis both proposed schemes are adopted for application in an asynchronous cooperative wireless relay network with relay nodes deploying the distributed EO-STBC scheme with relay selection, with and without outer coding (Chapters 3 and 4), and a new one-bit feedback scheme is deployed in Chapter 5.

2.2.6 Maximum Likelihood and Pairwise Error Probability

As mentioned in the previous sections, all STBC's are designed for quasi-static channels, where the channels are constant over a block of T symbols periods and the channels are unknown at the transmitter, while the receiver has knowledge of the channel matrix **H**. Under ML detection it can be shown using similar techniques as in (2.2.16), giving received matrix **Y**, the ML estimate of the matrix $\hat{\mathbf{S}}$ can be represented as follows [34]

$$\hat{\mathbf{S}} = \arg \min_{\mathbf{S} \in \chi^{N_T \times T}} ||\mathbf{Y} - \mathbf{HS}||_F^2 = \arg \min_{\mathbf{S} \in \chi^{N_T \times T}} \sum_{k=1}^T ||\mathbf{y}_k - \mathbf{Hs}_k||^2 \quad (2.2.33)$$

where $||\mathbf{A}||_F$ is the Frobenius norm⁷ of matrix \mathbf{A} and minimization is taken over all possible space time input matrix $\chi^{M_T \times T}$. The pairwise error probability (PEP) for mistaking a transmit matrix \mathbf{S} for another matrix $\hat{\mathbf{S}}$, represented as $P_e(\hat{\mathbf{S}} \to \mathbf{S})$, depends only on the distance between the two matrices after transmission through channel matrix \mathbf{H} and noise power of \mathbf{N} equal to σ^2 can be represented as follows

$$P_e(\hat{\mathbf{S}} \to \mathbf{S}) = Q\left(\sqrt{\frac{||\mathbf{HS_s}||_F^2}{2\sigma^2}}\right)$$
(2.2.34)

where Q(.) represents the Gaussian Q function and $\mathbf{S}_{\mathbf{s}} = \mathbf{S} - \hat{\mathbf{S}}$ represents the difference between these two matrices. By applying the Chernoff bound to (2.2.34), an upper bound of (2.2.34) becomes

$$P_e(\hat{\mathbf{S}} \to \mathbf{S}) \le \exp\left[-\frac{||\mathbf{H}\mathbf{S}_s||_F^2}{4\sigma^2}\right]$$
 (2.2.35)

⁷The Frobenius norm of a matrix **A** is defined to be $||\mathbf{A}|| = \sqrt{\sum_{ij} |\mathbf{a}_{ij}|^2}$.

Let \mathbf{h}_k denote the kth row of $\mathbf{H}, k \in 1, ..., N_R$. Then

$$||\mathbf{H}\mathbf{S}_{\mathbf{s}}||_{F}^{2} = \sum_{k=1}^{N_{R}} \mathbf{h}_{k} \mathbf{S}_{\mathbf{s}} \mathbf{S}_{\mathbf{s}}^{H} \mathbf{h}_{k}^{H}$$
(2.2.36)

Substituting (2.2.36) into (2.2.35) and taking statistical expectation relative to all possible channel realizations yields [34]

$$P_e(\mathbf{S} \to \hat{\mathbf{S}}) \le \left(det \left[\mathbf{I}_{N_T N_R} + \frac{\mathbf{E}[\mathbf{S}_{\mathbf{s}}^H \mathbf{H}^H \mathbf{H} \mathbf{S}_{\mathbf{s}}]}{4\sigma^2} \right] \right)^{-1}$$
(2.2.37)

Assuming that the channel matrix \mathbf{H} has elements which are uncorrelated Gaussian random variables with zero-mean and unit-variance, then (2.2.37) becomes

$$P_e(\mathbf{S} \to \hat{\mathbf{S}}) \le \left(\frac{1}{\det[\mathbf{I}_{N_T} + 0.25\rho\mathbf{B}]}\right)^{N_R}$$
(2.2.38)

where \mathbf{I}_{N_T} is the $N_T \times N_T$ identity matrix and $\mathbf{B} = (1/P) \mathbf{S}_{\mathbf{s}} \mathbf{S}_{\mathbf{s}}^H$, where P denote the signal power. This can be simplified to

$$P_e(\mathbf{S} \to \hat{\mathbf{S}}) \le \prod_{k=1}^{R_{\mathbf{B}}} \left(\frac{1}{1 + 0.25\rho\lambda_k(\mathbf{B})}\right)^{N_R}$$
(2.2.39)

where $\rho = P/\sigma^2$ is the SNR per input symbol, $\lambda_k(\mathbf{B})$ is the *kth* nonzero eigenvalue of $\mathbf{B}, k \in 1, ..., R_{\mathbf{B}}$, where $R_{\mathbf{B}}$ is the rank of \mathbf{B} . In the case of high SNR i.e., for $\rho \gg 1$, this can be simplified to

$$P_e(\mathbf{S} \to \hat{\mathbf{S}}) \le \left(\prod_{k=1}^{R_{\mathbf{B}}} \lambda_k(\mathbf{B})\right)^{-N_R} \left(\frac{1}{4}\rho\right)^{-R_{\mathbf{B}}N_R}$$
(2.2.40)

The result of PEP in (2.2.40) indicates that the probability of error decreases as ρ^{-D_g} for $D_g = R_{\mathbf{B}}N_R$, where $R_{\mathbf{B}}N_R$ is the diversity gain of the STBC [34]. Therefore, the PEP in (2.2.40) gives rise to the main criteria for design of STBC, described in Section 2.3.

2.3 Uncoded Versus Coded Transmission

2.3.1 Coding Gain

Coding gain is the measure in the difference between the SNR levels between the uncoded system and coded system required to reach the same bit error rate (BER) levels. Therefore, when SNR is in dB the coding gain is defined as [66]

$$C_g = (SNR)_{uncoded} - (SNR)_{coded} \tag{2.3.1}$$

This also can reduce error rate to improve system performance, but, compared with diversity gain, the nature of coding gain is different. Diversity gain attests itself by rising the magnitude of the slope of the



Figure 2.3. The difference in the effects of coding gain and diversity gain on bit error rate [7].

BER curve, whereas coding gain generally just shifts the error rate curve to the left [10] as shown in Figure 2.3.

2.3.2 Convolution Coding

Convolutional codes are used extensively in numerous applications in order to achieve reliable data transfer, i.e. third generation (3G) cellular communication system [67] and [68]. A convolution code generates coded symbols by passing the information bits through a linear finitestate shift register as shown in Figure 2.4. The shift register consists of K stages with I_{Bits} bits per stage. There are O_{Bits} binary addition operations with input taken from all K stages: these operators produce a codeword of length O_{Bits} for each I_{Bits} bit input sequence. Moreover, the rate of the code C_R is I_{Bits}/O_{Bits} , because the binary input data is shifted into each stage of the shift register I_{Bits} bits at a time, and each of these shifts produces a coded sequence of length O_{Bits} . The number



Figure 2.4. Basic structure of a simple binary linear convolutional encoder.

of shift register stages K is called the constraint length. In Chapters 3, 4 and 5, a half rate⁸ ($O_{Bits} = 2, I_{Bits} = 1, K = 3$) convolution coding will be used to improve the BER performance. A well known scheme

⁸For every binary digit that enters the encoder, two code digits are output, hence code rate $C_R = 1/2$.

can be employed to decode the convolution coding, which is the Viterbi algorithm, full details of which can be found in [7], [67] and [68].

2.4 The Parallel Interference Cancellation Detection

The PIC detector is a popular method for interference cancellation and was presented in [69] and a block diagram of a multistage PIC detector is shown in Figure 2.5. The PIC detector is based on a technique that employs multiple iterations in detecting the desired information bits and cancelling the ISI at the receiver side. As shown in Figure 2.5, the PIC detector at the receiver uses a conventional bank of matched filter at the front end followed by two stages of cancellation. In the first and



Figure 2.5. Basic structure of parallel interference cancelation (PIC) scheme that uses a two cancellation stage.

the second stages, the reconstruction, subtraction and re-estimation operations are repeated using the outputs of the previous stage as the information inputs. In this way, a new set of better information bit estimates will be generated at the output of the next stage. Multiple cancellation stages maybe used to yield improved estimates of the received signals at the destination node for further cancellation. In general, the PIC algorithm can be represented as follows

- Initialization
- Set the iteration number q = 0
- Remove ISI from the $\mathbf{r}^{(0)}(i)$
- Set the iteration number q = 1, 2, ..., n
- Subtract more ISI from the received signals r^(q)(i) using the previous estimated signals
- Repeat the process from point 4 until q = n

One problem with the multistage PIC detector is that it cannot guarantee a performance improvement with more stages; when a wrong estimation of the received signal is applied, degradation is introduced. Furthermore, increasing the number of stages of the multistage PIC detector also makes the system computationally more expensive as a greater number of operations have to be preformed. In Chapter 3, the PIC detection scheme is adapted for application in an asynchronous cooperative relay network with the relay nodes utilizing the distributed closed-loop STBC.

2.5 Orthogonal Frequency Division Multiplexing

Recently, wireless communications has evolved towards broadband systems to achieve high data transmission rate and better transmission quality. OFDM is an emerging technology for high data transmission rates, with increased robustness to shadowing and multipath fading, through insertion of the guard period G_p and using digital signal processing. OFDM modulation for wireless communication networks is designed to more fully utilize existing bandwidth, and can be very effective against the frequency selective fading and narrowband interference [33] [70]. Due to these characteristics of OFDM, it is used widely in many applications such as digital audio broadcasting and the IEEE 802.11a/g WLAN standard, and is the preferred candidate for next generation mobile systems, such as 4G LTE. A review of OFDM is given in the following sections and it is used in Chapter 5 within a distributed closed-loop EO-STBC-OFDM for asynchronous cooperative relay networks and in Chapter 6 within a distributed closed-loop EO space-frequency block code (EO-SFBC-OFDM) for cognitive cooperative relay networks.

2.5.1 OFDM principle

A typical baseband OFDM transmission system is shown in Figure 2.6., which can be divided into three main parts. The first part is the transmitter, the second the channel and finally the receiver. To generate OFDM successfully, the relationship between all the subcarriers must be carefully controlled to maintain the orthogonality of the subcarriers [71]. Orthogonality is achieved by making the peak of each subcarrier signal coincide with the nulls of other subcarrier signals. For this reason, OFDM is generated by firstly choosing the frequency spectrum required, based on the input data, and the modulation scheme used can, for example, be binary phase shift keying (BPSK), quadrature phase shift keying (QPSK) and quadrature amplitude modulation (QAM). Then the frequency spectrum is converted into the time domain signal by applying an inverse fast Fourier transform (IFFT). The IFFT is useful for OFDM because it generates samples of a subcarrier



Figure 2.6. Basic diagram of a baseband OFDM transmitter and receiver, exploiting the FFT and serial to parallel and parallel to serial converters.

with frequency components satisfying the orthogonality condition, assuming perfect synchronization. In addition to that the FFT algorithm provides an efficient way to implement the DFT and the IDFT. The implementation of OFDM is very straightforward by using the FFT algorithm and it reduces the number of complex multiplications from of the order of N^2 to $N/2\log_2 N$ for an N-point DFT or IDFT [33]. The IFFT calculated from the subcarrier S(k) can be represented as follows

$$s(m) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S(k) e^{j2\pi km/N} \qquad \text{for} \qquad m \in 0, \dots, N-1 \quad (2.5.1)$$

where N denotes the duration of one OFDM symbol, k is the frequency index for the IFFT and m is the index for the samples of s(m). That means, the IFFT converts the discrete frequency domain signal into a discrete time domain signal which maintains the orthogonality. On the other hand, in the receiver side the FFT converts the time domain signal into the frequency domain to recover the information that was originally sent. However, multipath interference and timing errors are an important phenomena in wireless communication systems. Different versions of the transmitted signals arrive at the receiver side by different paths showing different time delays. Such delayed signals are the result of refections and refractions from objects, which are surrounding the receiver. This makes the correct synchronization to all signals impossible. To avoid this phenomenon of intersymbol interference (ISI) completely, a cyclic prefix (CP) is introduced, with the constraint that the CP duration at least matches the maximum delay spread of the channel. The time domain duration of the OFDM symbol is extended by the CP, which adds a copy of the last part of the OFDM symbol to the front to create a guard period as shown in Figure 2.7. To avoid destroying the orthogonality the length of the CP should be longer than the delay spread. At the receiver side this extension of the guard period is removed.

As long as the length of the guard period is greater than the channel



Figure 2.7. Illustration of cyclic prefix (CP) concept to the original OFDM signal.

delay spread, then all the time delays of the reflections are removed and the orthogonality still exists. The next step is to convert the received signal from the time domain to the frequency domain by applying the FFT to recover the transmitted signals.

2.5.2 OFDM Advantages and Disadvantages

As discussed earlier, OFDM is commonly implemented in many emerging wireless communication systems, because it provides several advantages over a traditional single carrier modulation approach to wireless communication channels as follows

- OFDM has very high frequency spectrum efficiency.
- By dividing the channel into narrowband flat fading sub channels, OFDM is more resistant to frequency selective fading, which greatly simplifies the structure of the receiver complexity.
- OFDM is computationally efficient by using simple FFT techniques to implement the modulation and demodulation functions.
- The OFDM signal has robustness in the multipath propagation environment and is more tolerant to delay spread⁹.

However, OFDM also has some shortcomings as follows

- Relatively higher peak-to-average power (PAPR) compared to a single carrier system, which tends to reduce the power efficiency of the radio frequency (RF) amplifier.
- OFDM is sensitive to frequency offsets, timing errors and phase noise.

More detailed discussion of these points can be found in [33] and [72].

⁹OFDM is robust against multipath effect, first because of using many narrowband subcarriers instead of one wide-band carrier for transmission and secondly because of the inserting of a cyclic prefix.

2.6 Synchronous Cooperative Relay Networks

As mentioned earlier in this chapter, a MIMO system can be used to mitigate the fading phenomena in wireless communication and increase the data transmission rate of the system [34]. It is affordable to have multiple antennas at the base station however it is impractical to equip the small mobile nodes with multiple antennas due to space constraints of the mobile node¹⁰, power limitation and hardware complexity of the mobile node [73]. In order to overcome these limitations, cooperative communications (also known as cooperative diversity or user cooperation), has been proposed in order to reap the benefits of MIMO communications systems in a wireless scenario with single antenna network nodes cooperatively transmitting and receiving by forming a virtual antenna array (VAA) [47]. Since cooperative communications is a recent paradigm for wireless relay networks, it has become a very active area of communications and networking research with promising developments, where the signals can be relayed from the source node to the destination node through half-duplex relay nodes. The basic idea behind cooperative communications was proposed in [74] wherein the relay channel was introduced and then investigated extensively in [75] to provide a number of relaying strategies to find achievable regions and provide upper bounds upon the capacity of a general relay channel in [74], a more detailed description of which can be found in [76]. The recent surge of interest in cooperative communications is represented by [24], [26], [19], [47], [77] and [78]. In [19] and [77], the concept and the potential benefits of cooperative communications have been intro-

¹⁰If the antennas are located close to each other, the channel fades may have some correlation which reduces the achievable diversity.

duced for a two user code division multiple access (CDMA) cellular network, where both users are active and orthogonal codes are used to avoid multiple access interference. Likewise, cooperative communication strategies have been analyzed in terms of their outage behavior in [24]. It has been shown that the outage gains and the cooperative diversity gain are proportion to the number of transmitting relay nodes. These benefits are achieved by deploying the relaying strategies rather than a direct link. In [47], the ideas in [24] have been extended to a more refined distributed MIMO multi-stage cooperative communications system, where cooperation between spatial distributed nodes not only takes place at the relaying stage, but also at the source node and the destination node. This will achieve the maximum cooperative diversity gain, nevertheless, at the expense of reducing the data rate¹¹. Furthermore, in [47] a significant discussion on the fractional power allocation at every stage in the network for different cooperative communications scenarios mentioned. Another technique known as coded cooperation has been introduced in [78] to achieve cooperative diversity and further analysis for coded cooperation can be found in [79]. A different approach, which can obtain the maximum spatial diversity gain, without suffering from the rate loss of repetition coding consists of combining the idea of cooperation with STBC, resulting in what is known as a distributed STBC. The case of distributed STBC has been first applied to wireless relay networks under the framework of cooperative communications in [26] and [17], in order to improve the bandwidth efficiency without losing any spatial diversity order (cooperative diversity) available in cooperative networks. As mentioned earlier

 $^{^{11}\}mathrm{If}$ the number of relaying stages increases, the total data transmission rate D_R will decrease.

in this section, this approach will involve a two phase protocol; in the first phase the relay nodes receive the signals from the source node and then in the second phase the relay nodes linearly process the signals received from the source node and forward them to the destination node such that the signals at the destination node appear as a STBC [29]. One such method can be found in [47], where orthogonal STBCs are deployed in distributed fashion for a VAA multi-stage network. Likewise, in [80] and [81] a simple Alamouti design [29] is introduced into cooperative relay networks because of its symbol-wise ML detector which achieves high diversity gain to combat channel impairments in wireless communications environments. In [27] and [28] a distributed linear dispersion coding scheme for a wireless relay network in which every node is equipped with a single antenna has been proposed, where the signals are sent by the relay nodes in the second phase as a linear function¹² of its received information. It has been shown that by cooperating distributively, the relay nodes generate a linear STBC codeword at the destination node and a full cooperative diversity gain of a multiple antenna system¹³ can be obtained, where the relay nodes are assumed to be synchronized at the destination node and also the destination node has full knowledge of the fading channels from the source node to the destination node. Furthermore, distributed linear dispersion STBC [56] among the relay nodes can be found in [82] and [83]. However, in [84], the idea of distributed STBC has been extended to wireless cooperative relay networks, in which every node has multiple antennas

¹²No decoding or demodulation is needed at relay nodes.

¹³Cooperative diversity gain equal to $R \times 1$, where R denotes the number of transmitting relay node antennas.

and the achievable cooperative diversity has been analyzed¹⁴. After these works, [28] has attempted to exploit the distributed coding techniques using orthogonal designs [29] and quasi-orthogonal designs [58] with more cooperating nodes, which leads to reliable communications in wireless relay networks. This approach provides many advantages such as maximum cooperative diversity, low ML decoding complexity and linear encoding of the information symbols. Regrettably, the maximum data transmission rate of orthogonal design [29] can not be achieved for more than two relay nodes [85] and the linear decoding complexity is no longer available. On the other hand, a quasi-orthogonal design [58] is only applicable to four relay nodes and its decoding complexity is higher than that of an orthogonal design. However, by applying feedback in the form of phase rotations [30], [31] and [32] at the relay nodes, it is possible to extract full data transmission rate at each hop and full cooperative diversity in proportion to the transmitting relay nodes from the orthogonal design [29] for more than two relay nodes and quasi-orthogonal design [58] for four relay nodes with simple linear decoding at the destination node. These approaches are presented in this thesis with the assumption of imperfect synchronization among the relay nodes and other distinctions are outlined in the introductory parts of Chapter 3 and 4. Most cooperative relay communication scenarios investigated so far are built upon the assumption that all nodes are equipped with a single half-duplex antenna, more recent work has however exploited the benefits of multiple antennas at the relay nodes. This model for the cooperative relay node is considered in Chapters 3, 4

¹⁴Cooperative diversity equal to $min\{M, N\}R$ if $M \neq N$ and equal to MR if M = N, where M denotes antennas at the source node, N denotes antennas at the destination node and R represent antennas at all the relay nodes.

and 5. The next section presents a review of research works that led to the deployments of STBCs in the context of asynchronous cooperative relay networks.

2.7 Asynchronous Cooperative Relay Networks

As mentioned in the previous section to reap the benefits of cooperative communications, it is necessary for the network to operate synchronously. However, perfect synchronization in cooperative relay networks is highly unlikely in practice [35]. This lack of synchronization results in ISI between the received signals from the relay nodes at the destination node, and also when the relay nodes use STBC to forward the received signals from the source node to the destination node, the code structure at the destination node is not orthogonal thereby preventing the transmitted signals from being successfully detected at the destination node with the normal STBC decoder. Therefore, the impact of imperfect synchronization between the relay nodes is one of the most critical and challenging issues when designing a space time cooperative relay network. Recently, there have been studies for asynchronous cooperative systems based on STCs to exploit cooperative diversity. In [86] and [87], a DF two-phase protocol with distributed space time trellis codes (STTCs) that achieves full cooperative diversity and coding gain without the synchronization assumption between relay nodes has been proposed. However, the decoding complexity becomes high when the number of relay nodes is not small, this is because the memory size of the corresponding STTCs grows exponentially in term of the number of relay nodes. To overcome this problem, in [88] the systematic construction of such a trellis code with minimum memory size for any number of relay nodes is developed to achieve full cooperative diversity in asynchronous cooperative communications. Furthermore, there has been limited work reported in the literature addressing distributed STBC regarding the mitigation of interference at the symbol level in [87] and [89], where an equalization technique is employed at the destination node to reduce the effect of imperfect synchronization among the relay nodes at the destination node, however, this leads to increased receiver complexity at the destination node. While in [90], [8], [91] and [92], the distributed STBC with PIC detection scheme for the case of two and four relay nodes is proposed and shown to be a very effective approach to mitigate the impact of imperfect synchronization at the destination node. Nevertheless, the schemes in [90] and [8] are limited by the number of relay nodes and can only achieve the diversity order between the relay nodes and the destination of two, also the work in [91] and [92] is limited since complex O-STBC with data transmission rate $D_R = 3/4^{15}$ is used between the relay nodes and the destination node. To achieve both full cooperative diversity and full data transmission rate for four relay nodes under the assumption of imperfect synchronization, distributed closed-loop EO-STBC [93] and closed-loop QO-STBC [94] with PIC detection scheme as in Section 2.5 are proposed. All these proposed schemes have the potential of delivering cooperative diversity order of cooperating relay nodes using a DF protocol, which leads to complexity increase at the relay nodes. To avoid this complexity, the proposed scheme in [27] and [28] is extended to the case of imperfect synchronization in [50] and [95], where the signals broadcast by all relay nodes are designed as linear func-

¹⁵Because three symbols transmit from the relay nodes over four time symbol transmission periods.

tions of their received information, therefore, no decoding operation at the relay nodes is required as compared to the DF scheme. However, all distributed STBCs based on the PIC detection scheme mentioned above have the disadvantage that their computational complexity is dependent upon the number of PIC detection iterations. It has been shown in simulation results that the PIC detection scheme is very effective in mitigating synchronization error and normally two or three iterations will deliver most of the performance gain. While in [48], an equalizer has been proposed to combat the interference components caused by timing misalignment in a decision feedback manner, and then a symbol-by-symbol ML detection is implemented. Nevertheless, the low computational complexity of the Alamouti code remains but this only applies in the case of two cooperative relay nodes and cannot be extended to more than two cooperative relay nodes. The sub-optimum detection scheme in [9] canceled the interference components caused by asynchronism effectively with near-Alamouti simplicity, unity rate code and achieved the full diversity order between the relay nodes and the destination node. However, this detection scheme relied mainly on the existence of a direct transmission (DT) link between the source node and the destination node which is unrealistic and can be difficult to achieve in practice. Furthermore, the work in [35] - [96] dealt with the timing error problem in cooperative relay networks under the assumption that the relaying process operates with the DF type of communication protocol. Whereas [95] adopted the AF type of transmission utilizing the iterative PIC detection scheme to mitigate the impact of asynchronism in addition to the utilization of a two-bit feedback scheme in [30] and assumes that a DT link must exist for a successful detection. Recently, a novel detection scheme for two dual-antenna relay nodes under imperfect synchronization is proposed [97] to overcome the drawback in [48], [9] and [95]. All the above works have been performed under the assumption that the wireless transmission channels are frequency flat fading channels. In [98] and [99] a distributed A-STBC transmission scheme based on OFDM for asynchronous AF protocol is proposed to cancel the timing error among the relay nodes and achieve asynchronous cooperative diversity, where the relay nodes only need to perform a few very simple operations, time reversion and complex conjugation, and the destination node has the Alamouti code structure. However, this proposed scheme is only valid for the case of two relay nodes. The idea in [98] is extended to the case of any number of relay nodes in [100]. Unfortunately, full cooperative diversity and the data transmission rate¹⁶ decreases as the number of transmitting relay nodes increase, in other words, O-STBC generated from complex constellations for more than two relay nodes do not exist [28]. Furthermore, the work in [101] presented a closed-loop scheme for use in asynchronous cooperative networks over frequency selective channels. In addition to implementing four relay nodes which incurred more time delay, the feedback scheme also required some quantization technique to reduce the overhead incurred by this proposed scheme. Most proposed approaches in wireless relay networks are built upon the assumption that all relay nodes are equipped with a single antenna. As mentioned earlier above in this literature survey to achieve both full cooperative diversity and full data transmission rate in each hop for more than two relay nodes using O-STBC, feedback schemes are required. Therefore,

¹⁶Since a half-duplex protocol is used, the data transmission rate of two-hop relay network is actually unity for each hop.

in [102] an asynchronous cooperative system is assumed using the M-D-QO-STBC scheme in which distributed OFDM transmission is used with two relay nodes each equipped with two antennas over frequency flat channels. Whereas, in [103] a new closed-loop scheme is proposed for D-EO-STBC over frequency selective fading channels in wireless relay networks utilizing one-bit feedback per subcarrier based on selection cyclic phase rotation. Distributed OFDM transmission is used with two relay nodes each equipped with two antennas rather than four relay nodes each has a single antenna to reduce timing error and decoding complexity at the destination node.

2.8 Chapter Summary

This chapter has reviewed research work that has been undertaken prior to the work presented in this thesis. The first section of this chapter focussed on the achievements made in the area of cooperative wireless communications in information theoretic terms to the various developments in distributed STBC under both cases of perfect synchronization and imperfect synchronization. In MIMO, the STBCs used in order to exploit the transmit diversity offered by multiple antennas are introduced, leading to the A-STBC, O-STBCs, QO-STBCs and EO-STBCs. The performance advantages of each were highlighted and also the principle of the OFDM technique with its advantages and disadvantages were highlighted. The chapter then proceeded to discuss earlier attempts to expand wireless network coverage with the use of cooperative relay networks. Various theoretical and practical issues were also reviewed. Motivated by the difficulty in deploying multiple antennas in mobile terminals, it was then highlighted that researchers have focused their attention into ways of using multiple single antenna elements to create a MIMO channel through the use of cooperation, this was later extended to cooperative relay networks and the advantages of distributed STBCs was also exploited. Furthermore, the problem of synchronization among the relay nodes was addressed, which is the main research focus in this thesis. In the next chapter, the design of a coded D-CL-EO-STBC and coded D-CL-QO-STBC with PIC detection for four relay nodes and two dual antenna relay nodes respectively, under imperfect synchronization is proposed. The PIC detection scheme in both designs is deployed to mitigate asynchronism between the relay nodes and the destination node which proved to be efficient from the complexity point of view. The LD cooperative strategy is employed for both scenarios achieving cooperative diversity order of four in both cases with a unity data rate in each transmission hop.

DISTRIBUTED CLOSED-LOOP STBC WITH OUTER CODING FOR ASYNCHRONOUS COOPERATIVE RELAY NETWORKS

This chapter investigates the cooperative strategy for distributed spacetime block coding (STBC) design which is based on a linear dispersion code and exploits outer convolutive coding for asynchronous cooperative relay networks. Furthermore, to overcome the lack of synchronization and achieve high performance, a parallel interference cancelation (PIC) detection scheme with distributed closed-loop extended orthogonal STBC (CL-EO-STBC) design with outer coding for four relay nodes and distributed closed-loop quasi orthogonal STBC (CL-QO-STBC) design with outer coding for two dual-antenna relay nodes are considered. Finally, the pairwise error probability (PEP) is analyzed to reveal the available cooperative diversity which can be utilized with distributed distributed CL-EO-STBC and distributed CL-QO-STBC within cooperative relay networks.

3.1 Introduction

Multiple antennas have been shown to greatly increase the reliability of a wireless communication link in a fading environment using space-time coding [27]. However, cost constraints, size limitations and hardware complexity make it difficult to co-locate multiple antennas at one mobile communication node. Therefore, researchers have been looking for different techniques to exploit spatial diversity using antennas of different users in the network [19] [104], and in [17] STBC is extended to cooperative systems in the form of distributed STBC (D-STBC) enhancing the reliability of the transmission and significantly improving the performance gain [19] [17] [18].

In practical systems, the assumption that cooperative relay nodes transmit the corresponding symbols to the destination node at the same time in a perfectly synchronized manner is difficult if not impossible to achieve due to different locations of relay nodes and individual clocks/timing circuits. The asynchronism among relay nodes, in turn, causes channels to be dispersive even under flat fading conditions and will damage the orthogonality of STBC by introducing inter-symbolinterference (ISI) between received symbols from different relay nodes at the destination node, which can lead to substantial degradation in performance.

On the other hand, the most commonly utilized strategies with various cooperative relaying protocols are either amplify-and-forward (AF) or decode-and-forward (DF) [51] [17] [18]. In [8] and [93] asynchronous cooperative relay networks are demonstrated using uncoded distributed A-STBC [29] for two relay nodes and distributed CL-EO-STBC for four relay nodes with the PIC technique at the destination node, respectively. These proposed schemes have the potential to deliver cooperative diversity order equal to the number of cooperating relay nodes between the relay nodes and the destination node using a DF strategy, which leads to relay node complexity. Whereas in [27], [28] and [8] authors considered a new cooperative relaying strategy in which the relay nodes forward a distributed linear dispersion (LD) code to the destination node without any decoding operations at the relay nodes. This is a form of AF type of transmission strategy, which does not need channel information at the relay nodes, but full channel information from the source node to the relay nodes and from the relay nodes to the destination node is required at the destination node. Furthermore, compared to the DF strategy, it does not need any decoding operation at the relay nodes, which leads to saving in power and time at the relay nodes. Most importantly, it achieves cooperative diversity gain in proportion to the number of transmitting antennas at the relay nodes.

To overcome the complexity at the relay nodes, a distributed STBC strategy in [27] and [28] is designed for asynchronous cooperative relay networks utilizing the distributed A-STBC design with outer convolutive coding for two relay nodes in [50]. Applying PIC detection at the destination node can mitigate the impact of imperfect synchronization caused by timing misalignment from the relay nodes, delivers full data transmission rate in each stage and achieves full cooperative diversity gain with some coding gain. However, using distributed A-STBC at

the relay nodes is strictly limited by the number of cooperating relay nodes and only cooperative diversity order of two can be exploited between the relay nodes and the destination. Within this context, this chapter focuses on the design of a coded D-CL-EO-STBC and coded D-CL-QO-STBC with PIC detection for four relay nodes and two dual antenna relay nodes respectively, under imperfect synchronization. The PIC detection scheme in both designs is deployed to mitigate asynchronism between the relay nodes and the destination node which proved to be efficient from the complexity point of view. The LD cooperative strategy is employed for both scenarios achieving cooperative diversity order of four in both cases with a unity data rate in each transmission hop.

Although, both design approaches exploit the fourth diversity order with coding gain and unity data transmission rate in each transmission hop when utilizing CL-QO-STBC, the complexity was reduced as compared to CL-EO-STBC due to the reduction of the timing error among the relay nodes when equipped with two antennas on each relay node. Finally, the PEP analysis for the proposed designs is presented to confirm the available cooperative diversity.

3.2 Coherent D-CL-STBC in asynchronous cooperative relay networks

This section investigates the effects, and assesses the impact, of asynchronism on cooperative relay networks. The four relay nodes are each equipped with a single antenna and implementing the CL-EO-STBC is discussed first, followed by the two dual-antenna model which utilizes the CL-QO-STBC scheme.

3.2.1 Using the D-CL-EO-STBC Design for four Relay Nodes

A coded two-hop cooperative relay network with single source and destination nodes communicating via four parallel relay nodes, used for both transmission and reception, is depicted in Figure 3.1. It is assumed that there is a direct transmission link (DT) between the source node and the destination node and every participating node in the network communicates utilizing a single transmit and receive antenna configuration and assumed to operate in a half-duplex mode, thereby generating a virtual multiple antennas system. The narrowband flat



Figure 3.1. Basic structure of distributed CL-EO-STBC with outer coding using two-bit feedback based on phase rotation for four relay nodes with single antenna in each relay node and one antenna in the source and the destination node with two phases for the cooperative transmission process.

fading broadcasting channel gain from the source node to each relay node is represented as f_k , the relaying channel gain from the k^{th} relay node to the destination node is represented as g_k , where $k \in \{1, 2, ..., 4\}$, and the channel between the source node and the destination node is represented as h_{sd} . It is assumed that all random channel parameters¹ f_k , g_k and h_{sd} are spatially uncorrelated and they are assumed to be zero mean circularly symmetrical complex Gaussian random variable with unity variance. It is assumed that the relays know only the statistical distribution of the channels. However, the assumption that the receiver knows all the fading coefficients f_k and g_k is made valid. Knowledge of the channels can be obtained by sending training signals from the relays and the transmitter. Many types of gains can be obtained from the network, for example, gains on the capacity and gains on the error rate. This chapter focuses on the improvement in the end-to-end error rate, which can be obtained by the pairwise error probability (PEP). The distributed open loop EO-STBC code word **S** that needed to be generated at the destination node has the following form

$$\mathbf{S} = \begin{bmatrix} s(1,n) & s(1,n) & s(2,n) & s(2,n) \\ -s^*(2,n) & -s^*(2,n) & s^*(1,n) & s^*(1,n) \end{bmatrix}$$
(3.2.1)

The cooperative communication is divided into two phases (hops). During the first phase, the transmitter at the source node broadcasts the mapped QPSK symbols $\mathbf{s}(n) = [s(1,n), -s^*(2,n)]^T$, where *n* denotes the discrete pair index in two different symbol transmission periods, to the relay nodes as well as the destination node through the DT link. This transmission process takes place after performing the encoding and interleaving processes as shown in Figure 3.1. The received signal vector at the destination node through the DT link can be modelled as

¹In practice, the channel parameters can be correlated due to small angular spread and/or not large enough antenna spacing, this will reduce the diversity gain. Estimation errors in these parameters are not considered in this Thesis.

follows

$$\mathbf{r}_{sd}(n) = \sqrt{P_1} h_{sd} \mathbf{s}(n) + \mathbf{n}_{sd}(n)$$
(3.2.2)

where P_1 is the transmit power used at the source node for every channel use and $\mathbf{n}_{sd}(n)$ is the corresponding additive Gaussian noise vector at the destination node with elements with zero-mean and unit-variance. Similarly, the received signals at every relay node can be written as follows

$$\mathbf{r}_k(n) = \sqrt{P_1} f_k \mathbf{s}(n) + \mathbf{v}_k(n)$$
(3.2.3)

where $\mathbf{v}_k(n)$ is the additive Gaussian noise vector at each relay node with elements with zero-mean and unit-variance. During the second phase, the relay nodes will process and transmit the received noisy signals to the destination node. The transmitted signal vectors $\mathbf{t}_k(n)$, $k \in 1, ..., 4$ from the relay nodes are designed to be a linear function of the received signal and its conjugate as follows

$$\mathbf{t}_{k}(n) = \sqrt{\frac{P_{1}P_{2}}{P_{1}+1}} (f_{k}A_{k}\mathbf{s}(n) + f_{k}^{*}B_{k}\mathbf{s}^{*}(n)) + \sqrt{\frac{P_{2}}{P_{1}+1}} (A_{k}\mathbf{v}_{k}(n) + B_{k}\mathbf{v}_{k}^{*}(n))$$
(3.2.4)

where P_2 is the average transmission power on every relay node. Due to the unity variance assumption of the elements of the additive noise $\mathbf{v}_k(n)$ vector from the source node to the relay nodes in (3.2.3), the average power of the signal on every relay node is P_1+1 . If the total power per symbol transmission used in the whole network is fixed as P, the optimal power allocation used as in [105] (and the references therein) that maximizes the expected received signal-to-noise ratio (SNR) at the destination node can be such that

$$P_1 = \frac{P}{2}$$
 and $P_2 = \frac{P}{2R}$

where R represents the number of relay nodes. The matrices A_k and B_k at each relay node are used as in [28] where $k \in 1, ..., 4$ and the relay nodes have been designed to use the following matrices

$$A_{1}, A_{2} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, B_{1}, B_{2} = 0_{2 \times 2},$$
$$A_{3}, A_{4} = 0_{2 \times 2}, B_{3}, B_{4} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$$
(3.2.5)

In order to achieve the maximum cooperative diversity gain between the source node and the destination node, the relay nodes are scheduled to broadcast their signals simultaneously. However, synchronous reception at the destination node is often unrealistic due to propagation delay among the relay nodes as shown in Figure 3.2. There is normally a timing misalignment of $\tau_2 = \tau_3 = \tau_4 \neq 0$ from relay nodes R_k , where $k \in 2, 3, 4$. The time delay τ_k is assumed to be smaller than a sample period and, without loss of generality, all time delays τ_k are also assumed to be identical due to their co-location in the same geographical area as shown in Figure 3.2. Such a relative time delay from R_k , where $j \in 2, 3, 4$ will induce ISI from neighboring symbols at the destination node owing to sampling or matched filtering (whatever pulse shaping is used) [8]. Nevertheless, the destination node is assumed to be fully synchronized to R_1 i.e. $\tau_1 = 0$.

Therefore, the received signal vectors at the destination node at two



Figure 3.2. Representation of time delay of imperfect synchronization between relay nodes at the destination node, which induces ISI from R_2 , R_3 and R_4 [8].

different transmission periods can be represented as follows

$$\mathbf{r}_{rd}(n) = \sum_{k=1}^{4} g_k \begin{bmatrix} t_k(1,n) \\ t_k(2,n) \end{bmatrix} + \underbrace{\sum_{k=2}^{4} g_k(-1) \begin{bmatrix} t_k(2,n-1) \\ t_k(1,n) \end{bmatrix}}_{\mathbf{I}_{int}(n)} + \begin{bmatrix} w_{rd}(1,n) \\ w_{rd}(2,n) \end{bmatrix}$$
(3.2.6)

where $\mathbf{r}_{rd}(n) = [r_{rd}(1, n), r_{rd}(2, n)]^T$ is the received signal at the destination node over two different transmission periods under imperfect synchronization level, $w_{rd}(1, n)$ and $w_{rd}(2, n)$ are the total additive Gaussian noise terms at the destination node, $\mathbf{I}_{int}(n) = [I_{int}(1, n), I_{int}(2, n)]^T$ denotes the total ISI vector at the destination node and $g_k(-1)$ reflects the ISI from the previous symbols under synchronization error, where $k \in 2, 3, 4$. Such assumptions will damage the orthogonality of distributed EO-STBC, also this will lead to substantial system performance degradation. Therefore, the coefficients of $g_k(-1)$ depends upon timing delay τ_k , where $k \in 2, 3, 4$ and the particular pulse shaping waveform used and its relative strength can be represented by a ratio as follows

$$\beta_k = \frac{|g_k(-1)|^2}{|g_k|^2} \tag{3.2.7}$$

By taking the conjugate of $r_{rd}(2, n)$ in (3.2.6), the equivalent channel matrix **H** corresponding to the codeword in (3.2.1) used over four relay nodes is given by

$$\mathbf{H} = \begin{bmatrix} f_1 g_1 + f_2 g_2 & f_3^* g_3 + f_4^* g_4 \\ f_3 g_3^* + f_4 g_4^* & -f_1^* g_1^* - f_2^* g_2^* \end{bmatrix}$$
(3.2.8)

Applying matched filtering at the destination node, namely forming the $\mathbf{H}^{H}\mathbf{H}$ matrix, the Grammian matrix Δ can be represented by

$$\Delta = \begin{bmatrix} \lambda_d + \lambda_f & 0\\ 0 & \lambda_d + \lambda_f \end{bmatrix}$$
(3.2.9)

Although the decoding complexity is low, the λ_f term may be negative which will lead to some diversity loss. In order to achieve a full cooperative diversity gain between the relay nodes and the destination node, a two-bits feedback scheme $U_1 = (-1)^a$ and $U_2 = (-1)^b$, where $a, b \in 0, 1$ is used to rotate the phases of signals before they are transmitted from the first and third relay nodes respectively to ensure that the λ_f term is positive during the whole transmission, while the other two relay nodes are kept unchanged as shown in Figure 3.1 [31]. Therefore, the received signal in (3.2.6) at two independent transmission periods after taking the conjugate of $r_{rd}(2,n)$ can be re-written as follows

$$r_{rd}(1,n) = t_1(1,n)U_1g_1 + t_2(1,n)g_2 + t_3(1,n)U_2g_3 + t_4(1,n)g_4$$

+ $I_{int}(1,n) + w(1,n)$
$$r_{rd}^*(2,n) = t_1^*(2,n)U_1^*g_1^* + t_2^*(2,n)g_2^* + t_3^*(2,n)U_2^*g_3^* + t_4^*(2,n)g_4^*$$

+ $I_{int}^*(2,n) + w^*(2,n)$ (3.2.10)

where

$$I_{int}(1,n) = \sqrt{\frac{P_1 P_2}{P_1 + 1}} (g_2(-1)f_2 s(2, n - 1) + (g_3(-1)U_2 f_3^*) + g_4(-1)f_4^*) s^*(1, n - 1))$$

$$I_{int}^*(2,n) = \sqrt{\frac{P_1 P_2}{P_1 + 1}} (g_2^*(-1)f_2^* s^*(1, n) + (g_3^*(-1)U_2^* f_3) + g_4^*(-1)f_4) s^*(2, n))$$
(3.2.11)

and

$$w(1,n) = \sqrt{\frac{P_2}{P_1 + 1}} (U_1 g_1 v_1(1,n) + g_2 v_2(1,n) + U_2 g_3 v^*(1,n) + g_4 v_4^*(1,n)) + w_{rd}(1,n) w^*(2,n) = \sqrt{\frac{P_2}{P_1 + 1}} (U_1^* g_1^* v_1^*(2,n) + g_2^* v_2^*(2,n) + U_2^* g_3^* v(2,n) + g_4^* v_4(2,n)) + w_{rd}^*(2,n)$$
(3.2.12)

where $w_{rd}(1,n)$ and $w_{rd}^*(2,n)$ are the noise vectors at the destination node whose entries represent independent complex Gaussian random variables with zero-mean and unity variance and the equivalent channel matrix \mathbf{H} in (3.2.8) becomes

$$\mathbf{H} = \begin{bmatrix} U_1 f_1 g_1 + f_2 g_2 & U_2 f_3^* g_3 + f_4^* g_4 \\ U_2^* f_3 g_3^* + f_4 g_4^* & -U_1^* f_1^* g_1^* - f_2^* g_2^* \end{bmatrix}$$
(3.2.13)

Consequently, the SNR at the destination node can be calculated as follows

$$SNR = (\underbrace{\sum_{k=1}^{4} (|f_k g_k|^2)}_{\lambda_d} + \underbrace{2\Re(U_1 f_1 g_1 f_2^* g_2^* + U_2 f_3^* g_3 f_4 g^*)}_{\lambda_f} \frac{\sigma_s^2}{\sigma_n^2} \quad (3.2.14)$$

where σ_s^2 is the total transmit power of the desired signal and σ_n^2 is the noise power at the destination node. It is clear from the above analysis in (3.2.14) that if $\lambda_f > 0$, the designed CL-EO-STBC can obtain additional performance gain, which leads to an improved overall channel gain and correspondingly the SNR at the destination node. According to the analysis in [31], the values of a and b that, in turn, determine the exact values of U_1 and U_2 can be proposed according to the following design criteria

1

$$a = \begin{cases} 0, & \text{if } f_1 g_1 f_2^* g_2^* \ge 0 \\ 1, & \text{otherwise} \end{cases}$$
(3.2.15)

$$b = \begin{cases} 0, & \text{if } f_3^* g_3 f_4 g_4^* \ge 0 \\ 1, & \text{otherwise} \end{cases}$$
(3.2.16)

Since the equivalent channel matrix \mathbf{H} is known at the destination node and the equivalent noise vector $\mathbf{w}(n)$ is circularly symmetric complex Gaussian random vector whose mean is zero and covariance matrix

is
$$\left[1 + \frac{P_2}{P_1 + 1}\left(\sum_{k=2,4} |g_k|^2\right) + |U_1 g_1|^2 + |U_2 g_3|^2\right] \mathbf{I}_2$$
. Therefore, the ML decoding at the destination node is

$$\hat{\mathbf{s}}(n) = \arg\min_{s_k(n)\in\mathbf{S}} \left\| \mathbf{r}_{rd}(n) - \sqrt{\frac{P_1P_2}{P_1+1}} \mathbf{H}\mathbf{s}_k(n) \right\|_F^2$$
(3.2.17)

where **S** is the set of all possible vector symbols. The ML detection in (3.2.17) will suffer from synchronization error at the destination node, because the $\mathbf{I}_{int}(n) = [I_{int}(1,n), I_{int}^*(2,n)]^T$ component in (3.2.11) will destroy the orthogonality of distributed CL-EO-STBC at the destination node.

Similarly, the analysis is considered for the case of CL-QO-STBC scheme.

3.2.2 Using CL-QO-STBC Design for Two Dual Antenna Relay Nodes

In this section the use of CL-QO-STBC design in networks with two relay nodes is considered, where each relay node is equipped with two half-duplex antennas that can be used for both transmission and reception, while the source node as well as the destination node are equipped with a single antenna. Furthermore, the DT is assumed to be available between the source node and the destination node and the imperfect synchronization issue is considered as well. The basic structure of this model is depicted in Figure 3.3. The narrowband flat fading broadcasting channel gain from the source node to each antenna on each relay node is represented as f_{ik} , the relaying channel gain from the i^{th} antenna on the k^{th} relay node to the destination node is represented as g_{ik} , where $i, k \in 1, 2$, and the channel between the source node and the destination node is represented as h_{sd} . It is assumed that



Figure 3.3. Basic structure of distributed CL-QO-STBC with outer coding using two-bit feedback based on phase rotation for two relay nodes equipped with two antennas in each relay node and one antenna in the source and the destination node with two phases for the cooperative transmission process

all random channel parameters² f_{ik} , g_ik and h_{sd} are spatially uncorrelated and they are assumed to be zero mean circularly symmetrical complex Gaussian random variable with unity variance. It is assumed that the relays know only the statistical distribution of the channels. However, the assumption that the receiver knows all the fading coefficients f_{ik} and g_{ik} is made valid. Knowledge of the channels can be obtained by sending training signals from the relays and the transmitter. During the first phase, the transmitter at the source node broadcasts $\mathbf{s}(n) = [s(1,n), -s^*(2,n), -s^*(3,n), s(4,n)]^T$ to the destination node through the DT link as well as to the relay nodes, where *n* denotes the discrete pair index in four different symbol transmission periods.

²In practice, the channel parameters can be correlated due to small angular spread and/or not large enough antenna spacing, this will reduce the diversity gain. Estimation errors in these parameters are not considered in this Thesis.
modelled as in (3.2.2). Also the received signal vectors at each antenna on every relay node can be written as follows

$$\mathbf{r}_{ik}(n) = \sqrt{P_1} f_{ik} \mathbf{s}(n) + \mathbf{v}_{ik}(n)$$
(3.2.18)

where P_1 denotes the average power per transmission used at the source node, and $\mathbf{v}_{ik}(n)$ is the additive Gaussian noise vector at the i^{th} antenna on the k^{th} relay node with zero-mean and unit-variance. The distributed open-loop QO-STBC codeword **S** that needed to be generated at the destination node has the following form

$$\mathbf{S} = \begin{bmatrix} s(1,n) & s(2,n) & s(3,n) & s(4,n) \\ -s^{*}(2,n) & s^{*}(1,n) & -s^{*}(4,n) & s^{*}(3,n) \\ -s^{*}(3,n) & -s^{*}(4,n) & s^{*}(1,n) & s^{*}(2,n) \\ s(4,n) & -s(3,n) & -s(2,n) & s(1,n) \end{bmatrix}$$
(3.2.19)

Therefore, the code matrix **S** in (3.2.19) presents a unity data transmission rate on each transmission hop. In the second phase, the transmitted signal vectors from each antenna on every relay node $\mathbf{t}_{ik}(n)$ is designed to be a linear function of its received signal and its conjugate as follows

$$\mathbf{t}_{ik}(n) = \sqrt{\frac{P_1 P_2}{P_1 + 1}} (f_{ik} A_{ik} \mathbf{s}(n) + f_{ik}^* B_{ik} \mathbf{s}^*(n)) + \sqrt{\frac{P_2}{P_1 + 1}} (A_{ik} \mathbf{v}_{ik}(n) + B_{ik} \mathbf{v}_{ik}^*(n))$$
(3.2.20)

where A_{ik} and B_{ik} are the matrices at the i^{th} antenna on the k^{th} relay node used as in [28] and can be modelled as follows

$$A_{11} = I_4, \ A_{21} = 0_4, \ A_{12} = 0_4, \ A_{22} = \begin{bmatrix} 0 & 0 & 0 & 1 \\ 0 & 0 & -1 & 0 \\ 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{bmatrix}$$
$$B_{11} = B_{22} = 0_4, \ B_{21} = \begin{bmatrix} 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \end{bmatrix}, \ B_{12} = \begin{bmatrix} 0 & 0 & -1 & 0 \\ 0 & 0 & 0 & -1 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}$$

The average transmission power at each antenna on every relay node is denoted by P_2 . As mentioned in the previous section, due to the unity variance assumption of the elements of additive noise vector $\mathbf{v}_{ik}(n)$ from the source node to the relay nodes in (3.2.20), the average power of the signal at each antenna on every relay nodes is $P_1 + 1$. If the total power per symbol transmission used in the whole network is fixed as P, the optimal power allocation used as in [105] (and the references therein) that maximizes the expected received signal-to-noise ratio (SNR) at the destination node can be such that

$$P_1 = \frac{P}{2}$$
 and $P_2 = \frac{P}{4R}$

where R represents the number of relay nodes. Due to various factors such as different propagation delays, the signals from different antennas on different relays will most likely arrive at the destination node at different time instants. Since accurate synchronization is difficult or impossible [8], there is practically a time misalignment of τ_{i2} between



Figure 3.4. Representation of time delay of imperfect synchronization between relay nodes and the destination node, which induces ISI from the two antennas of R_2 .

the received copies of these signals. Considering the required effort of synchronization due to the fact that relay nodes are usually close to the source node, the assumption of τ_{i2} being smaller than the symbol period is valid as mentioned in the previous section. This relative time delay will induce ISI from adjacent symbols at the destination node as shown in Figure 3.4. Without loss of generality, the receiver at the destination node is assumed to be synchronized to R_1 and the time delays at the received symbols from R_2 , i.e. τ_{i2} , are assumed to be identical. The received signal vector at the destination node can be represented as follows

$$\mathbf{r}_{rd}(n) = \sum_{i=1}^{2} \left(g_{i1} \mathbf{t}_{i1}(n) + g_{i2} \mathbf{t}_{i2}(n) + g_{i2}^{(-1)} \mathbf{I}_{int}(n) \right) + \mathbf{w}(n)$$
(3.2.21)

where $\mathbf{r}_{rd}(n)$ is the received signal vector at the destination node over four different transmission periods under imperfect synchronization, $\mathbf{t}_{ik}(n)$ denotes the transmitted signal vector from each antenna of each relay node, $\mathbf{I}_{int}(n)$ denotes the total ISI vector at the destination node, $g_{i2}^{(-1)}$ reflects the ISI from the previous symbols under synchronization error from the i^{th} antenna of the second relay, where $i, k \in 1, 2$, and $\mathbf{w}(n)$ is the total additive Gaussian noise vector term at the destination node. The relative strength of $g_{i2}^{(-1)}$ will be represented by the ratio [8]

$$\beta_{i2} = |g_{i2}^{(-1)}|^2 / |g_{i2}|^2 \quad \text{for} \quad i \in 1, 2$$
(3.2.22)

The equivalent Grammian matrix $\Delta = \mathbf{H}^H \mathbf{H}$ can be presented as

$$\Delta = \begin{bmatrix} \alpha & 0 & \beta & 0 \\ 0 & \alpha & 0 & \beta \\ \beta & 0 & \alpha & 0 \\ 0 & \beta & 0 & \alpha \end{bmatrix}$$
(3.2.23)

The appearing nonzero off-diagonal terms cause coupling between the estimated symbols, therefore, reducing the cooperative diversity gain of the code matrix in (3.2.19) between the relay nodes and the destination node. The bit error rate (BER) performance of this distributed openloop QO-STBC therefore is degraded. To overcome this problem, two feedback methods for this distributed QO-STBC are used to achieve full cooperative diversity gain with unity data transmission rate between the relay nodes and the destination node as presented in Section 2.2. The signals transmitted from the first and the second antennas of second relay node relay node are instead rotated by $U_1 = e^{j\phi}$ and $U_2 = e^{j\theta}$ respectively as shown in Figure 3.3, while the other two antennas of the first relay node are kept unchanged. The phase rotation on the transmitted symbols is equivalent to rotating the phases of the corresponding channel coefficients as shown in Figure 3.3. As a result, the received signal vectors from the four antennas at the two relay nodes after applying the feedback scheme are:

$$\mathbf{r}_{rd}(n) = \sum_{i=1}^{2} \left(g_{i1} \mathbf{t}_{i1}(n) + U_i g_{i2} \mathbf{t}_{i2}(n) + g_{i2}^{(-1)} \mathbf{I}_{int}(n) \right) + \mathbf{w}(n)$$
(3.2.24)

where

$$I_{int}(1,n) = \sum_{i=1}^{2} U_i t_{i2}(4,n-1)$$

$$I_{int}(t,n) = \sum_{i=1}^{2} U_i t_{i2}(t,n) \quad \text{for} \quad t = 2, 3, 4 \quad (3.2.25)$$

Equivalently, the diagonal and off-diagonal elements in (3.2.23) will be multiplied by U_1 and U_2

$$\alpha = \sum_{i=1}^{2} (|g_{i1}f_{i1}|^2 + |U_ig_{i1}f_{i1}|^2)$$
(3.2.26)

and

$$\beta = 2\Re\{f_{11}g_{11}f_{22}^*g_{22}^*U_1 - f_{21}g_{21}f_{12}^*g_{12}^*U_2\}$$
(3.2.27)

Therefore, the received signals at the destination node in (3.2.24) can be represented in vector form as follows

$$\mathbf{r}_{rd}(n) = \mathbf{Hs}(n) + \mathbf{I}_{int}(n) + \mathbf{w}(n)$$
(3.2.28)

where **H** becomes

$$\mathbf{H} = \begin{bmatrix} f_{11}g_{11} & f_{12}^{*}g_{12} & U_{1}f_{21}^{*}g_{21} & U_{2}f_{22}g_{22} \\ f_{12}g_{12}^{*} & -f_{11}^{*}g_{11}^{*} & U_{2}^{*}f_{22}^{*}g_{22}^{*} & -U_{1}^{*}f_{21}g_{21}^{*} \\ U_{1}^{*}f_{21}g_{21}^{*} & U_{2}^{*}f_{22}^{*}g_{22}^{*} & -f_{11}^{*}g_{11}^{*} & -f_{12}g_{12}^{*} \\ U_{2}f_{22}g_{22} & -U_{1}f_{21}^{*}g_{21} & -f_{12}^{*}g_{12} & f_{11}g_{11} \end{bmatrix}$$
(3.2.29)

The ML decoding is used to detect which symbols reach the destination node as in (3.2.17). As mentioned in the previous section, the interference term induced in the received signals in (3.2.17) will destroy the orthogonality of the distributed CL QO-STBC and the ML detection will suffer from this.

So far, the analysis has shown the effect of asynchronism and its impact on wireless cooperative relay networks introducing an interference factor that causes the conventional ML detector to fail regardless of the type of STBC utilized in transmission. Therefore, it is considered crucial to mitigate the interference factor at the destination to achieve a robust and reliable communications. With this intention, the next section introduces the PIC detection method for both system models discussed earlier in this chapter.

3.3 Parallel Interference Cancellation Detection (PIC)

As discussed in Sections (3.2.1) and (3.2.2), signals transmitted from the antennas of each relay node are received asynchronously at the destination node. Such assumptions will damage the orthogonality for any kind of distributed STBC, also this will lead to substantial system performance degradation. Therefore, the coefficients of $g_k(-1)$, $k \in 2, 3, 4$, in the case of four relay nodes when EO-STBC is utilized, and the coefficients of $g_{i2}^{(-1)}$, $i \in 1, 2$, in the case of two relay nodes when QO-STBC is utilized, depend upon timing delay τ_k and τ_{i2} respectively. The use of PIC is a promising detection technique for interference cancellation [8]. Furthermore, this technique relies on simple processing elements constructed around the matched filter concept. Therefore, the PIC detection scheme is used with the proposed schemes in Sections (3.2.1) and (3.2.2) to remove the interference term $\mathbf{I}_{int}(n)$ in (3.2.6) and (3.2.28) for both designs. From (3.2.11) and (3.2.25), the interference term of $I_{int}(1, n)$ can be expressed as:

• In the case of CL-EO-STBC

$$I_{int}(1,n) = (g_2(-1)f_2s(2,n-1) + (g_3(-1)U_2f_3^* + g_4(-1)f_4^*)s^*(1,n-1))$$
(3.3.1)

• In the case of CL-QO-STBC

$$I_{int}(1,n) = -U_1 g_{12}(-1) f_{12}^* s(2,n-1) + U_2 g_{22}(-1) f_{22} s(1,n-1)$$
(3.3.2)

Since $s^*(1, n-1)$ and s(2, n-1) in (3.3.1) and s(1, n-1) and s(2, n-1)in (3.3.2) are in fact known if the detection process has been initialized properly, $I_{int}(1, n)$ in (3.3.1) and (3.3.2) can be removed during the initialization stage. Therefore, the general PIC iteration process can be applied to eliminate the interference term of $\mathbf{I}_{int}(n)$ for both proposed schemes as follows

- 1. Initialization
- 2. Set iteration q = 0
- 3. From the received signal vector $\mathbf{r}_{rd}(n)$ in (3.2.6) for the case of distributed CL-EO-STBC and in (3.2.28) for the case of distributed CL-QO-STBC, calculate
 - In the case of CL-EO-STBC

$$\mathbf{r}_{rd}^{\prime(0)}(n) = \begin{bmatrix} r_{rd}(1,n) - I_{int}(1,n) \\ r_{rd}^{*}(2,n) \end{bmatrix}$$

• In the case of CL-QO-STBC

$$\mathbf{r}_{rd}^{\prime(0)}(n) = \begin{bmatrix} r_{rd}(1,n) - I_{int}(1,n) \\ r_{rd}^{*}(2,n) \\ r_{rd}^{*}(3,n) \\ r_{rd}(4,n) \end{bmatrix}$$

Due to the poor performance of the ML detector within distributed STBC in (3.2.17) [8] and [91], from the received signal of the DT link between the source node and the destination node in (3.2.2), the DT link detection result can be calculated and used as follows

$$\hat{s}_{sd}(t,n) = \arg\{\min_{s_k \in S} |\Lambda - \sqrt{P_1}|h_{sd}|^2 s_k|^2\}$$

where $t \in 1, 2$ in the case of CL-EO-STBC and $t \in 1, 2, 3, 4$ in the case of CL-QO-STBC, and

• In the case of CL-EO-STBC

$$\Lambda = \begin{cases} h_{sd}^* r_{sd}(1,n) & \text{if } t = 1\\ -h_{sd} r_{sd}^*(2,n) & \text{if } t = 2 \end{cases}$$

• In the case of CL-QO-STBC

$$\Lambda = \begin{cases} h_{sd}^* r_{sd}(1,n) & \text{if} \quad t = 1,4 \\ -h_{sd} r_{sd}^*(2,n) & \text{if} \quad t = 2,3 \end{cases}$$

Therefore, the detection result of the DT link is next used to initialize

$$\mathbf{s}^{(0)}(n) = [s^{(0)}(1,n), ..., s^{(0)}(t,n)]^T = [\hat{s}_{sd}(1,n), ..., \hat{s}_{sd}(t,n)]^T$$

- 4. Set iteration q = 1, 2, ..., N
- 5. Remove additional ISI from the received signal $r_{rd}(t, n)$, where $t \in 2$ in the case of CL-EO-STBC and $t \in 2, 3, 4$ in the case of CL-QO-STBC,
 - In the case of CL-EO-STBC

$$\mathbf{r}_{rd}^{\prime(q)}(n) = \begin{bmatrix} r_{rd}^{\prime(0)}(1,n) \\ r_{rd}^{*}(2,n) - I_{int}^{*(q-1)}(2,n) \end{bmatrix}$$

• In the case of CL-QO-STBC

$$\mathbf{r}_{rd}^{\prime(q)}(n) = \begin{bmatrix} r_{rd}^{\prime(0)}(1,n) \\ r_{rd}^{*}(2,n) - I_{int}^{*(q-1)}(2,n) \\ r_{rd}^{*}(3,n) - I_{int}^{*(q-1)}(3,n) \\ r_{rd}(4,n) - I_{int}^{(q-1)}(4,n) \end{bmatrix}$$

where $I^{(q-1)}(2,n)$ in the case of distributed CL-EO-STBC is

$$I^{(q-1)}(2,n) = \sqrt{\frac{P_1 P_2}{P_1 + 1}} ((g_2(-1)f_2 s^{(q-1)}(1,n) + (g_3(-1)U_2 f_3^* + g_4(-1)f_4^*)s^{(q-1)}(2,n))$$

and $I^{(q-1)}(t,n),\,t\in 2,3,4$ in the case of distributed CL-QO-STBC is

$$I_{int}(2,n) = \sqrt{\frac{P_1 P_2}{P_1 + 1}} (U_1 g_{12}^{(-1)} f_{12}^* s^{(q-1)}(3,n) + U_2 g_{22}^{(-1)})$$

$$f_{22} s^{(q-1)}(4,n)$$

$$I_{int}(3,n) = \sqrt{\frac{P_1 P_2}{P_1 + 1}} (-U_1 g_{12}^{(-1)} f_{12}^* s^{*(q-1)}(4,n) + U_2 g_{22}^{(-1)})$$

$$f_{22} s^{*(q-1)}(3,n)$$

$$I_{int}(4,n) = \sqrt{\frac{P_1 P_2}{P_1 + 1}} (U_1 g_{12}^{(-1)} f_{12}^* s^{*(q-1)}(1,n) + U_2 g_{22}^{(-1)})$$

$$f_{22} s^{*(q-1)}(2,n)$$

6. Substitute $\mathbf{r}_{rd}(n)$ in (3.2.17) with $\mathbf{r}_{rd}^{'(q)}(n)$ to detect the $\hat{\mathbf{s}}^{(q)}(n)$

7. Repeat the process from point 4 until q = N

As will be seen in the simulation results, the above approach shows that, the PIC detection can mitigate the impact of imperfect synchronization among the relay nodes at the destination node. In distributed CL-EO- STBC and distributed CL-QO-STBC, the performance of cooperative diversity gain can be achieved in 2-3 iterations of the PIC detection process as demonstrated in the simulation results.

3.4 Pairwise Error Probability Analysis

Pairwise error probability (PEP) is one of the common performance measure metrics used in evaluating wireless communication systems as presented in Section 2.3. It is simply referred to as the probability of detecting one symbol when another symbol is transmitted. Therefore, the system with the minimum PEP can be considered as the best system performance. In this section, the PEP of distributed STBC is analyzed by employing the distributed CL-EO-STBC and distributed CL-QO-STBC assuming perfect synchronization among the relay nodes. Following the approach in [27] and assuming the destination node knows the channel coefficients μ_k and ψ_k , where

$$\mu_k = \begin{cases} f_k & \text{if } \text{Using CL-EO-STBC} \\ \\ \\ f_{ik} & \text{if } \text{Using CL-QO-STBC}, \end{cases}$$

and

where $k \in 1, 2, 3, 4$ in the case of using CL-EO-STBC and $i, k \in 1, 2$ in the case of using CL-QO-STBC, the PEP for D-CL-STBC has the following Chernoff bound

$$P_e(\mathbf{s}_k(n) \to \hat{\mathbf{s}}_k(n)) \le \mathbf{E}_{\mu_k,\psi_k} \exp^{-\left(\hat{P}(\mathbf{h}_c^H \mathbf{S}_e^H \mathbf{S}_e \mathbf{h}_c + \zeta)\right)} (3.4.1)$$

where $\mathbf{E}(.)$ represents the statistical expectation operation with respect to the channel coefficients μ_k and ψ_k , and ζ^3 can be represented as follows

$$\zeta = \begin{cases} 0 & \text{if } \text{Using CL-QO-STBC} \\ \\ \lambda_f \mathbf{s}_e^H \mathbf{s}_e & \text{if } \text{Using CL-EO-STBC}, \end{cases}$$

where, $\mathbf{s}_e = \mathbf{s}_k(n) - \hat{\mathbf{s}}_k(n)$, $\mathbf{S}_e = \mathbf{S}_k(n) - \hat{\mathbf{S}}_k(n)$, $\mathbf{h}_c = [\mu_1 \psi_1, \mu_2^* \psi_2, U_1 \mu_3^* \psi_3, U_2 \mu_4 \psi_4]^T$ in the case of using CL-QO-STBC and $\mathbf{h}_c = [U_1 \mu_1 \psi_1, \mu_2 \psi_2, U_2 \mu_3^* \psi_3, U_2 \mu_4^* \psi_4]^{T4}$, and $\hat{P} = \frac{P_1 P_2}{4(1+P_1+P_2g)}$ with $g = (\sum_{k \in 2,4} |\psi_k|^2 + |U_1 \psi_1|^2 + |U_2 \psi_3|^2)$ in the case of using CL-EO-STBC and $g = (\sum_{k \in 1,2} |\psi_k|^2 + |U_1 \psi_3|^2 + |U_2 \psi_4|^2)$ in the case of using CL-QO-STBC. Since $|U_1|^2 = |U_2|^2 = 1$, then the channel gain for the cooperative relay nodes becomes $g = \sum_{k=1}^4 |\psi_k|^2$. It is important to note that $\mathbf{S}_k(n)$ and $\hat{\mathbf{S}}_k(n)$ ($\mathbf{S}_k(n) \neq \hat{\mathbf{S}}_k(n)$) are four possible codewords of QO-STBC and two possible codewords of EO-STBC. Next, integrating over μ_k as in [27] yields

$$P_e(\mathbf{s}_k(n) \to \hat{\mathbf{s}}_k(n)) \le \mathbf{E}_{\psi_k} \exp^{-(\hat{P}\zeta)} \det^{-1} \left[\mathbf{I}_4 + \hat{P} \mathbf{S}_e^H \mathbf{S}_e D \right] \qquad (3.4.2)$$

With the optimum power allocation in [27], when $P \gg 1$ and the number of transmitting antennas on each relay node R is large g =

 $^{^{3}\}zeta$ denotes the array gain which can be achieved by using CL-EO-STBC.

⁴Where in general $\mathbf{h}_c = [\hat{\mu}_1 \psi_1, ..., \hat{\mu}_R \psi_R]^T$ and the coefficient $\hat{\mu}_k$ can be calculated as $\hat{\mu}_k = \mu_k$ if $B_k = 0$ and $\hat{\mu}_k = \mu_k^*$ if $A_k = 0, k \in [1, 2, 3, 4]$ in the case of using both CL-QO-STBC and CL-EO-STBC.

 $\sum_{k=1}^{R} |\psi_k|^2 \approx R$, where R = 4 in the case of both distributed CL-EO-STBC and distributed CL-QO-STBC, then (3.4.2) becomes

$$P_e(\mathbf{s}_k(n) \to \hat{\mathbf{s}}_k(n)) \le \mathbf{E}_{\psi_k} \exp^{-\left(\frac{P}{16R}\zeta\right)} \det^{-1} \left[\mathbf{I}_R + \frac{P}{16R} \mathbf{S}_e^H \mathbf{S}_e D\right] 3.4.3)$$

where $D = diag\{|\psi_1|^2, |psi_2|^2, |psi_3|^2, |psi_4|^2\}$ and define $\mathbf{B} = \mathbf{S}_e^H \mathbf{S}_e$. The upper bound for the PEP using the minimum nonzero singular value of \mathbf{B} , which is denoted as σ_{min}^2 , from (3.4.3) becomes

$$P_{e}(\mathbf{s}_{k}(n) \rightarrow \hat{\mathbf{s}}_{k}(n)) = \mathbf{E}_{\psi_{k}} \exp^{-\left(\frac{P}{64}(\lambda_{f}\mathbf{s}_{e}^{H}\mathbf{s}_{e})\right)}$$
$$\prod_{k=1}^{\operatorname{rank} \mathbf{B}} \left(1 + \frac{P\sigma_{min}^{2}}{64}|\psi_{k}|^{2}\right)$$
$$= \exp^{-\left(\frac{P}{64}(\lambda_{f}\mathbf{s}_{e}^{H}\mathbf{s}_{e})\right)} \left(\frac{P\sigma_{min}^{2}}{64}\right)^{-\operatorname{rank} \mathbf{B}}$$
$$\left[-\exp^{\frac{64}{P\sigma_{min}^{2}}}\mathbf{Ei}\left(-\frac{64}{P\sigma_{min}^{2}}\right)\right]^{\operatorname{rank} \mathbf{B}} (3.4.4)$$

where

$$\mathbf{Ei} = \int_{-\infty}^{\chi} \frac{e^t}{t} dt, \qquad \chi < 0$$

is the exponential integral function and $R_{\mathbf{B}}$ is the rank of \mathbf{B}

$$P_{e}(\mathbf{s}_{k}(n) \to \hat{\mathbf{s}}_{k}(n)) \lesssim \exp^{-\left(\frac{P}{64}(\lambda_{f}\mathbf{s}_{e}^{H}\mathbf{s}_{e})\right)} \left(\frac{64}{P\sigma_{min}^{2}}\right)$$
$$P^{-\operatorname{rank} \mathbf{B}\left(1 - \frac{\log\log P}{\log P}\right)}$$
(3.4.5)

which yields the same diversity as in [27], therefore when **B** is full rank, the diversity gain is $4\left(1 - \frac{\log \log P}{\log P}\right)^5$ in both cases with a smaller PEP due to the array gain λ_f in the case of distributed CL-EO-STBC only.

⁵When P is very large (log $P \gg \log \log P$), the diversity gain is approximately equal to the number of cooperating relay nodes R.

3.5 Simulation Results

In this section, the simulation results shows how the synchronization error can cause a degradation in the system's end-to-end BER performance and how it can be mitigated. In all figures, the horizontal axis indicates the total power used in the network, while the vertical axis indicates end-to-end BER. Furthermore, all simulations which are presented in this section are simulated using the QPSK mapping scheme. In order to show the BER performance, it is assumed that the channels from the source node to the relay nodes and from the relay nodes to the destination node are both quasi-static flat fading channels.

The simulated performance of distributed STBC with outer coding and



Figure 3.5. The end-to-end BER performance comparisons of proposed distributed STBC with and without outer coding schemes.

the comparison between distributed A-STBC in [50] and distributed

CL-EO-STBC is demonstrated in Figure 3.5 where it can be observed that the end-to-end BER performance of the relaying system based on distributed STBC without outer convolutive performs worse when compared to end-to-end BER performance of distributed STBC with outer convolutive coding. The performance of the proposed distributed CL-EO-STBC with outer convolutive coding provides approximately 3 dB of gain at BER 10^{-3} as compared with distributed A-STBC with outer convolutive coding. When the total power P is in dB, the coding gain is defined as $C_g = (P)_{Uncoded} - (P)_{Coded}$. Approximately $C_g=25$ -16.5=8.5 at BER 10^{-3} in the case of distributed CL-EO-STBC and approximately $C_g=28$ -19.5=8.5 at BER 10^{-3} in the case of distributed A-STBC.



Figure 3.6. The end-to-end BER performance of distributed CL-EO-STBC with outer convolutive coding utilizing ML detection under imperfect synchronization $\beta_k = 3, 0, -3, and -6$ dB.

Figure 3.6 shows the results of ML detection with different time misalignment among the relay nodes at the destination node for the CL-EO-STBC scheme, including the performance of the scheme under perfect synchronization (PS) as a reference. It can be observed from the figure and Table 3.1 that the ML detection does not achieve the available cooperative diversity gain and cannot deliver a good end-to-end performance under imperfect synchronization errors even under small time misalignment $\beta_k = -6$ dB, $k \in 2, 3, 4$.

Table 3.1. End-to-end BER comparison between ML detection and PIC detection for distributed CL-EO-STBC under imperfect synchronization.

Distributed			ML	number	PIC
STBC	β_k (dB)	BER	Power (dB)	of	Power (dB)
scheme				iterations	
CL-EO-STBC	0	10^{-2}	None	3	~ 16
	-6	10^{-2}	None	3	~ 14

Figure 3.7 shows the BER of PIC detection with distributed CL-EO-STBC under imperfect synchronization when $\beta_k = 3, 0, -3, and - 6$ dB and when the PIC number of iterations is q = 3. It can be observed that the PIC detection is an effective approach to improve the performance of distributed STBC and mitigates the ISI and also improves its efficiency over large amounts of time misalignments.

For example as shown in Table 3.1, the value of 10^{-2} BER cannot be achieved by ML detection in the proposed scheme even under small time misalignment $\beta_k = -6$ dB, whereas with the PIC detector involved, it just required approximately 14 dB of gain to achieve the value of 10^{-2} BER when $\beta_k = -6$ dB and q = 3.



Furthermore, as it can be observed in Figure 3.7, even under large time

Figure 3.7. The end-to-end BER performance of distributed CL-EO-STBC with outer convolutive coding utilizing PIC detection under different time error $\beta_k = 3, 0, -3, -6$ dB and when the iteration q = 3.

misalignment $\beta_k = 3$ dB and q=3, it required approximately 18 dB of gain to achieve 10^{-2} as compared to DT link performance, which required approximately 22 dB to achieve 10^{-2} BER. Similarly, simulation results for the CL-QO-STBC scheme under imperfect synchronization is depicted. In Figure 3.8, the results of conventional (synchronized) CL-QO-STBC and the impact of DT link on this scheme is shown. The figure shows that the conventional CL-QO-STBC with DT link provides an improved performance over the conventional CL-QO-STBC without DT link in terms of power saving. For example, at a BER of 10^{-3} the scheme with DT link provides a power saving of approximately 5 dBcompared to the CL-QO-STBC without DT link.



Figure 3.9 shows the results of conventional detector under different

Figure 3.8. The end-to-end BER performance of synchronized CL-QO-STBC.

values of time misalignment (τ_k) calculated as per Table 3.3 below. It can be observed from the figure and Table 3.2 that the ML detection does not achieve the available cooperative diversity gain and cannot deliver a good end-to-end performance under imperfect synchronization errors even under small time misalignment $\tau_k = 0.125T$. On the other hand, Figure 3.10 shows the effect of the PIC detector to synchronization error even under large time misalignment $\tau_k = 0.5T$ where the number of PIC iterations (q) implemented are only 3 iterations.

For example as shown in Table 3.2, the value of 10^{-2} BER cannot be achieved by ML detection in the proposed scheme even under small time misalignment $\tau_k = 0.125T$ dB, whereas with the PIC detector involved, it just required approximately 14 dB of gain to achieve the



Figure 3.9. The end-to-end BER performance of asynchronous CL-QO-STBC under different time misalignment τ_k values.

Table 3.2. End-to-end BER comparison between ML detection and PIC detection for distributed CL-QO-STBC under imperfect synchronization.

Distributed			ML	number	PIC
STBC	$\tau_k (\mathrm{dB})$	BER	Power (dB)	of	Power (dB)
scheme				iterations	
CL OO STRC	$0.5\mathrm{T}$	10^{-2}	None	3	~ 16
	0.125T	10^{-2}	None	3	~ 14

value of 10^{-2} BER when $\tau_k = 0.5T$ dB and q = 3.



Figure 3.10. The end-to-end BER performance of asynchronous CL-QO-STBC under different time misalignment τ_k values.

Table 3.3. Shows the calculation of the impact of time delay between relay node R_1 and the other relay nodes.

β	$dB = 10\log_{10}(\beta_k)$	$ au_k$
1	0	0.5T
0.5	-3	0.25T
0.3	-5	0.15T
0.25	-6	0.125T

3.6 Summary

This chapter investigated applying distributed STBC in [27], and outer convolutive coding for a relay network with two dual-antenna relay nodes and four single-antenna relay nodes by utilizing distributed CL-EO-STBC and CL-QO-STBC respectively. The impact of imperfect synchronization among the antennas of relay nodes at the destination node was considered. To overcome this issue a signal detector was proposed based on the principle of PIC detection to cancel the ISI caused by imperfect synchronization and achieve full diversity order with coding gain and full data transmission rate in each hop for both proposed schemes, even without channel knowledge at the relay nodes. The implementation of feedback scheme with distributed EO-STBC and QO-STBC ensured diversity gain of the order of the number of transmitting antennas on the cooperative relay nodes in addition of achieving array gain with full data transmission rate between the relay nodes and the destination node. Furthermore, the PEP utilizing distributed CL-EO-STBC and CL-QO-STBC for asynchronous cooperative relay network was analyzed to confirm that the cooperative diversity gain is equal to the number of transmitting antennas on the relay nodes in the case of distributed CL-QO-STBC and with a smaller PEP due to array gain in the case of distributed CL-EO-STBC. This result was based on the destination node having full knowledge of all fading channels. Finally, the simulation results of the PIC approach with both proposed schemes showed that there was significant performance improvement over ML detection under synchronization error and the performance close to the perfect synchronized case was achieved in just three iterations, whereas ML detection fails to mitigate the impact of imperfect synchronization even under small time misalignment. Although both distributed STBC schemes showed similar results, however, in the case of CL-QO-STBC a significant amount of complexity was reduced at the relay nodes in terms of time misalignment calculations due to the dual-antenna system model incorporated. With the complexity of the number of PIC iterations, the next chapter will focus on analyzing asynchronous cooperative relay networks and proposing two novel detection schemes with low detection complexity which is dependent only on the constellation size.

NEAR-OPTIMUM DETECTION SCHEME WITH RELAY SELECTION TECHNIQUE FOR ASYNCHRONOUS COOPERATIVE RELAY NETWORKS

In this chapter, a novel detection scheme for decode-and-forward (DF) and amplify-and-forward (AF) asynchronous cooperative relay networks is proposed utilizing distributed closed-loop extended orthogonal space time block coding (CL-EO-STBC). These techniques are both designed to effectively remove the interference at the destination node induced by different time delays from the antennas of each relay node and achieve full cooperative diversity gain with unity data transmission rate between the relay nodes and the destination node. Moreover, a simple max-min relay selection scheme is proposed for cooperative relay networks to enhance the system performance. The best two relays are selected based on the overall path gain and the smallest timing error, then the CL-EO-STBC in a distributed manner is applied over the selected two relay nodes.

4.1 Introduction

The most existing research on cooperative transmission assume perfect synchronization among cooperative users, which means that the users' timing, carrier frequency and propagation delay are identical [51]. Under this assumption, there is in fact little diversity difference between cooperative STBC and traditional STBC except certain cooperation overhead. Unfortunately, it is difficult, and in most cases impossible, to achieve perfect synchronization among distributed transmitters. This is even more a reality when low-cost, small-sized transmitters are used. The effect of asynchronism among cooperative relay nodes might lead to channel dispersion which will damage the orthogonality of the STBC causing significant degradation in the overall system performance. The issue of asynchronism in D-STBC among cooperating nodes has been recently addressed in [35] [106] utilizing an equalization technique at the destination. This technique increases the overhead at the receiver. An alternative method is using the near-optimum detection scheme in [48] which effectively mitigates the asynchronism among relay nodes with simple near-Alamouti decoding, but suffers from the limitation that the number of cooperating relay nodes must not exceed two, thereby limiting the diversity order. In contrast, the approach introduced in [92] for mitigating the effect of imperfect synchronization achieves fourth order diversity by utilizing a parallel interface cancelation (PIC) detection scheme for four relay nodes. This approach, however, does not achieve full data rate between the relay nodes and the destination node because a complex Distributed orthogonal STBC (D-OSTBC) with 3/4 data rate is utilized; moreover the computational complexity of the PIC iteration process required for overcoming the effects of interference is significant.

The sub-optimum detection scheme in [96] was proposed for the case of four relay nodes to mitigate asynchronism effects utilizing the CL-EO-STBC with phase rotated feedback as in [30]. This scheme canceled the interference components caused by asynchronism effectively with near-Alamouti simplicity, unity rate code and achieved the full diversity order between the relay nodes and the destination node. Furthermore, the complexity of this scheme is only related to the size of the utilized signal constellation. However, this detection scheme relied mainly on the existence of a direct transmission (DT) link between the source node and the destination node which is unrealistic and can be difficult to achieve in practice. Furthermore, the work in [35] - [96] dealt with the timing error problem in cooperative relay networks under the assumption that the relaying process operates with the DF type of communication protocol. Whereas [95] adopted the AF type of transmission utilizing the iterative PIC detection scheme to mitigate the impact of asynchronism in addition to the utilization of a two-bit feedback scheme in [30] and assumes that a DT link must exist for a successful detection. In this chapter, a near-optimum detection scheme is proposed for two dual-antenna cooperative relay nodes selected from

a set of candidate relays and utilizing both the DF and AF cooperation strategies. Outer convolutional coding is employed to further improve the coding gain. The approach eliminates the interference components at the destination due to time misalignment caused by the asynchronism in the transmission from the antennas of the relay nodes with the equivalent simplicity of decoding as in [48] and [96], but with the advantage of overcoming their drawbacks. In particular of full diversity order of four from the cooperating relay nodes to the destination node is achieved as well as completely dispensing with the DT link assumed in [96] and hence reducing the detection complexity at the destination node and modeling a more realistic wireless cooperative transmission scenario. Moreover, the AF type of cooperating strategy adopted in the scheme reduces the complexity imposed in the cooperative relay nodes as well as saving power and time as compared to the DF strategy. Moreover, the non-iterative detection algorithm avoids undesirable complexity at the destination node due to the utilization of a PIC detection algorithm as in [95] and achieves a near-optimum performance with a significant reduction of feedback amount compared to the proposed scheme in [30]. In addition, the relay selection technique utilized in this work selects the best channel quality as well as the channel with the minimum time misalignment.

4.2 System Model

The cooperative system model comprises of a single-antenna source and destination node and R_k dual-antennas relay nodes, where $k \in 1, ..., N_R$, and N_R denotes the number of relay nodes in the network as depicted in Figure 4.1. There is no DT connection between the source node and



Figure 4.1. Basic structure of general cooperative relay network with one source, N_R relay nodes and one destination node; each relay node is equipped with two antennas with outer convolutive coding.

the destination node because it is assumed that the signal through the DT link fails to reach the destination node due to pathloss effects [25]. Therefore, the destination node relies only on the signal from the relay nodes. Furthermore, it is assumed that the separation between antennas is sufficient to have the required independent branches for space-diversity application [107]. It is also assumed that the source-relays and the relays-destination channel coefficients are estimated perfectly at the destination ¹. Let f_{ik} denote the channel coefficients from the source node to *i*-th antenna of the *k*-th relay node and g_{ik} is the channel coefficients from the *i*-th antenna of the *k*-th relay node to the destination node and that all channel coefficients f_{ik} and g_{ik} are constant during the transmission of a signal code block ². The wireless network is assumed to transmit in an outdoor flat-fading environment

 $^{^1 {\}rm In}$ practice these are estimated by training sequences which introduce CSI error but considering its effect is beyond the scope of this thesis

²That is a quasi-static channel

where there is insignificant multipath propagation; however, without perfect synchronization, channels become dispersive even in flat fading environments. Due to the transmitting/receiving pulse shaping filters, if the sampling time instants are not ideal, intersymbol interference (ISI) is introduced. This certainly brings performance degradation or performance loss in cooperative STBC. More important, asynchronism among the transmitters may break the orthogonal STBC signal structure, which makes most of the existing STBC decoders fail due to performance loss rather than simply complexity increase. Coding gain is exploited in this model by combining outer convolutive coding at the source node with a Viterbi decoder at the destination node as shown in Figure 4.1. All relay nodes are assumed to operate in half-duplex mode, which means all information transmission from the source node to destination node occurs in two phases. In the first phase, the information sequence $\mathbf{s}(n) = [s(1,n), s(2,n)]^T$, which is encoded by the convolutional encoder is then passed through the interleaver and mapped into QPSK symbols, where n denotes the discrete pair index. Then the source node broadcasts them to each antenna of each relay node R_k , $k \in 1, ..., N_R$, in two different time transmission periods. In the second phase, the DF or AF cooperation strategies are used by the relay nodes. Then the received signals at each relay node are processed using the distributed CL-EO-STBC technique and components of the matrix code are effectively transmitted from each antenna of each relay node to the destination node in two different time transmission periods. The cooperating relays are selected according to certain criteria in order to enhance the overall system performance. In the following section, These criteria are discussed.

4.3 Relay Selection Technique for Two Dual-Antenna Relay Nodes

As mentioned previously, in a cooperative relay network, the cooperative relay nodes can assist the source node to broadcast the signals to the destination node, however the cooperative relay nodes have different locations so each transmitted signal from the source node to the destination node must pass through distinct paths causing different attenuations within the signals received at the destination node which result in reducing the overall system performance. Therefore, to overcome this effect and benefit from cooperative communication, certain paths should be avoided by using selection techniques [108]. In this section, the conventional relay selection technique based on maximizing the minimum instantanous SNR is used to ensure that the relay node with the best end-to-end paths between the source node and the destination is used, as shown in Figure 4.2(a), to improve the system performance and provide diversity gain on the order of the number of transmitting antenna at the relay nodes equal to $2N_{R_s}$. As mentioned above each relay node is equipped with two antennas to assist the source node to transmit its signals to the destination node. The resulting signal-to-noise ratio SNR_D at the destination node assuming maximum ratio combining is given by

$$SNR_D = \sum_{k \in N_R} \sum_{i \in 1,2} \frac{SNR_{SR_{ik}}SNR_{R_{ik}D}}{1 + SNR_{SR_{ik}} + SNR_{R_{ik}D}}$$
(4.3.1)

where N_R denotes the set of relay indices for the relay nodes chosen in the multi-relay selection scheme. $SNR_{SR_{ik}} = |f_{ik}|^2 \frac{\sigma_s^2}{\sigma_n^2}$ is the instantaneous SNR of the paths between the source node node and the relay node antennas and $SNR_{R_{ik}D} = |g_{iik}|^2 \frac{\sigma_s^2}{\sigma_n^2}$ is the instantaneous SNR of



Figure 4.2. Source transmits to destination and neighboring nodes overhear the communication. (a) A set of relays with the best end-toend path among N_R candidates is selected to relay information, via a distributed mechanism based on instantaneous channel measurements. (b) The relay with the best end-to-end path and the relay with the minimum time misalignment are selected assuming that the relay node with best SNR is synchronized to the destination node.

the paths between the relay node antennas and the destination node, where $k \in 1, ..., N_R$ and $i \in 1, 2$. These expressions give the instantaneous signal strength SNR between the source node and the relay node antennas, and the relay node antennas and the destination node respectively. Then the selected relay node is chosen to maximize the minimum between them. The two antenna relay selection policy can be expressed as

$$\Omega_{1k} = \min \{ SNR_{SR_{1k}}, SNR_{R_{1k}D} \}$$

$$\Omega_{2k} = \min \{ SNR_{SR_{2k}}, SNR_{R_{2k}D} \} \quad \text{for} \quad k \in 1, ..., N_R (4.3.2)$$

and

$$\mathcal{R} = \max_{k \in 1, \dots, N_R} \left\{ \min\{\Omega_{1k}, \Omega_{2k}\} \right\}$$
(4.3.3)

After the best relay node has been chosen as shown in Figure 4.2(a), then it can be used to forward the received signals toward the destination node. In this work, only two relay nodes in which each relay node is equipped with two antennas are selected to perform an appropriate encoding process to generate distributed STBC at the destination node. The relay selection technique applied in this work is also utilized to minimize the time misalignment among the set of selected relay nodes in addition to selecting the best relay node with the smallest time misalignment as shown in Figure 4.2(b). The procedure to perform this operation based on channel state information (CSI) at the destination node is as follows

- Denote the set of available relay nodes by N_R each of which has two antennas.
- Select the best group of relay nodes N_{R_s} from the available set of relay nodes N_R using the max-min selection scheme in (4.3.3).
- Without loss of generality, the destination node is assumed to be synchronized with the best selected relay node i.e. τ₁ = τ₁₁ = τ₂₁ = 0 as shown in Figure 4.2(b).
- Without loss of generality, the time delays from each antenna of each remaining relay nodes N_{R_s-1} are assumed to be the same, because both antennas are co-located in the same relay node τ_k = τ_{1k} = τ_{2k} ≠ 0, k ∈ 2, ..., N_{R_s-1}.
- At this point the selection process is repeated among the remaining set of the best relay nodes N_{R_s-1} where the relay nodes with

minimum relative interference strength β are selected such that

$$\beta(l) = \min\left\{\beta_2, \beta_3, \dots, \beta_{N_{R_s-1}}\right\} \quad \text{for} \quad l \in -1, -2, -3... \quad (4.3.4)$$

The relative interference strength β is normally $\beta = 1$ (i.e. 0 dB) for $\tau_k = 0.5T$ and $\beta = 0.25$ (i.e. -6 dB) for $\tau_k = 0.125T$ [8].

• Finally, the best relay node chosen above and the relay node with the smallest time delay error are used to forward the received signals toward the destination node.

After selecting the best relay node and the relay node with the smallest time delay error, the EO-STBC is distributed over these two dualantenna relay nodes during the second broadcasting phase. In the following section, the distribution is explained first for both the DF and AF types of the cooperative relaying.

4.3.1 Distributed CL-EO-STBC for Two Dual-Antenna Relay Nodes

Utilizing the DF Type of Transmission

The distributed open loop EO-STBC can achieve full data rate at the expense of losing some diversity order between the relay nodes and the destination node due to the interference factor between the estimated symbols. Therefore, in this section, a CL-EO-STBC is considered for two dual-antenna relay nodes as shown in Figure 4.3. The application of a feedback scheme, [30], to the relay network will ensure full data-rate together with full diversity order equal to the number of employed antennas on the transmitting relay nodes. The phases of the different code symbols which are transmitted from the first antennas at both relay nodes must be rotated by appropriate phase angles U_1 and U_2 ,



Figure 4.3. Basic structure of distributed CL-EO-STBC with outer coding using two-bit feedback based on phase rotation for an asynchronous wireless relay with two antennas in each relay node and one antenna in the source and the destination node with two phases for the cooperative transmission process and time delay offset between the antennas of R_2 and the destination node.

respectively, before they are transmitted, while the symbols from the second antennas of each relay node are kept unchanged as in (4.3.5).

$$\begin{bmatrix} U_1 s(1,n) & s(1,n) & U_2 s(2,n) & s(2,n) \\ -U_1 s^*(2,n) & -s^*(2,n) & U_2 s^*(1,n) & s^*(1,n) \end{bmatrix}$$
(4.3.5)

The application of phase rotations to the transmitted symbols is equivalent to phase rotations of the corresponding channel coefficients g_{ik} . This is important to enhance the effective diversity between the relays and the destination at the expense of some feedback overhead.

As assumed previously, the transmitted symbols from the relay nodes will suffer from the asynchronism due to different propagation delays and hence will not arrive simultaneously at the destination node. This asynchronism will induce intersymbol interference (ISI) at the destination and will damage the orthogonality of the transmitted codeword as shown in Figure 4.4. It is assumed that both antennas of R_1 are fully



Figure 4.4. Representation of misalignment of received signals at the destination node which induces ISI in the case of two relay nodes each relay node is equipped with two antennas.

synchronized to the destination node, that is, $\tau_{i1} = 0$, $i \in 1, 2$. Therefore, the received signal at the destination node via the relay nodes in two different time transmission periods due to time synchronization error between the antennas of each relay node can be expressed as follows

$$r_{rd}(1,n) = (U_1g_{11} + g_{21})s(1,n) + (U_2g_{12} + g_{22})s(2,n) + I_{int}(1,n) + w_{rd}(1,n)$$
(4.3.6)

$$r_{rd}(2,n) = -(U_1g_{11} + g_{21})s^*(2,n) + (U_2g_{12} + g_{22})s^*(1,n) + I_{int}(2,n) + w_{rd}(2,n)$$
(4.3.7)

where $I_{int}(1, n)$ and $I_{int}(2, n)$ are the interference terms from both antennas of relay node R_2 in two different time transmission periods and are expressed as:

$$I_{int}(1,n) = (U_2g_{12}(-1) + g_{22}(-1))s^*(1,n-1) + (U_2g_{12}(-2) + g_{22}(-2))s(2,n-1)$$
(4.3.8)

$$I_{int}(2,n) = (U_2g_{12}(-1) + g_{22}(-1))s(2,n) + (U_2g_{12}(-2) + g_{22}(-2))s^*(1,n-1)$$
(4.3.9)

and $w_{rd}(1, n)$ and $w_{rd}(2, n)$ represent additive Gaussian noise with zeromean and unity variance at the destination node and g_{i1} and g_{i2} , $i \in$ 1, 2, denote complex channel coefficients from the antennas of each relay node and the destination node. As shown in Figure 4.4, the effect of ISI from the previous symbols is represented by $g_{i2}(l)$, $i \in 1, 2$, and $l \in -1, -2$. It is noted that that $g_{i2}(-2)$ is generally a much smaller coefficient [48], [9], therefore, the strengths of $g_{i2}(l)$ can be expressed as a ratio as

$$\beta_{i2}(l) = \frac{|g_{i2}(l)|^2}{|g_{i2}|^2}$$
 for $i = 1, 2$ and $l = -1, -2$ (4.3.10)

where $\beta_{12}(l) = \beta_{22}(l) = \beta(l)$ denotes a sample of a pulse shaping waveform and the effect of the time delay τ_{i2} , $i \in [1, 2]$ [48] [9], between transmission from the antennas of the relay node R_1 and the antennas of the relay node R_2 at the destination node.

The received signals $r_{rd}(1,n)$ and $r^*_{rd}(2,n)$ conjugated for convenience

in (4.3.6) and (4.3.7) can be represented in vector form as

$$\mathbf{r}_{rd}(n) = \mathbf{Hs}(n) + \mathbf{I}_{int}(n) + \mathbf{w}_{rd}(n)$$
(4.3.11)

where

$$\mathbf{H} = \begin{bmatrix} U_1 g_{11} + g_{21} & U_2 g_{12} + g_{22} \\ U_2^* g_{12}^* + g_{22}^* & -U_1^* g_{11}^* - g_{21}^* \end{bmatrix}, \quad (4.3.12)$$

 $\mathbf{w}_{rd}(n) = [w_{rd}(1,n), w_{rd}^*(2,n)]^T$ is an additive Gaussian noise vector, with elements having distribution $CN(0, \sigma_w^2)$, at the destination node and $\mathbf{I}_{int}(n) = [I_{int}(1,n), I_{int}^*(2,n)]^T$ is the interference vector containing terms from the antennas of R_2 at the destination node, which can be modelled as in (4.3.8) and (4.3.9). The feedback terms $U_1 = e^{j\theta_1}$ and $U_2 = e^{j\theta_2}$ are determined by two feedback information angles θ_1 and θ_2 which are obtained by maximizing the array gain. The design criterion of the two-bit feedback scheme in [30] is applied ensuring that each element of the feedback performance gain should be real and positive and can be achieved by:

$$\theta_1 = -\angle (g_{11}g_{21}^*)$$

$$\theta_2 = -\angle (g_{12}g_{22}^*)$$
(4.3.13)

The exact values of the phase angles in (4.3.13) are real valued and would require infinite precision for perfect precision which in practice it is not possible to use due to the limited feedback bandwidth. Hence, these angles should be quantized first before they get fed back to the transmitting antennas. For each phase angle, a two-bit feedback requires four phase level angles to be chosen from the set of $\{\theta_1, \theta_2\} \in \psi$
= $[0, \pi/2, \pi \text{ or } 3\pi/2]$. The selection of the discrete feedback information that corresponds to the phase adjustment of the first antennas of each of the two relays is performed according to

$$\theta_1 = \arg \max_{\theta_1 \in \psi} \Re \left\{ (g_{11}g_{21}^*)e^{j\theta_1} \right\}$$

$$(4.3.14)$$

and

$$\theta_2 = \arg \max_{\theta_2 \in \psi} \Re \left\{ (g_{12} g_{22}^*) e^{j\theta_2} \right\}$$

$$(4.3.15)$$

The selection that gives the largest values of (4.3.14) and (4.3.15) will provide the largest array gain and achieve full diversity advantage. At this stage, the received signal at the destination node is ready to be processed, however in the following section, the transmission process is explained for the AF type of relaying transmission.

4.3.2 Distributed CL-EO-STBC for Two Dual-Antenna Relay Nodes Utilizing the AF Type of Transmission

Most cooperative wireless relay networks communicate through a twophase communication process as mentioned previously. In the first phase, the source node broadcasts the mapped QPSK symbols, after encoding and interleaving process, which are grouped as $\mathbf{s}(n) =$ $[s(1,n)), s(2,n)]^T$ where n denotes the discrete index as shown in Figure 4.1. When the information vector $\mathbf{s}(n)$ is transmitted, the received signal vector at the *i*-th antenna of the k-th relay node is given by

$$\mathbf{r}_{ik}(n) = \sqrt{P_1} f_{ik} \mathbf{s}(n) + \mathbf{v}_{ik}(n), \quad i = 1, 2 \text{ and } k = 1, 2, ..., N_R \quad (4.3.16)$$

where P_1 is the average transmit power at the source node for every channel use and $\mathbf{v}_{ik}(n)$ is the additive Gaussian noise vector at each antenna of each relay node with zero-mean and unit-variance. In the second phase, all the relay nodes are scheduled to process and transmit the received noisy signals $\mathbf{r}_{ik}(n)$ to the destination node simultaneously. In general, the transmitted signals from different antennas at each relay node is designed to be a linear function of the received signal and its conjugate and can be written as:

$$\mathbf{t}_{ik}(n) = \sqrt{\frac{P_2}{P_1 + 1}} (\mathbf{A}_k \mathbf{r}_{ik}(n) + \mathbf{B}_k \mathbf{r}_{ik}^*(n))$$
(4.3.17)

where $\mathbf{t}_{ik}(n) = [t_{ik}(1,n), t_{ik}(2,n)]^T$, P_2 is the average transmitted power at the relay nodes for every channel use. If the total power per symbol transmission used in the whole network is fixed as P, the optimal power allocation that maximizes the expected receive SNR is

$$P_1 = \frac{P}{2}$$
 and $P_2 = \frac{P}{4N_R}$

where N_R is the number of candidate relays in the network, and each of the relay nodes is equipped with a pair of 2×2 unitary matrices \mathbf{A}_k and \mathbf{B}_k , where $k \in 1, 2$ [28]. Then, the conventional relay selection method in [109] is used to ensure that the relay node with the best end-to-end path between the source node and the destination is used, as shown in Figure 4.2, to improve the system performance and provide diversity gain on the order of the number of transmitting antenna at the relay nodes equal to $2N_{R_s}$. Only two dual-antenna relay nodes are chosen to perform an appropriate encoding process to generate distributed EO-STBC at the destination node. Moreover, the relay selection technique was used to select the relay node with the smallest time misalignment among the set of the best selected relay nodes N_{R_s} as shown in Figure 4.5 [109]. As shown in Figure 4.5, in order to achieve the full cooperative



Figure 4.5. Basic structure of distributed CL-EO-STBC with outer coding using one-bit feedback based on phase rotation for an asynchronous wireless relay with two antennas in each relay node and one antenna in the source and the destination node with two phases for the cooperative transmission process and time delay offset between the antennas of R2 and the destination node.

diversity and full rate between the relay node and destination node, the one-bit feedback EO-STBC is distributed over the two selected relays as follows:

$$\begin{bmatrix} t_{11}(1,n)e^{j\theta_1} & t_{21}(1,n)e^{j\theta_2} & t_{12}(1,n)e^{j\theta_1} & t_{22}(1,n)e^{j\theta_2} \\ t_{11}(1,n)e^{-j\theta_2} & t_{21}(1,n)e^{-j\theta_1} & t_{12}(1,n)e^{-j\theta_2} & t_{22}(1,n)e^{-j\theta_1} \end{bmatrix} (4.3.18)$$

where $t_{ik}(t, n)$ can be represented as in (4.3.16) and $\mathbf{r}_{ik}(n)$ can be represented as in (4.3.17), where i, t, k = 1, 2 and $\mathbf{s}(n) = [s(1, n), -s^*(2, n)]^T$ is the transmitted signal from the source node during two transmission periods, and matrices used in this model for designing the EO-STBC over the relay nodes are

$$\mathbf{A}_{1} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \mathbf{B}_{1} = \mathbf{A}_{2} = 0 \text{ and } \mathbf{B}_{2} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$$
(4.3.19)

Due to the asynchronism the relayed signal will most likely reach the destination node at different time instants, causing ISI at the destination from adjacent symbols. As shown in Figure 4.4, there exists a time misalignment of τ_{i2} , i = 1, 2, between the received copies of the signal, which for convenience are shown identical, due to the fact that both antennas are co-located on the same relay node. Without loss of generality, the relay node R_1 is assumed to be synchronized to the destination node, i.e. $\tau_{11} = \tau_{21} = 0$. Therefore, the received signal at the destination node can be modelled as:

$$\mathbf{r}_{rd}(n) = \sum_{k=1}^{2} \left(g_{1k} \mathbf{t}_{2k} U_1 + g_{2k} \mathbf{t}_{2k} U_2 \right) + \mathbf{I}_{int}(n) + \mathbf{w}_{rd}(n)$$
(4.3.20)

where $\mathbf{w}(n) = [w(1, n), w(2, n)]^T$ is the additive Gaussian noise vector at the destination node, $\mathbf{I}_{int}(n) = [I(1, n), I(2, n)]^T$ is the interference components at the destination node,

$$U_{1} = \begin{bmatrix} e^{j\theta_{1}} & 0\\ 0 & e^{-j\theta_{2}} \end{bmatrix} \text{ and } U_{2} = \begin{bmatrix} e^{j\theta_{2}} & 0\\ 0 & e^{-j\theta_{1}} \end{bmatrix}$$
(4.3.21)

are the one-bit phase rotation matrices applied at the relay nodes. By substituting (4.3.17) into (4.3.20) and taking the conjugate of y(2), the received signal can be rewritten as:

$$r_{rd}(1,n) = \sqrt{\frac{P_2 P_1}{P_1 + 1}} \Big[\left(g_{11} f_{11} e^{j\theta_1} + g_{21} f_{21} e^{j\theta_2} \right) s(1,n) \\ + \left(g_{12} f_{12}^* e^{j\theta_1} + g_{22} f_{22}^* e^{j\theta_2} \right) s(2,n) + I_{int}(1,n) \Big] + w_{rd}(1,n)$$

$$(4.3.22)$$

Similarly,

$$r_{rd}(2,n)^{*} = \sqrt{\frac{P_{2}P_{1}}{P_{1}+1}} \Big[\left(g_{12}^{*}f_{12}e^{j\theta_{2}} + g_{22}^{*}f_{22}e^{j\theta_{1}} \right) s(1,n) \\ - \left(g_{11}^{*}f_{11}^{*}e^{j\theta_{2}} + g_{21}^{*}f_{21}^{*}e^{j\theta_{1}} \right) s(2,n) + I_{int}(2,n) \Big] + w_{rd}^{*}(2,n)$$

$$(4.3.23)$$

where

$$\mathbf{I}_{int}(1,n) = \left(g_{12}(-1)f_{12}^*e^{j\theta_1} + g_{22}(-1)f_{22}^*e^{j\theta_2}\right)s^*(1,n-1) \quad (4.3.24)$$

and

$$\mathbf{I}_{int}(2,n) = \left(g_{12}^*(-1)f_{12}e^{j\theta_2} + g_{22}^*(-1)f_{22}e^{j\theta_1}\right)s^*(2,n) \tag{4.3.25}$$

The coefficient of $g_{i2}(-1)$, which signifies the ISI from the previous symbol under synchronization error, is dependent upon the timing delay $\tau_{i2} \neq 0$, i = 1, 2 and the pulse shaping waveform used. Its relative strength can be represented as in (4.3.20) [95]. The received signals in (4.3.22) and (4.3.23) can be represented as in (4.3.20) where the channel matrix **H** is given by:

$$\begin{bmatrix} g_{11}f_{11}e^{j\theta_1} + g_{21}f_{21}e^{j\theta_2} & g_{12}f_{12}^*e^{j\theta_1} + g_{22}f_{22}^*e^{j\theta_2} \\ g_{12}^*f_{12}e^{j\theta_2} + g_{22}^*f_{22}e^{j\theta_1} & -g_{11}^*f_{11}^*e^{j\theta_2} - g_{21}^*f_{21}^*e^{j\theta_1} \end{bmatrix}$$
(4.3.26)

On the basis of (4.3.11) and (4.3.20), it is well known from estimation theory that the matched filter is the optimum front-end receiver to obtain statistics for detection in the sense that it preserves information. The detailed conventional ML detection for both cases (DF and AF) is next discussed.

4.4 Conventional Distributed EO-STBC for Dual-Antenna Relay Network

The matched filtering is performed at the destination by pre-multiplying (4.3.11) and (4.3.20) by \mathbf{H}^{H} [29]. Therefore, the conventional distributed CL-EO-STBC detection can be carried out as follows

• Step 1: Applying the linear transformation

$$\hat{\mathbf{y}}(n) = [\hat{y}(1,n), \hat{y}(2,n)]^T = \mathbf{H}^H \mathbf{r}_{rd}(n)$$
$$= \Delta \mathbf{s}(n) + \mathbf{H}^H \mathbf{I}_{int}(n) + \mathbf{H}^H \mathbf{w}_{rd}(n) \qquad (4.4.1)$$

The Grammian matrix Δ can be obtained by applying the matched filter at the destination node

$$\Delta = \begin{bmatrix} \lambda & 0\\ 0 & \lambda \end{bmatrix} \tag{4.4.2}$$

- In the case of DF type of transmission

$$\lambda = \underbrace{\sum_{i=1}^{2} \sum_{k=1}^{2} (|g_{ik}|^2)}_{\lambda_d} + \underbrace{2\Re(g_{11}g_{21}^*e^{j\theta_1} + g_{12}g_{22}^*e^{j\theta_2})}_{\lambda_f} \quad (4.4.3)$$

– In the case of AF type of transmission

$$\lambda = \underbrace{\left|f_{11}g_{11}\right|^{2} + \left|f_{21}g_{21}\right|^{2} + \left|f_{12}g_{12}\right|^{2} + \left|f_{22}g_{22}\right|^{2}}_{\lambda_{d}} + \underbrace{2\Re(f_{11}^{*}g_{11}^{*}f_{21}g_{21} + f_{12}g_{12}^{*}f_{22}^{*}g_{22})e^{j(\theta_{2}-\theta_{1})}}_{\lambda_{f}}$$
(4.4.4)

where λ_d is the total channel gain for dual-antenna relay nodes and λ_f can be interpreted as the channel dependent interference parameters. It is evident from (4.4.3) and (4.4.4) that when the interference parameters are negative there will be loss in the cooperative diversity. Therefore, in order to overcome this problem and have full cooperative diversity:

- In the case of DF type of transmission, the phases of the different code symbols which are transmitted from the first antennas at both relay nodes must be rotated by appropriate phase angles $e^{j\theta_1}$ and $e^{j\theta_2}$, respectively, before they are transmitted, while the symbols from the second antennas of each relay node are kept unchanged as shown in Figure 4.3, where θ_1 and θ_2 can be obtained from (4.3.13).
- In the case of AF type of transmission, as shown in Figure 4.5, the one-bit feedback information are fed back to all antennas of both relay nodes where the value of the phase rotation angles θ_1 and θ_2 are determined such that

 $\theta_1 = \theta_2 = 0$ or $\theta_1 = \theta_2 = \pi$ to ensure that $e^{j(\theta_2 - \theta_1)} = 1$, and $\theta_1 = \pi$ and $\theta_2 = 0$ or $\theta_1 = 0$ and $\theta_2 = \pi$ to ensure that $e^{j(\theta_2 - \theta_1)} = -1$ [110]. The *SNR* is related to the channel gain λ as:

$$SNR = \frac{\lambda}{4} \frac{\sigma_s^2}{\sigma_w^2} \tag{4.4.5}$$

where σ_s^2 is the total transmit power of the desired signal and σ_w^2 is the noise power at the destination node.

- Step 2: Applying the ML detection at the destination to detect which symbols were actually transmitted from each antenna of each relay, the least squares (LS) method can be employed as follows
 - In the case of DF type of transmission

$$\hat{s}(t,n) = \arg \min_{s_t \in S} |\hat{r}_{rd}(t,n) - \lambda s_t|^2 \quad \text{for} \quad t \in 1, 2,$$

(4.4.6)

- In the case of AF type of transmission

$$\hat{\mathbf{s}}(n) = \arg \min_{\mathbf{s}_t(n) \in \mathbf{S}} ||\mathbf{r}_{rd}(n) - \sqrt{\frac{P_1 P_2}{P_1 + 1}} \mathbf{H} \mathbf{s}_t(n)||$$
(4.4.7)

where \mathbf{S} denotes the set of all possible vector symbols.

It can be noticed from (4.4.6) and (4.4.7) that the ML detector will suffer from the destination synchronization error due to the presence of the interference component $\mathbf{I}_{int}(n)$ in (4.3.11) and (4.3.20) which will destroy the orthogonality of the received signal causing a severe degradation in the system performance. In order to overcome the synchronization error, a novel near-optimum detection scheme is applied for both transmission scenarios as explained hereinafter.

4.5 Near-optimum detection for two dual-antenna relay nodes

The proposed near-optimum detection scheme for the DF and AF type of transmission can be used to eliminate the effect of $\mathbf{I}_{int}(n)$ in (4.4.1), where s(1, n - 1) is in fact already known if the detection process has been initialized properly. Also, the interference components

$$(g_{12}(-1)U_2 + g_{22}(-1))s^*(1, n-1)$$

and

$$(g_{12}(-1)f_{12}^*U_1 + g_{22}(-1)f_{22}^*U_2)s^*(1, n-1)$$

presented in (4.3.8) and (4.3.24) can be removed before applying the LT in (4.4.1) and can be rewritten as follows

$$\hat{\mathbf{y}}(n) = [\hat{y}(1,n), \hat{y}(2,n)]^T = \mathbf{H}^H \mathbf{r}_{rd}(n)$$
$$= \Delta \mathbf{s}(n) + \mathbf{z}(n) s^*(2,n) + \mathbf{v}(n)$$
(4.5.1)

where
$$\Delta = \mathbf{H}^{H}\mathbf{H} = \begin{bmatrix} \lambda & 0 \\ 0 & \lambda \end{bmatrix}$$
, $\mathbf{v}(i) = \mathbf{H}^{H}\mathbf{w}_{rd}(n)$,

• In DF

$$\mathbf{z}(n) = \begin{bmatrix} z(1,n) \\ z(2,n) \end{bmatrix} = \mathbf{H}^{H} \begin{bmatrix} 0 \\ (U_{2}^{*}g_{12}^{*}(-1) + g_{22}^{*}(-1)) \end{bmatrix}$$

and

• In AF

$$\mathbf{z}(n) = \begin{bmatrix} z(1,n) \\ z(2,n) \end{bmatrix} = \mathbf{H}^{H} \begin{bmatrix} 0 \\ g_{12}^{*} f_{12} e^{j\theta_{2}} g_{22}^{*} f_{22} e^{j\theta_{1}} \end{bmatrix}$$

Hence, (4.5.1) can then be rewritten as

$$\hat{y}(1,n) = \lambda \ s(1,n) + z(1,n) \ s^*(2,n) + v(1,n)$$
(4.5.2)

and

$$\hat{y}(2,n) = \lambda \ s(2,n) + z(2,n) \ s^*(2,n) + v(2,n)$$
 (4.5.3)

 $\hat{y}(2,n)$ in (4.5.3) is only related to s(2,n) and therefore s(2,n) can be detected by using the LS method as follows

$$\hat{s}(2,n) = \arg \min_{s_m \in S} |\hat{y}(2,n) - \lambda \ s_m - z(2,n) \ s_m^*|^2 \qquad (4.5.4)$$

Then, the detection of s(1, n) can be carried out using the LS method by substituting $\hat{s}(2, n)$ detected in (4.5.2), as follows:

$$\hat{s}(1,n) = \arg \min_{s_m \in S} |\hat{y}(1,n) - \lambda \ s_m - z(1,n) \ \hat{s}^*(2,n)|^2 \quad (4.5.5)$$

The above procedure totally mitigates the interference induced by different time delays from the antennas of the second relay node at the destination node. The optimality of the above procedure in terms of ML detector can be achieved if there is no decision feedback error s(t, n-1), $t \in 1, 2$, therefore the above procedure is termed near-optimum detection. Moreover, the above analysis has shown that the detection complexity of this approach is only dependent upon the constellation size as compared with detection schemes presented in [90] which also depends on the number of PIC iterations. Also the above detection approach does not rely on the detection result of the DT link as compared with the sub-optimum detection approach in [9] and by reducing the timing error among the relay nodes when equipped with two antennas on each relay node, the complexity is significantly reduced. Furthermore, the fourth order diversity with coding gain and unity data transmission rate between the relay nodes and the destination node are exploited by this approach with the same decoding complexity as [48]. On the other hand, in the AF type of transmission, the complexity on the relay nodes was further reduced as compared with the DF type of transmission.

4.6 Simulation Results

In this section, simulation results for the proposed near-optimum detections scheme for both DF and AF types of transmission with two dual-antenna relay nodes utilizing two different feedback approaches is shown. In simulation, it is assumed that all channels are quasi-static Rayleigh fading frequency flat channels, all simulations are coded using QPSK symbols. The CL-EO-STBC with outer convolutive coding under perfect synchronization (PS) and with relay selection (RS) is included as reference in all figures. The destination node has full knowledge of the CSI.

a. For DF Network Using Two-Bit Feedback

Assuming perfect detection at relays, the average BER is plotted against the SNR. The SNR is defined as $\text{SNR} = \sigma_s^2 / \sigma_n^2$ (dB), and all relays transmit at 1/2 power. In Figure 4.6, the performance of the RS scheme discussed in Section 4.3 for perfectly synchronized distributed CL-EO-STBC relay network using two-bit feedback scheme is shown. It can be observed that there is a significant improvement in the BER when the RS scheme is applied, for example, at BER= 10^{-3} , there is approximately a 2dB gain with RS.



Figure 4.6. The probability of error performance comparison of the perfectly synchronized, with and without RS, conventional detector for distributed CL-EO-STBC and outer convolutive coding using two relay nodes each relay node has two antennas.

Figure 4.7 shows the performance of the distributed CL-EO-STBC network when time misalignment is introduced. The perfectly synchronized distributed CL-EO-STBC with RS is also included in the figure as a reference. In this figure, the performance comparison in terms of BER is made between the asynchronous transmision with RS as well as minimum selection of time misalignment (β) in the range of [-6 ... 6] dB and without RS with fixed time misalignment of -6 dB.

It can be noticed that the performance improvement is due to the RS scheme adopted, for example, approximately 9dB SNR is required



Figure 4.7. Probability of error performance of asynchronous distributed CL-EO-STBC with relay selection and randomly selected minimum time misalignment versus fixed time misalignment without relay selection.

to achieve a BER of 10^{-3} whereas, 12dB is required for achieving the same value of BER when the RS scheme is not adopted. The performance of asynchronous distributed CL-EO-STBC network with RS and different time misalignment values is shown in Figure 4.8. The value of β is in the range of [-6 ... 6]dB, [-3 ... 6]dB, and [0 ... 6]dB. The effect of selecting the smallest time misalignment is clear on the error performance where 9dB SNR attains a BER of 10^{-3} in the case of selecting the smallest value of β (i.e. in the range of [-6 ... 6]dB, whereas, with increasing β value, the greater the degradation in BER . In Figure 4.9, the effect of applying the RS scheme with selecting the minimum time misalignment on the distributed CL-EO-STBC for four single-antenna relay nodes is shown [9] as compared to the proposed distributed CL-EO-STBC for two dual-antenna relay nodes. It can be seen that the proposed network significantly



Figure 4.8. Probability of error performance of asynchronous distributed CL-EO-STBC with relay selection and randomly selected minimum time misalignment with different ranges.

outperforms the scheme in [9] when applying the RS with minimum delay algorithm. The results in this figure shows that no matter what SNR is employed a BER of 10^{-3} is not attainable with [9]. Figure 4.10 shows the effectiveness of the proposed near-optimum detection scheme in mitigating the effect of asynchronism. The proposed scheme completely eliminates the effect of ISI due to time misalignment and enhancing the diversity order. The results also show that the smaller the time misalignment, the better the performance in terms of BER. For example, at a BER of 10^{-3} , the proposed nearoptimum scheme with RS and minimum time misalignment (in the range [-6 ... 6]dB) requires approximately 4.7 dB of SNR and in the range [-3 ... 6]dB requires approximately 5 dB of SNR, while the conventional perfectly synchronized distributed CL-EO-STBC scheme requires approximately 4 dB of SNR for achieving the same value of BER. Table 4.1 illustrates the comparison between the proposed



Figure 4.9. Probability of error performance of asynchronous distributed CL-EO-STBC with relay selection and randomly selected minimum time misalignment for the case of two dual-antenna relay network versus four single-antenna relay network [9].

near-optimum and previous work in [48] and [9]. It can be seen that the cooperative diversity order of four can be obtained by using the proposed near-optimum and sub-optimum detection scheme in [9] for CL-EO-STBC, while in the near-optimum detection scheme in [48] cooperative diversity order cannot exceed more than two by utilizing the Alamouti code. However, the sub-optimum detection [9] requires the existence of the DT link between the source node and the destination node for a successful detection. In contrast, the proposed near-optimum detection scheme does not require the DT link between the source node and the destination node. In all simulations no error propagations were encountered.



Figure 4.10. Probability of error performance of the proposed nearoptimum detection scheme for distributed CL-EO-STBC using two dual-antenna relay nodes for different randomly selected minimum time misalignment.

b. For AF Network Using One-Bit Feedback

In the simulation, the average BER is plotted against the total transmission power. The one-bit feedback EO-STBC scheme is utilized. The cooperative AF transmission type with relay selection is used. In Figure 4.11, the one-bit feedback EO-STBC in [110] is compared with the conventional EO-STBC and the two-bit feedback EO-STBC in [30] without the RS in Figure 4.11a, and with the RS as in Figure 4.11b. It can be noticed that the one-bit feedback scheme outperforms the open-loop and the two-bit feedback schemes. For example it requires approximately 13.5 dB of total power to achieve a BER of 10^{-3} for the one-bit feedback scheme without RS whereas, approximately 11 dB is required for the same value of BER when RS is applied. The interference factor resulting from the asynchronism of the antennas on the second relay severely degrade the system

Detection	Near-optimum	Sub-optimum	Proposed near
Scheme	detection in [46]	detection in [47]	-optimum
			detection
Number of			
Relay nodes	2	4	2
Number of			
antennas at			
relay nodes	1	1	2
Number of			
time			
misalignments	1	3	2
Complexity at			
destination	Low	High	Low
Complexity at			
relay node	Low	Low	High
DT Link	Not needed	Needed	Not needed
Cooperative			
diversity			
order	2	4	4

Table 4.1. Comparison of proposed near-optimum scheme with previous detection scheme in [46] and [47].

performance as seen in Figure 4.12. The BER of one-bit feedback EO-STBC detector under PS have been simulated to show the impact of imperfect synchronization under different values of β . When the time misalignment, for instance, $\beta = 0$ dB and 3 dB, the one-bit feedback EO-STBC detector cannot deliver good performance comparing with the PS case. However, when the proposed RS scheme is used to randomly select the minimum time misalignment from a uniformly distributed values of $\beta = [0,...10]$ dB among the set of candidate relay nodes without prior knowledge of the values of timing errors, the performance of the one-bit feedback scheme was effectively enhanced achieving a BER of 10^{-3} with approximately 14.5 dB of total transmission power.



Figure 4.11. The end-to-end BER performance comparison of the conventional, one-bit feedback and two-bit feedback EO-STBC with outer coding. (a) Perfectly synchronized without RS (b) Perfectly synchronized with RS.

On the other hand, Figure 4.13 illustrates the performance of the near-optimum detection which significantly eliminates the effect of the interference components due to asynchronism even when the time misalignment is large with no computational complexity as compared with the work in [95]. The figure shows that it just requires approximately 11.2 dB to achieve a BER of 10^{-3} power when the near-optimum approach was combined with the minimum-delay RS technique. Furthermore, when $\beta = 3$ dB, the near-optimum approach requires approximately 17 dB of power to get a BER of 10^{-3} .



Figure 4.12. The end-to-end BER performance of the one-bit feedback EO-STBC with outer coding utilizing ML detection under imperfect synchronization when $\beta = 0$ dB, 3 dB and with minimum selection in the range from 0 dB to -10 dB.

4.7 Summary

In this chapter, a near-optimum detection approach with RS and minimum time misalignment selection was proposed and analyzed employing a half-duplex DF and AF type of relaying transmission based on distributed closed-loop EO-STBC with phase rotation [30] [111] and outer convolutive coding for wireless relay networks over frequency flat fading under imperfect synchronization. A system with two dual-antenna relay nodes was demonstrated in particular. Through simulation results and analysis process, it was shown that in both cases (DF and AF), this near-optimum approach is effective at removing ISI at the destination node caused by time misalignment among relay nodes with computational detection complexity at the destination node dependant only on the constellation size and without the need for the direct transmission link as compared to the previous work in [9]. Furthermore,



Figure 4.13. The end-to-end BER performance of the near-optimum detection approach for the one-bit feedback EO-STBC with outer coding when $\beta = 0$ dB, 3 dB and with minimum selection in the range from 0 dB to -10 dB.

the fourth order diversity with coding gain and unity data transmission rate between the relay nodes and the destination node is exploited by the proposed approach that employed two dual-antenna relay nodes with the equivalent decoding complexity as in [48]. Moreover, it has been shown through simulation results that a simple max—min relay selection method was effective in enhancing the system performance by selecting the best links and the smallest time delay error for cooperative transmission together with exploiting the available cooperative diversity order of four as compared to distributed CL-EO-STBC with outer coding under perfect synchronization among the relay nodes without relay selection. Finally, when the AF protocol is used, the complexity at the relay nodes is reduced and the feedback resolution was restricted to one-bit only which is practically achievable in the bi-directional control channels present in many communication systems, reducing by which the high cost and high complexity due to the increased feedback information utilized in the cased of DF transmission protocol. In the next chapter a one-bit feedback scheme for asynchronous relay network utilizing EO-STBC and orthogonal frequency division multiplexing (OFDM) type transmission over frequency flat channels will be presented.

OFDM-BASED SPACE-TIME BLOCK CODING TRANSMISSION SCHEMES FOR ASYNCHRONOUS COOPERATIVE RELAY NETWORKS

Employing multiple relay nodes between source and destination nodes can provide enhanced cooperative diversity in wireless relay systems as shown in the previous chapters. However, one of the key challenges in practice when designing high-performance distributed space-time code systems is symbol-level synchronization among these multiple relay nodes. This effective problem of synchronization occurs due to several factors such as different propagation delays, different relay locations, and different relay oscillators. As a result, the relay signals arrive at different time instants at the destination which may cause inter-symbol interference (ISI). The cyclic prefix can also be used for symbol synchronization because it is a known repetition of some part of the received signal that can be detected through autocorrelation. It is especially well-suited for fast-changing channels, because the delay time can be updated on a per-symbol basis. Aditionaly, it does not add any additional overhead [112]. In this chapter, a novel robust scheme for two dual-antenna relay nodes to employ in cooperative relay networks without the requirement of exact synchronization between relay nodes is proposed. The design exploits distributed modified quasi orthogonal space-time block coding (M-QO-STBC) type of transmission that can achieve full cooperative diversity and code gain distance, and distributed extended orthogonal space-time block coding type of cooperative transmission (D-EO-STBC). Orthogonal frequency division multiplexing (OFDM) is implemented at the source node. A cyclic prefix (CP) is added at the source and relay nodes to combat the effects of random delays at the relay nodes. The relays operate in a simple amplify-and-forward (AF) mode. In this chapter, a narrowband system, where the two-hop channels are assumed to be flat Rayleigh fading channels is initially considered with the M-QO-STBC transmission technique. Next, a new low complexity one-bit feedback scheme based on the selection of the cyclic phase rotation scheme is proposed for broadband systems, where the two-hop channels are assumed to be frequency-selective Rayleigh fading channels, and the D-EO-STBC transmission technique is employed. This approach attains unity rate over each hop in the network and full cooperative spatial diversity. Simulation results are included to confirm that these schemes provide better performance as compared with previous schemes.

5.1 Introduction

Cooperative transmission protocols can commonly be categorized into two general forms amplify-and-forward (AF) or decode-and-forward (DF). In practice, the AF scheme is more practical for its low-complexity relay transceivers and low power consumption since the cooperative terminals do not need to re-encode the received signals. In order to maximize the diversity advantage in cooperative systems, the deployment of orthogonal space time block coding (OSTBC) [29] [57] is presented in a distributed fashion, i.e. a distributed-STBC (D-STBC), thereby providing cooperative diversity gain in proportion to the number of transmitting relay nodes for wireless relay networks with a single antenna at each relay node [17] [27]. The major issues of timing synchronization among cooperative relay nodes were considered and dealt with using approaches that were proposed for flat fading channels which limits their application to narrowband communication systems [35] [113] [98]. Moreover, the approaches used the DF scheme at relay nodes, where each relay node has a single antenna. In [98] a simple distributed STBC transmission scheme based on OFDM is proposed to combat the timing error between relay nodes, but this is only valid for the case of two relay nodes and is still limited to narrowband communication systems. However, in [49] the idea of exploiting OFDM type of transmission to mitigate timing errors and ISI from multipath fading in asynchronous cooperative relay networks over frequency selective fading channels in

the case of two relay nodes was proposed. Nevertheless, full cooperative diversity with full data rate can not both be achieved in the case of more than two relay nodes [58] [28].

In [30], a closed-loop method for STBC has been proposed to exploit full diversity and full transmission rate for point-to-point transmission for four transmit and one receiver antenna. They apply phase rotation to two transmit antennas based on CSI feedback from the receiver antenna, but this approach is limited since two-bits are required. Furthermore, the work in [101] presented a closed-loop scheme for use in asynchronous cooperative networks over frequency selective channels. In addition to implementing four relay nodes which incurred more time delay, the feedback scheme also required some quantization technique to reduce the overhead incurred by this proposed scheme. Most proposed approaches in wireless relay networks are built upon the assumption that all relay nodes are equipped with a single antenna.

In this chapter, an asynchronous cooperative system is assumed using the M-D-QO-STBC scheme in which distributed OFDM transmission is used with two relay nodes each equipped with two antennas over frequency flat channels. Furthermore, a new closed-loop scheme is proposed for D-EO-STBC over frequency selective fading channels in wireless relay networks utilizing one-bit feedback per subcarrier based on selection of cyclic phase rotation in which distributed OFDM transmission is used with two relay nodes each equipped with two antennas. The performance improvement is also investigated when the proposed schemes are combined with outer convolutive coding to exploit further coding gain.

5.2 Asynchronous Cooperative Relay Networks

In these networks, the source node transmits its information to the destination node via a collection of cooperative relay nodes. This section will consider narrowband and broadband communication systems where synchronization among the relays nodes is not required.

5.2.1 System model and problem statement

Consider a wireless relay system with one source node, one destination node, and two dual-antenna half-duplex relay nodes, as shown in Figure 5.1. The relay nodes are distributively located between the source and



Figure 5.1. Asynchronous cooperative relay network constituting of source, destination and two dual-antenna relay nodes.

destination nodes. The source coverage extends to include the relay nodes but not the destination node due to deep fading, heavy path loss or shadowing effects, or the coverage design of the source. So, there is no direct path between the source and the destination nodes. The relay nodes assist the source in transmitting signals to the destination. Therefore, the system needs two phases to transmit the signals from the source node to the destination node. In the first phase, the source node broadcasts the information symbols to the two relay nodes and then stops sending during the second phase. While in the second phase, the two relay nodes process the received signals and then retransmits them to the destination node. The quasi-static fading coefficients from the source node to each antenna on each relay node are represented as f_{ik} , the relaying channel gain from the i^{th} antenna on the k^{th} relay node to the destination node is represented as g_{ik} , where $i, k \in 1, 2$. Each channel coefficient of f_{ik} and g_{ik} is modelled as an independent complex circular Gaussian random variable with zero-mean and unit variance. In real-life wireless applications, each relay node in a cooperative relay



Figure 5.2. Asynchronous cooperative relay network constituting of source, destination and two dual-antenna relay nodes together with relative time delays.

system is in a different location and has its own oscillator so the relayed signals arrive at the destination at different time instants, however, timing synchronization is easily performed for the shortest path from the relays to the destination. Without much loss of generality, this will be assumed to be the path from both antennas of R_1 . Hence, the delays for different relayed signals $\tau_{(i2)}$, for $i \in 1, 2$, are assumed relative to the received signal from each antenna of R_1 (i.e. $\tau_{(i1)} = 0$), as shown in Figure 5.2. Such relative delays will cause ISI between subcarriers. To solve this problem, modified D-QO-STBC and closed-loop D-EO-STBC schemes based on OFDM pre-coding with CP insertion are proposed next in this chapter.

5.2.2 M-QO-STBC scheme for narrowband systems

The term narrowband system is used here to refer to the type of wireless channels employed in this section which are assumed to be frequencyflat. Specifically, the channels between any two terminals are assumed to be quasi-static flat Rayleigh fading. Therefore, f_{ik} and g_{ik} are modelled as independent complex circular Gaussian random variables with zero-mean and unit variance.

Broadcasting phase

In this phase, the source node broadcasts sequentially four consecutive OFDM blocks

$$\overline{\mathbf{X}} = \sqrt{P_1} \quad \mathbf{IFFT}(\mathbf{X})$$
$$= \sqrt{P_1} \quad [\mathbf{IFFT}(\mathbf{X}_1) \quad (\mathbf{IFFT}(\mathbf{X}_2))^* \quad \mathbf{IFFT}(\mathbf{X}_3) \quad (\mathbf{IFFT}(\mathbf{X}_4))^*]^T$$
$$= [\overline{\mathbf{X}}_1 \quad \overline{\mathbf{X}}_2 \quad \overline{\mathbf{X}}_3 \quad \overline{\mathbf{X}}_4]^T$$

where P_1 is the power allocation factor at the source and

$$\mathbf{X}_m = [x_{m,0}, x_{m,1}, \dots, x_{m,N-1}]$$

for $m \in \{1, 2, 3, 4\}$ represents the four consecutive time slots as shown in Figure 5.3 .

Each OFDM symbol is proceeded with a CP with length l_{cp} .



Figure 5.3. The baseband model of the OFDM at the source node.

Assume that l_{cp} is not less than the channel memory length L and the maximum of the possible relative timing errors τ_{max} of the signals arriving at the destination node, i.e. L = 0 due to flat fading assumption.

Relaying phase

In this stage, the received signals at the relay nodes are processed and forwarded to the destination. It is assumed that the channel coefficients are constant during four OFDM symbol intervals. The received signal at each antenna i of each relay node k, in four successive OFDM symbol durations after removing the CP at the relay nodes can be expressed as:

$$\mathbf{y}_{(ik)}^{m} = f_{ik} \,\overline{\mathbf{X}}_{m} + \mathbf{V}_{(ik)}^{m} \quad \text{for} \quad m \in \{1, 2, 3, 4\}$$
 (5.2.1)

where the vector $\mathbf{V}_{(ik)}^{m}$ includes the corresponding zero-mean unit-variance Additive White Gaussian Noise (AWGN) terms at the *i*-th antenna of the *k*-th relay node at time slot m.

The mean power of the signal $\mathbf{y}_{(ik)}^m$ at relay nodes is $P_1 + 1$ due to



Figure 5.4. The baseband model of OFDM for one dual-antenna relay node.

the unit variance assumption of the fading channels from the source node to relay nodes $\mathbf{f}_{(ik)}$ and the additive noise $\mathbf{V}_{(ik)}^m$ at the relays. Let P_2 be the average transmission power at each antenna of each relay node, then the optimum power allocation proposed in [14] is used in this scheme, which yields:

$$P_1 = \frac{P}{2}, \ P_2 = \frac{P}{2R}$$

where P is the total transmission power for the whole scheme, and R is the number of relay nodes.

As mentioned above, an AF type of transmission is deployed, so the two dual-antenna relay nodes will perform QO-STBC on the received noisy signal at each antenna i of each relay node k as shown in Table 5.1.

Therefore, the transmitted signals for two consecutive OFDM symbols

	v	1	0	v
	Relay1 Ant.1	Relay1 Ant.2	Relay2 Ant.1	Relay2 Ant.2
OFDM1	$\sqrt{rac{P_2}{P1+1}} \mathbf{y}_{11}^1$	$-\sqrt{rac{P_2}{P1+1}}{f y}_{21}^{*2}$	$\sqrt{rac{P_2}{P1+1}}\mathbf{y}_{21}^3$	$-\sqrt{rac{P_2}{P1+1}}\mathbf{y}_{22}^{*4}$
OFDM2	$\sqrt{\frac{P_2}{P1+1}}\zeta(\mathbf{y}_{11}^2)$	$\sqrt{\frac{P_2}{P_{1+1}}}\zeta(\mathbf{y}_{21}^{*1})$	$\sqrt{\frac{P_2}{P_{1+1}}}\zeta(\mathbf{y}_{12}^4)$	$\sqrt{\frac{P_2}{P_{1+1}}}\zeta(\mathbf{y}_{22}^{*3})$
OFDM3	$\sqrt{rac{P_2}{P1+1}}\mathbf{y}_{11}^3$	$-\sqrt{rac{P_2}{P_{1+1}}}\mathbf{y}_{21}^{*4}$	$\sqrt{rac{P_2}{P_{1+1}}}\mathbf{y}_{21}^1$	$-\sqrt{rac{P_2}{P_{1+1}}}\mathbf{y}_{22}^{*2}$
OFDM4	$\sqrt{\tfrac{P_2}{P1+1}}\zeta(\mathbf{y}_{11}^4)$	$\sqrt{\tfrac{P_2}{P1+1}}\zeta(\mathbf{y}_{21}^{*3})$	$\sqrt{\tfrac{P_2}{P1+1}}\zeta(\mathbf{y}_{12}^2)$	$\sqrt{\frac{P_2}{P1+1}}\zeta(\mathbf{y}_{22}^{*1})$

 Table 5.1. QO-STBC processing at the relay nodes

from all antennas of the two relay nodes are

$$\begin{bmatrix} \mathbf{t}_{11}^{1} & \mathbf{t}_{21}^{1} & \mathbf{t}_{12}^{1} & \mathbf{t}_{22}^{1} \\ \mathbf{t}_{21}^{2} & \mathbf{t}_{22}^{2} & \mathbf{t}_{22}^{2} \\ \mathbf{t}_{11}^{3} & \mathbf{t}_{21}^{3} & \mathbf{t}_{12}^{3} & \mathbf{t}_{22}^{3} \\ \mathbf{t}_{11}^{4} & \mathbf{t}_{21}^{4} & \mathbf{t}_{12}^{4} & \mathbf{t}_{22}^{4} \end{bmatrix} = \rho \begin{bmatrix} \mathbf{y}_{11}^{1} & -\mathbf{y}_{21}^{*2} & \mathbf{y}_{12}^{3} & -\mathbf{y}_{22}^{*4} \\ \zeta(\mathbf{y}_{11}^{2}) & \zeta(\mathbf{y}_{21}^{*1}) & \zeta(\mathbf{y}_{12}^{4}) & \zeta(\mathbf{y}_{22}^{*3}) \\ \mathbf{y}_{11}^{3} & -\mathbf{y}_{21}^{*4} & \mathbf{y}_{12}^{1} & -\mathbf{y}_{22}^{*2} \\ \zeta(\mathbf{y}_{11}^{4}) & \zeta(\mathbf{y}_{21}^{*3}) & \zeta(\mathbf{y}_{12}^{2}) & \zeta(\mathbf{y}_{22}^{*1}) \end{bmatrix}$$

$$(5.2.2)$$

where $\zeta(.)$ denotes the modulo-*M* time-reversal of the signal and $\rho = \sqrt{\frac{P_2}{P_{1+1}}}$ is the signal amplification factor scalar [49] [105].

Implementation at the destination node

As shown in Figure 5.4, the CP is removed and re-inserted at each antenna of each relay node in the second phase. The advantage of the distributed QO-STBC protocol at the relay nodes without the requirement of the decoding operation is reducing the complexity at the relay nodes as well as lower power consumption, since the relay nodes only implement very simple linear transform operation as shown in (5.2.2). The relayed signals arrive at the destination at different time instants, however, timing synchronization is easily performed for the shortest path from the relays to the destination. Without much loss of generality, this will be assumed to be the path from both antennas of R_1 . Hence, the delays for different relayed signals $\tau_{(i2)}$ are assumed relative to the received signal from each antenna of R_1 as shown in Figure 5.5. Such relative delays will cause ISI between subcarriers. However, the receiver still can overcome the effect of ISI and the or-



Figure 5.5. (a) OFDM frame structure after CP insertion. (b) CP removal with respect to relay 1 synchronization.

thogonality between subcarriers can still be maintained because the l_{cp} is not less than $L + \tau_{max}$, where L is the maximum channel memory length between all antennas of the two relay nodes and the destination node, $\tau_{max} = \max{\{\tau_{(ik)}\}}$ is the maximum overall relative delay from the source to all the receive antennas at the relays and from these antennas to the destination node, and $\tau_{(i1)} = 0$, for $i \in 1, 2$. Also, without loss of generality, the delays from both antennas of the second relay R_2 are assumed to be equal due to their vicinity and co-location on the same relay, i.e. $\tau_{(12)} = \tau_{(22)} = \tau_2$.

On the other hand, the delay in the time domain can be interpreted in the frequency domain as phase changes

$$\mathbf{D}^{\tau_2} = \begin{bmatrix} D_0^{\tau_2}, D_1^{\tau_2}, ..., D_{N-1}^{\tau_2} \end{bmatrix}$$
(5.2.3)

where $D_n^{\tau_2} = e^{-j2\pi n\tau_2/N}$, for n = 0, 1, ..., N - 1. The receiver then removes the CP from each OFDM symbol and performs the FFT operation for the four successive OFDM symbols $\mathbf{Z}_m = [Z_{m,0}, Z_{m,1}, ..., Z_{m,N-1}],$ (m=1,2,3,4) as shown in Figure 5.6.

Consequently, \mathbf{Z}_1 , \mathbf{Z}_2 , \mathbf{Z}_3 and \mathbf{Z}_4 in the frequency domain can be



Figure 5.6. The baseband model of OFDM at receiver node.

written as:

$$\begin{aligned} \mathbf{Z}_1 \ &= \ \rho[\mathbf{FFT}(\overline{\mathbf{X}}_1) \circ \Lambda_{11} - \mathbf{FFT}(\overline{\mathbf{X}}_2^*) \circ \Lambda_{21} \\ &+ \ \mathbf{FFT}(\overline{\mathbf{X}}_3) \circ \Lambda_{12} - \mathbf{FFT}(\overline{\mathbf{X}}_4^*) \circ \Lambda_{22}] + \mathbf{W}_1 \end{aligned}$$

$$\begin{split} \mathbf{Z}_2 \ &= \ \rho[\mathbf{FFT}((\mathbf{X}_2)) \circ \Lambda_{11} + \mathbf{FFT}((\mathbf{X}_1^*)) \circ \Lambda_{21} \\ &+ \ \mathbf{FFT}((\overline{\mathbf{X}}_4)) \circ \Lambda_{12} + \mathbf{FFT}((\overline{\mathbf{X}}_3^*)) \circ \Lambda_{22}] + \mathbf{W}_2 \end{split}$$

$$egin{aligned} \mathbf{Z}_3 \ &= \ &
ho[\mathbf{FFT}(\mathbf{X}_3) \circ \Lambda_{11} - \mathbf{FFT}(\mathbf{X}_4^*) \circ \Lambda_{21} \ &+ \ \mathbf{FFT}(\overline{\mathbf{X}}_1) \circ \Lambda_{12} - \mathbf{FFT}(\overline{\mathbf{X}}_2^*) \circ \Lambda_{22}] + \mathbf{W}_3 \end{aligned}$$

$$\mathbf{Z}_{4} = \rho[\mathbf{FFT}((\mathbf{X}_{4})) \circ \Lambda_{11} + \mathbf{FFT}((\mathbf{X}_{3}^{*})) \circ \Lambda_{21} \\ + \mathbf{FFT}((\mathbf{\overline{X}}_{2})) \circ \Lambda_{12} + \mathbf{FFT}((\mathbf{\overline{X}}_{1}^{*})) \circ \Lambda_{22}] + \mathbf{W}_{4}$$

$$(5.2.4)$$

with

$$\Lambda_{ik} = \begin{cases} \mathbf{FFT}(\mathbf{f}_{(i1)}) \circ \mathbf{FFT}(\mathbf{g}_{(i1)}) & \text{for } k=1, i=1, 2\\ \mathbf{D}^{\tau_2} \circ \mathbf{FFT}(\mathbf{f}_{(i2)}) \circ \mathbf{FFT}(\mathbf{g}_{(i2)}) & \text{for } k=2, i=1, 2. \end{cases}$$

where the operation \circ is the Hadamard product, i.e., the componentwise product, and $\mathbf{W}_m = \rho(\sum_{m=1}^4 \mathbf{FFT}(\mathbf{V}_{(i1)}^m) \circ (\mathbf{g}_{(i1)}) + \mathbf{D}^{\tau_2} \circ \mathbf{FFT}(\mathbf{V}_{(i2)}^m) \circ (\mathbf{g}_{(i2)})]) + \mathbf{n}_m$ where \mathbf{n}_m is the AWGN vectors at the destination node at four successive time slots (m=1,2,3,4) with elements having zero-mean and unit variance.

By taking into consideration the simple identities $(\mathbf{FFT}(X))^* = \mathbf{IFFT}(X^*)$, $\mathbf{FFT}(\zeta(\mathbf{FFT}(X))) = \mathbf{IFFT}(\mathbf{FFT}(X)) = X, \zeta(\mathbf{FFT}(X)) = \mathbf{IFFT}(X)$, $\zeta(\mathbf{IFFT}(X)) = \mathbf{FFT}(X), (\mathbf{IFFT}(X))^* = \mathbf{FFT}(X^*)$, and $\mathbf{FFT}(\zeta(X)) = \zeta(\mathbf{FFT}(X))$, then the equations in (5.2.4) can be written to have the QO-STBC structure on each subcarrier $n, 0 \le n \le N - 1$, as follows:

$$\begin{bmatrix} Z_{1,n} \\ Z_{2,n} \\ Z_{3,n} \\ Z_{4,n} \end{bmatrix} = \rho \begin{bmatrix} x_{1,n} & -x_{2,n}^* & x_{3,n} & -x_{4,n}^* \\ x_{2,n} & -x_{1,n}^* & x_{4,n} & x_{3,n}^* \\ x_{3,n} & -x_{4,n}^* & x_{1,n} & -x_{2,n}^* \\ x_{4,n} & -x_{3,n}^* & x_{2,n} & x_{1,n}^* \end{bmatrix} \begin{bmatrix} \Lambda_{(11),n} \\ \Lambda_{(21),n} \\ \Lambda_{(12),n} \\ \Lambda_{(22),n} \end{bmatrix} + \begin{bmatrix} w_{1,n} \\ w_{2,n} \\ w_{3,n} \\ w_{4,n} \end{bmatrix}$$
(5.2.5)

where $w_{1,n}$, $w_{2,n}$, $w_{3,n}$ and $w_{4,n}$ are the total AWGN at the *i*-th antenna of each relay node and the destination node with zero-mean and unity variance.

The code from (5.2.5) shows that the timing errors only cause phase shift in the channel frequency response, and the structure of the QO-STBC still holds. Therefore, pairwise decoding can be applied at the destination node.

Improved QO-STBCs can be achieved by using different constellations for different symbols. For example, in [28] the third and fourth symbols are rotated before transmission, and it is shown that it is possible to provide full diversity for distributed QO-STBCs. In the next section, the performance of D-QO-STBCs based on an amplify-and-forward (AF) type protocol is further improved by using a modified code.

5.2.3 Modified code design

The design of A_i and B_i which give the quasi-orthogonal code are

$$A_1 = I_4, \ A_2 = A_4 = 0_4, \ A_3 = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}$$

 B_2

$$B_{1} = B_{3} = 0_{4}$$

$$= \begin{bmatrix} 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \end{bmatrix}, B_{4} = \begin{bmatrix} 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \\ 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{bmatrix}$$
(5.2.6)

The matrix 0_4 indicates a 4×4 matrix with all zero elements. The rotation using modified D-STBCs is applied either on the first pair or the second pair of symbols [114], thus the codeword is:

$$X(\Phi_{1},\Phi_{2}) = \begin{bmatrix} x_{1}e^{j\Phi_{1}} & x_{2}e^{j\Phi_{1}} & x_{3}e^{j\Phi_{2}} & x_{4}e^{j\Phi_{2}} \\ -x_{2}^{*}e^{-j\Phi_{1}} & x_{1}^{*}e^{-j\Phi_{1}} & -x_{4}^{*}e^{-j\Phi_{2}} & x_{3}^{*}e^{-j\Phi_{2}} \\ x_{3}e^{j\Phi_{2}} & x_{4}e^{j\Phi_{2}} & x_{1}e^{j\Phi_{1}} & x_{2}e^{j\Phi_{1}} \\ -x_{4}^{*}e^{-j\Phi_{2}} & x_{3}^{*}e^{-j\Phi_{2}} & -x_{2}^{*}e^{-j\Phi_{1}} & x_{1}^{*}e^{-j\Phi_{1}} \end{bmatrix}$$
(5.2.7)

where x_1 , x_2 , x_3 , and x_4 are symbols from the constellation and Φ_1 and Φ_2 are the rotation angles. The matrices $X(\Phi_1, \Phi_2)$ consist of a whole quasi-orthogonal codebook. By changing (Φ_1, Φ_2) a family of codebooks can be generated, for BPSK modulation $\Phi_1 = \Phi_2 = \frac{\pi}{2}$. At the beginning, the union of two quasi-orthogonal codebooks, $\mathbb{T} \cup \mathbb{S}$ is taken, where $\mathbb{S} = X(\frac{\pi}{2}, 0)$ and $\mathbb{T} = X(0, \frac{\pi}{2})$. Given that $\mathbb{T} \cup \mathbb{S}$ has twice as many codewords the union is pruned down by one-half to get the new code \mathbb{C} , which has the original rate. The performance of the modified codebook \mathbb{C} is bounded below by the performance of quasiorthogonal codes \mathbb{S} and \mathbb{T} , because each of them is one possible pruning of $\mathbb{S} \cup \mathbb{T}$. Therefore, the process can only improve the code. Firstly, the codebooks \mathbb{S} and \mathbb{T} are partitioned such that each partition has good distance properties, then \mathbb{C} is constructed by combining these
partitions [114]. For instance, for the BPSK case, firstly, $X(0, \frac{\pi}{2})$ is set to generate 16 possible matrices which is S, then all the 16 matrices are presented by their binary values. For the second set \mathbb{T} , $X(\frac{\pi}{2}, 0)$ is taken to generate another 16 matrices, so the total number of matrices which is generated in $\mathbb{T} \cup \mathbb{S}$ is 32 matrices. Since $\mathbb{T} \cup \mathbb{S}$ has twice as many codewords, the first 16 matrices are taken from the first set and divided into:

$$\mathbb{S}_1 = \{0000, 0011, 0101, 0110, 1001, 1010, 1100, 1111\}$$

 $\mathbb{S}_2 = \{0001, 0010, 0100, 0111, 1000, 1011, 1101, 1110\}$

now there is a new codeset S_1 , which is made from 8 codewords, each codeword is represented by its four binary symbols (x_1, x_2, x_3, x_4) . The same steps are taken for the second set to obtain \mathbb{T}_1 and \mathbb{T}_2 . The new codebook $\mathbb{C} = \mathbb{S}_1 \cup \mathbb{T}_2$.

5.2.4 Code gain distance

The inter-distance of the subcodes can be generated in a manner similar to [115]. Thus for two possible codewords X and Y the code gain distance CGD = det(Π) where

$$\Pi = ((X - Y)(X - Y)^H).$$

The minimum CGD of the codebook S is defined as the minimum CGD of all non identical codeword pairs of $(S \times S)$, therefore the distance between two codebooks S and T is given by

$$D(\mathbb{S}, \mathbb{T}) = \det(\Pi(X, Y)) \tag{5.2.8}$$

The minimum CGD of $\mathbb S$ and $\mathbb T$ is found to be

$$D(\mathbb{S}_1, \mathbb{S}_2) = D(\mathbb{T}_1, \mathbb{T}_2) = 256 \tag{5.2.9}$$

and for \mathbb{S}_1 and \mathbb{T}_2 is:

$$D(\mathbb{S}_1, \mathbb{T}_2) = 2304 \tag{5.2.10}$$

It is clear that the distance of the new code is increased, in particular the new codebook is built from $S_1 \cup T_2$ which has higher distance and thereby yields better performance. For higher rate codes the distance properties of the code can be generated and for QPSK modulation it is

$$D(\mathbb{S}_1, \mathbb{T}_2) = 2.22 \tag{5.2.11}$$

It is clear that the modified code has a small distance thus the distance property of the code is not improved greatly which affects the performance of the system. In order to increase the distance of the modified code a unitary rotation matrix \mathbf{U} is used. The codewords in \mathbb{T}_2 are multiplied by the matrix \mathbf{U} as

$$\max_{\mathbf{U}} D(\mathbb{S}_1, \mathbb{T}_2 \mathbf{U})$$

where

$$\mathbf{U} = \text{diag}(e^{j\theta_1}, e^{j\theta_2}, e^{j\theta_3}, e^{j\theta_4})$$
(5.2.12)

and diag(.) denotes a diagonal matrix. A genetic algorithm is used to search for the optimum rotation matrix to maximize the distance property between the codes. The MATLAB fitness function reference of the genetic algorithm has four parameters for the rotation matrix the lower bound for each parameter is -2π and the upper bound is 2π . The size of populations which is used is 50000 with 100 generations. It is found that the optimum rotation which gives the maximum distance is

$$\mathbf{U} = \operatorname{diag}(e^{j1.5832\pi}, e^{j.4165\pi}, e^{j.9165\pi}, e^{j1.0832\pi})$$
(5.2.13)

The maximum distance from the optimum rotation is

$$D(\mathbb{S}_1, \mathbb{T}_2\mathbf{U}) \approx 64$$

therefore the distance is maximized because of the optimum rotation matrix which improves the performance of the modified code. A full maximum likelihood decoder (MLD) over the new codebook has been used in this work. In the following, the closed-loop D-EO-STBC scheme is presented.

5.2.5 D-EO-STBC scheme with Outer Coding for broadband systems

The system model in Figure 5.2 is considered. The channels between the source node and relay nodes, $\mathbf{f}_{(ik)}$, and between the relay nodes and the destination node, $\mathbf{g}_{(ik)}$, are assumed to be quasi-static frequencyselective Rayleigh fading channels so that the vectors $\mathbf{f}_{(ik)} = [f_{(ik),0}, f_{(ik),1}, \dots, f_{(ik),N-1}]$ and $\mathbf{g}_{(ik)} = [g_{(ik),0}, g_{(ik),1}, \dots, g_{(ik),N-1}]$, for $i, k \in$ 1, 2, and $n = 0, 1, \dots, N - 1$ corresponds to the channel length, have elements which are independent complex Gaussian random variables with elements having zero-mean and unit-variance. The half-duplex two dual-antenna relay nodes use AF relaying protocol.

Broadcasting phase

The source node broadcasts sequentially two consecutive OFDM blocks $\overline{\mathbf{X}} = \sqrt{P_1} \ \mathbf{IFFT}(\mathbf{X}) = \sqrt{P_1} \ [\mathbf{IFFT}(\mathbf{X}_1) \ (\mathbf{IFFT}(\mathbf{X}_2))^*]^T = [\overline{\mathbf{X}}_1 \ \overline{\mathbf{X}}_2]^T$, where P_1 is the power allocation factor at source and $\mathbf{X}_m = [x_{m,0}, x_{m,1}, \dots, x_{m,N-1}]$, for $m \in 1, 2$ represents the two consecutive time slots. Each OFDM symbol is proceeded with a CP with length l_{cp_1} . Assume that l_{cp_1} is not less than the maximum channel memory length l_1 from source node to all antennas of the two relay nodes. The channel impulse responses are assumed to be constant over two consecutive symbols. The received signal at each antenna *i* of each relay node *k*, in two successive OFDM symbol durations after removing the CP at the relay nodes can be expressed as:

$$\mathbf{y}_{(ik)}^m = \mathbf{F} \ \overline{\mathbf{X}}_m + \mathbf{V}_{(ik)}^m \quad \text{for} \quad m \in 1, 2$$
 (5.2.14)

where \mathbf{F} is the channel matrix in time domain, it is a circulant matrix and is expressed as

where n = 0, 1, ..., N-1 is the number of resolvable paths, **0** is a (4×1) zero vector, \mathbf{F}_n^m is a (4×1) vector at path n and time m and it can be written as

$$\mathbf{F}_{(n)}^{m} = \begin{bmatrix} f_{(11),n}^{m} \\ f_{(21),n}^{m} \\ f_{(12),n}^{m} \\ f_{(22),n}^{m} \end{bmatrix}$$

where $f_{(ik),n}^m$ is the channel impulse response between the source node and the *i*-th antenna of each relay node k at time slot m and path n. The mean power of the signal $\mathbf{y}_{(ik)}^m$ at relay nodes is $P_1 + 1$ due to the unit variance assumption of the fading channels from the source node to relay nodes $\mathbf{f}_{(ik)}$ and the additive noise $\mathbf{V}_{(ik)}^m$ at relays. Let P_2 be the average transmission power at each antenna of each relay node, then the optimum power allocation proposed in [105] is used in this scheme, which yields:

$$P_1 = \frac{P}{2}, \quad 2P_2 = \frac{P}{2R}$$

where P is the total transmission power for the whole scheme, and R is the number of relay nodes.

Relaying phase

The two dual-antenna relay nodes will perform EO-STBC on the received noisy signal at each antenna i of each relay node k as shown in Table 5.2. Therefore, the transmitted signals at two consecutive OFDM symbols from all antennas of the two relay nodes are

$$\begin{bmatrix} \mathbf{t}_{11}^{1} & \mathbf{t}_{21}^{1} & \mathbf{t}_{12}^{1} & \mathbf{t}_{22}^{1} \\ \mathbf{t}_{21}^{2} & \mathbf{t}_{21}^{2} & \mathbf{t}_{22}^{2} \end{bmatrix} = \rho \begin{bmatrix} \zeta(\mathbf{y}_{11}^{1}) & -\mathbf{y}_{21}^{*2} & \zeta(\mathbf{y}_{12}^{1}) & -\mathbf{y}_{22}^{*2} \\ \zeta(\mathbf{y}_{11}^{2}) & \mathbf{y}_{21}^{*1} & \zeta(\mathbf{y}_{12}^{2}) & \mathbf{y}_{22}^{*1} \end{bmatrix}$$
(5.2.15)

	Relay1 Ant.1	Relay1 Ant.2	Relay2 Ant.1	Relay2 Ant.2
OFDM1	$\sqrt{\frac{P_2}{P_{1+1}}}\zeta(\mathbf{y}_{11}^1)$	$-\sqrt{rac{P_2}{P_{1+1}}}\mathbf{y}_{21}^{*2}$	$\sqrt{\frac{P_2}{P_{1+1}}}\zeta(\mathbf{y}_{12}^1)$	$-\sqrt{rac{P_2}{P_{1+1}}}\mathbf{y}_{22}^{*2}$
OFDM2	$\sqrt{\frac{P_2}{P1+1}}\zeta(\mathbf{y}_{11}^2)$	$\sqrt{rac{P_2}{P1+1}}\mathbf{y}_{21}^{*1}$	$\sqrt{\frac{P_2}{P1+1}}\zeta(\mathbf{y}_{12}^2)$	$\sqrt{rac{P_2}{P1+1}}\mathbf{y}_{22}^{*1}$

 Table 5.2.
 EO-STBC processing at the relay nodes

where $\rho = \sqrt{\frac{P_2}{P_1+1}}$ is the signal amplification factor scalar [49] [105]. The CP l_{cp_1} is removed and l_{cp_2} is inserted at each antenna of each relay node in the second phase. This process is performed under the assumption that l_{cp_2} is not less than $l_2 + \tau_{max}$, where l_2 is the maximum channel memory length between all antennas of the two relay nodes and the destination node, $\tau_{max} = \max{\{\tau_{(ik)}\}}$ is the maximum overall relative delay from the source to all the receive antennas at the relays and from these antennas to the destination node, and $\tau_{(i1)} = 0$, for $i \in 1, 2$, the receiver still can overcome the effect of ISI and the orthogonality between subcarriers can still be maintained.

The advantage of D-EO-STBC protocol at the relay nodes without the requirement of decoding operation is reducing the complexity at the relay nodes as well as lower power consumption since the relay nodes only implement very simple linear transform operation as shown in (5.2.15).

Implementation at the destination node

The relayed signals arrive at the destination at different time instants, however, timing synchronization is easily performed for the shortest path from the relays to the destination. Without much loss of generality, this will be assumed to be the path from both antennas of R_1 . Hence, the delays for different relayed signals $\tau_{(i2)}$ are assumed relative to the received signal from each antenna of R_1 . Such relative delays will cause ISI between subcarriers. However, because the l_{cp_2} is not less than $l_2 + \tau_{max}$, where l_2 is the maximum channel memory length between all antennas of the two relay nodes and the destination node, $\tau_{max} = \max{\{\tau_{(ik)}\}}$ is the maximum overall relative delay from the source to all the receive antennas at the relays and from these antennas to the destination node, and $\tau_{(i1)} = 0$, for $i \in 1, 2$, the receiver still can overcome the effect of ISI and the orthogonality between subcarriers can still be maintained. Without loss of generality, the delays from both antennas of the second relay R_2 are assumed to be equal due to their vicinity and co-location on the same relay, i.e. $\tau_{(12)} = \tau_{(22)} = \tau_2$.

The delay in the time domain corresponds to a phase change in the frequency domain as follows

$$\mathbf{D}^{\tau_2} = \begin{bmatrix} D_0^{\tau_2}, D_1^{\tau_2}, ..., D_{N-1}^{\tau_2} \end{bmatrix}$$
(5.2.16)

where $D_n^{\tau_2} = e^{-j2\pi n\tau_2/N}$, for n = 0, 1, ..., N - 1. The receiver then removes the CP from each OFDM symbol and performs the FFT operation for the two successive OFDM symbols $\mathbf{Z}_m = [Z_{m,0}, Z_{m,1}, ..., Z_{m,N-1}]$. Consequently, \mathbf{Z}_1 and \mathbf{Z}_2 in the frequency domain can be written as:

$$\mathbf{Z}_{1} = \rho[\mathbf{FFT}(\overline{\mathbf{X}}_{1}) \circ \Lambda_{11} - \mathbf{FFT}(\overline{\mathbf{X}}_{2}^{*}) \circ \Lambda_{21} \\ + \mathbf{FFT}(\overline{\mathbf{X}}_{1}) \circ \Lambda_{12} - \mathbf{FFT}(\overline{\mathbf{X}}_{2}^{*}) \circ \Lambda_{22}] + \mathbf{W}_{1}$$

$$(5.2.17)$$

$$\mathbf{Z}_{2} = \rho[\mathbf{FFT}(\overline{\mathbf{X}}_{2}^{2}) \circ \Lambda_{11} - \mathbf{FFT}(\overline{\mathbf{X}}_{1}^{*1}) \circ \Lambda_{21} \\ + \mathbf{FFT}(\overline{\mathbf{X}}_{2}^{2}) \circ \Lambda_{12} - \mathbf{FFT}(\overline{\mathbf{X}}_{1}^{*1}) \circ \Lambda_{22}] + \mathbf{W}_{2}$$

$$(5.2.18)$$

with

$$\Lambda_{ik} = \begin{cases} \mathbf{FFT}(\mathbf{f}_{(i1)}) \circ \mathbf{FFT}(\mathbf{g}_{(i1)}) & \text{for } \mathbf{k} = 1, \mathbf{i} \in 1, 2\\ \mathbf{D}^{\tau_2} \circ \mathbf{FFT}(\mathbf{f}_{(i2)}) \circ \mathbf{FFT}(\mathbf{g}_{(i2)}) & \text{for } \mathbf{k} = 2, \mathbf{i} \in 1, 2. \end{cases}$$

where the operation \circ is the Hadamard product, i.e., the componentwise product, and $\mathbf{W}_m = \rho(\sum_{m=1}^2 \mathbf{FFT}(\mathbf{V}_{(i1)}^m) \circ \mathbf{g}_{(i1)} + \mathbf{D}^{\tau_2} \circ \mathbf{FFT}(\mathbf{V}_{(i2)}^m) \circ \mathbf{g}_{(i2)}]) + \mathbf{n}_m.$

By taking into consideration the simple identities $(\mathbf{FFT}(X))^* = \mathbf{IFFT}(X^*)$, $\mathbf{FFT}(\zeta(\mathbf{FFT}(X))) = \mathbf{IFFT}(\mathbf{FFT}(X)) = X, \zeta(\mathbf{FFT}(X)) = \mathbf{IFFT}(X)$, $\zeta(\mathbf{IFFT}(X)) = \mathbf{FFT}(X), (\mathbf{IFFT}(X))^* = \mathbf{FFT}(X^*)$, and $\mathbf{FFT}(\zeta(X)) = \zeta(\mathbf{FFT}(X))$, then the equations in (5.2.17) and (5.2.18) can be written to have the EO-STBC structure on each subcarrier $n, 0 \le n \le N - 1$, as follows:

$$\begin{bmatrix} Z_{1,n} \\ Z_{2,n} \end{bmatrix} = \rho \begin{bmatrix} x_{1,n} & -x_{2,n}^* & x_{1,n} & -x_{2,n}^* \\ x_{2,n} & -x_{1,n}^* & x_{2,n} & x_{1,n}^* \end{bmatrix} \begin{bmatrix} \Lambda_{(11),n} \\ \Lambda_{(21),n} \\ \Lambda_{(12),n} \\ \Lambda_{(12),n} \end{bmatrix} + \begin{bmatrix} w_{1,n} \\ w_{2,n} \end{bmatrix}$$
(5.2.19)

The code from (5.2.19) shows that the timing errors only cause phase shift in the channel frequency response, and orthogonality of the STBC still holds. Therefore, the fast symbol-wise ML decoding can be applied at the destination node. By conjugating the received signal at the second independent time interval, for convenience, then we can write

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the equivalent received signals at the destination at each subcarrier n, for $0 \le n \le N - 1$ as follows:

$$\begin{bmatrix} Z_{1,n} \\ Z_{2,n}^* \end{bmatrix} = \rho \begin{bmatrix} \Lambda_{(11),n} + \Lambda_{(12),n} & -\Lambda_{(21),n} - \Lambda_{(22),n} \\ \Lambda_{(21),n}^* + \Lambda_{(22),n}^* & \Lambda_{(11),n}^* + \Lambda_{(12),n}^* \end{bmatrix} \begin{bmatrix} x_{1,n} \\ x_{2,n}^* \end{bmatrix}$$
(5.2.20)
$$+ \begin{bmatrix} w_{1,n} \\ w_{2,n}^* \end{bmatrix}$$

From (5.2.20), the equivalent channel matrix corresponding to the openloop EO-STBC can be represented as:

$$\mathbf{H}_{n}^{OL} = \begin{bmatrix} \Lambda_{(11),n} + \Lambda_{(12),n} & -\Lambda_{(21),n} - \Lambda_{(22),n} \\ \Lambda_{(21),n}^{*} + \Lambda_{(22),n}^{*} & \Lambda_{(11),n}^{*} + \Lambda_{(12),n}^{*} \end{bmatrix}$$
(5.2.21)

Assuming that perfect CSI is available at the destination receiver and matched filtering is performed with the equivalent channel matrix (5.2.21), so the resultant Grammain matrix \mathbf{G}_n for each of n subcarriers can be obtained as follows

$$\mathbf{G}_{n} = (\mathbf{H}_{n}^{OL})^{H} \mathbf{H}_{n}^{OL} = \begin{bmatrix} \gamma_{n} & 0\\ 0 & \gamma_{n} \end{bmatrix}$$
(5.2.22)

where γ_n represents the channel gain, such that $\gamma_n = \alpha_n + \beta_n$, and

$$\alpha_n = \sum_{k=1}^{2} \left(|\Lambda_{(1k),n}|^2 + |\Lambda_{(2k),n}|^2 \right)$$
 (5.2.23)

$$\beta_n = 2\Re(\Lambda^*_{(11),n}\Lambda_{(21),n} + \Lambda_{(12),n}\Lambda^*_{(22),n})$$
(5.2.24)

where α_n is the cooperative diversity gain and βn is an interference factor due to the correlation between channel coefficients. The average value of the channel gain can therefore be expressed as

$$E(\gamma_n) = E(\alpha_n + \beta_n) = 4 + E(\beta_n) \tag{5.2.25}$$

As such, the β_n term may reduce the channel gain and correspondingly the SNR where the relation between channel gain and SNR can be expressed as

$$SNR = \gamma_n \frac{\sigma_x^2}{\sigma_\omega^2} = (|\Lambda_{(1k),n}|^2 + |\Lambda_{(2k),n}|^2) \frac{\sigma_x^2}{\sigma_\omega^2}$$
(5.2.26)

where $\frac{\sigma_x^2}{\sigma_\omega^2} = \frac{P_1 P_2}{P_1 + 4P_2 + 1}$ is the ratio of the total transmission power of the desired signal to the noise power at the receiver.

5.2.6 One-bit per subcarrier feedback scheme based on cyclic rotation

As can be observed from (5.2.26), it is clear that the βn term may reduce the system performance. In order to achieve full cooperative diversity and array gain and hence maximizing the system performance, a one-bit per subcarrier feedback scheme is proposed to modify the transmitted signals from all antennas of the two relay nodes as shown in Figure (5.7).

This CL-EO-STBC scheme is based on selection of cyclic rotation of $(\theta_{1,n}, \theta_{2,n})$ phase angles for each subcarrier. Only two choices of angles effectively makes the β_n term positive by multiplying them with the codeword matrix before they are transmitted from the *i*-th antenna of each relay node.



Figure 5.7. Basic structure of an asynchronous wireless relay network with two antennas on each relay node and single antennas on source and destination nodes. The network implements a cyclic rotation feedback at relay nodes.

In order to apply the weighted values $U_{1,n}$ and $U_{2,n}$ over each subcarrier n, all relay nodes require to perform an N-point OFDM transformation (**FFT/IFFT**) operation to manage to change the phase rotation of their signals to the proper phase as shown in Figure 5.8.

Therefore, the received signals in (5.2.14) at each antenna *i* of each



Figure 5.8. The baseband model of OFDM at one dual-antenna relay node.

relay node k, in two successive OFDM symbol durations after removing the CP at the relay nodes can be expressed as:

$$\mathbf{y}_{(ik)}^{m} = \mathbf{FFT}(\overline{\mathbf{X}}_{m} \circ \mathbf{f}_{(ik)} + \mathbf{V}_{(ik)}^{m}) \quad \text{for} \quad m, i, k \in 1, 2 \quad (5.2.27)$$

Accordingly, the encoding relay matrix in (5.2.2) will be

$$\mathbf{C}^{CL} = \rho \begin{bmatrix} \mathbf{IFFT}(\mathbf{u}_1 \circ \zeta(\mathbf{y}_{11}^1)) & -\mathbf{IFFT}(\mathbf{u}_2 \circ \mathbf{y}_{21}^{*2}) & \mathbf{IFFT}(\mathbf{u}_1 \circ \zeta(\mathbf{y}_{12}^1)) & -\mathbf{IFFT}(\mathbf{u}_1 \circ \mathbf{y}_{22}^{*2}) \\ \mathbf{IFFT}(\mathbf{u}_2 \circ \zeta(\mathbf{y}_{11}^2)) & \mathbf{IFFT}(\mathbf{u}_1 \circ \mathbf{y}_{21}^{*1}) & \mathbf{IFFT}(\mathbf{u}_2 \circ \zeta(\mathbf{y}_{12}^2)) & \mathbf{IFFT}(\mathbf{u}_2 \circ \mathbf{y}_{22}^{*1}) \\ \end{bmatrix}$$

where \mathbf{C}^{CL} denotes the encoding relay matrix for the broadband system. Therefore, the process over relay nodes will include **FFT/IFFT** operation over the two relays in addition to time-reversal and complex conjugation operations. However, after the four relay nodes process their received signals accordingly and forward them towards the destination node.

At the destination node, the received signals are subjected to the same procedure as in the open-loop scheme which yields the following equivalent channel matrix,

$$\mathbf{H}_{n}^{CL} = \begin{bmatrix} \Lambda_{(11),n}U_{1,n} + \Lambda_{(12),n}U_{2,n} & -\Lambda_{(21),n}U_{1,n} - \Lambda_{(22),n}U_{1,n} \\ \Lambda_{(21),n}^{*}U_{2,n} + \Lambda_{(22),n}^{*}U_{1,n} & \Lambda_{(11),n}^{*}U_{2,n} + \Lambda_{(12),n}^{*}U_{1,n} \end{bmatrix}$$
(5.2.28)

where $U_{k,n} = e^{j\theta_{k,n}}$, and k=1,2. After applying the feedback, the term β_n in (5.2.24) will be equal to

$$\beta_n = 2\Re(\Lambda^*_{(11),n}\Lambda_{(21),n} + \Lambda_{(12),n}\Lambda^*_{(22),n})e^{j(\theta_{2,n}-\theta_{1,n})}$$
(5.2.29)

It is clear that if the term $\beta_n > 0$ in (5.2.29), the designed feedback scheme will provide full cooperative diversity gain of $(2R)(1 - \frac{\log \log P}{\log P})$ [14]. Assuming full CSI at the destination node, the design criterion of one-bit/subcarrier feedback scheme over the frequency selective fading channels can be designed by adopting a one bit of b = 0 or b = 1 to be fedback to each subcarrier to indicate whether

$$\Re(\Lambda^*_{(11),n}\Lambda_{(21),n} + \Lambda_{(12),n}\Lambda^*_{(22),n}) \ge 0, \quad b = 0$$
 is fedback

or

$$\Re(\Lambda^*_{(11),n}\Lambda_{(21),n} + \Lambda_{(12),n}\Lambda^*_{(22),n}) < 0, \quad b = 1$$
 is fedback

Then this one-bit information will be fedback to each subcarrier of the i-th antenna of each relay node. At the relay nodes, we first judge the

value of b: if b=0, then $\theta_{n,1}=\theta_{n,2}=0$ or $\theta_{n,1}=\theta_{n,2}=\pi$ which ensures that $e^{j(\theta_{2,n}-\theta_{1,n})}=1$, on the other hand if b=1, then $\theta_{n,1}=\pi$ and $\theta_{n,2}=0$ or $\theta_{n,1}=0$ and $\theta_{n,2}=\pi$ which ensures that $e^{j(\theta_{2,n}-\theta_{1,n})}=-1$.

The solution for the proposed closed-loop scheme can be expressed as follows

$$b = \begin{cases} 0, & \Re(\Lambda_{(11),n}^* \Lambda_{(21),n} + \Lambda_{(12),n} \Lambda_{(22),n}^*) \ge 0 \\ & \text{set } \theta_{n,1} = \theta_{n,2} = 0 \text{ or } \theta_{n,1} = \theta_{n,2} = \pi \end{cases}$$
$$1, & \Re(\Lambda_{(11),n}^* \Lambda_{(21),n} + \Lambda_{(12),n} \Lambda_{(22),n}^*) < 0 \\ & \text{set } \theta_{n,1} = \pi \ \theta_{n,2} = 0 \text{ or } \theta_{n,1} = 0 \ \theta_{n,2} = \pi \end{cases}$$

It can be seen that the proposed scheme feeds back just one-bit per subcarrier to maximize the value of the β_n term which leads to an improved channel gain and correspondingly improving the SNR at destination node.

At the destination node, the matched filter is performed by pre-multiplying \mathbf{H}_{n}^{CL} in each subcarrier with $\mathbf{Z}_{n} = [Z_{1,n}, Z_{2,n}^{*}]$ to detect the vector $\hat{\mathbf{x}}_{n} = [\hat{s}_{1,n}, \hat{s}_{2,n}]$ as follows

$$(\mathbf{H}_{n}^{CL})^{H}\mathbf{Z}_{n} = (\mathbf{H}_{n}^{CL})^{H}\mathbf{H}_{n}^{CL} \begin{bmatrix} x_{1,n} \\ x_{2,n} \end{bmatrix} + (\mathbf{H}_{n}^{CL})^{H}\mathbf{W}_{n}$$
$$\hat{\mathbf{x}}_{n} = \Delta_{n} \begin{bmatrix} x_{1,n} \\ x_{2,n} \end{bmatrix} + \mathbf{W}_{n}$$
(5.2.30)

where $\mathbf{W}_n = [w_{1,n}, w_{2,n}^*]^T$ is the total additive Gaussian noise vector at the relay nodes and the destination node. Applying least squares (LS) detection to obtain the detection results $\hat{\mathbf{s}}_{k,n} = [\hat{s}_{1,n}, \hat{s}_{2,n}]$

$$\begin{bmatrix} \hat{s}_{1,n} \\ \hat{s}_{2,n} \end{bmatrix} = \arg \min_{x_{k,n} \in S} \left\| \begin{bmatrix} \hat{x}_{1,n} \\ \hat{x}_{2,n} \end{bmatrix} - \mathbf{H}_n^{CL} \begin{bmatrix} x_{1,n} \\ x_{2,n} \end{bmatrix} \right\|$$
(5.2.31)

where S is the alphabet containing 4 for QPSK symbols.

5.2.7 Simulation results

In this section, simulation results show the performance of the proposed asynchronous M-D-QO-STBC and D-EO-STBC schemes using QPSK modulation. For M-D-QO-STBC scheme simulation comparisons, the average end-to-end codeword error rate (CWER) is calculated. On the other hand, for D-EO-STBC scheme simulation comparisons, the end-to-end frame error (FER) performance against the average transmitted power at the source node P_1 in dB over quasi-static frequency selective 11 taps channels is verified using simulation and compared with the previous feedback methods in wireless relay networks. In both schemes, it is assumed that the length of data block N=64 with CP lengths l_{cp} 16. The delay between relay nodes is chosen between 0 to 6 with uniform distribution. The simulation assumes perfect CSI knowledge at the destination node. The transmitted power is fixed and identical for all schemes used in the comparison. Figure 5.9 shows that the OFDM-based scheme can significantly mitigate the delay when $\tau < CP$ which enhances the performance of the system under imperfect synchronization. In addition the simulations show that the OFDMbased M-D-QO-STBC scheme provides a better performance over the OFDM-based conventional D-QO-STBC [28] due to achieving full spatial diversity and the improvement in the code gain distance (CGD) which was obtained from the maximization of the distance properties of the original codebooks of the modified code. In Figure 5.10, the



Figure 5.9. Performance of OFDM-based M-D-QO-STBC for one-way asynchronous cooperative relay network with relay selection technique using QPSK signal.

performance of the proposed un-coded D-EO-STBC-OFDM for two relay nodes each equipped with two antennas is shown. It can be seen that the proposed D-EO-STBC-OFDM scheme with one-bit feedback is better than the closed-loop D-EO-STBC-OFDM in [30] as shown in Figure 5.10.and Figure 5.11. In particular it is clear that the un-coded proposed scheme, when FER is 10^{-3} , provides a power saving of approximately 8 dB as compared to the un-coded open-loop scheme and approximately 4 dB as compared to the previous feedback scheme [30] in wireless relay networks respectively. It is shown that the proposed closed-loop D-EO-SFBC-OFDM with one-bit/subcarrier feedback can get the conventional diversity gain equal to the number of transmit antennas co-located in each relay and more additional array gain (feedback gain) at the same time. Moreover, the distributed



Figure 5.10. The un-coded FER performance as a function of total transmit power of the proposed scheme and with previous feedback scheme in two relay nodes with two antennas in each relay for a wireless network.

Alamouti (D-Alamouti) OFDM using one relay equipped with two antenna is included as a reference. The proposed scheme achieves a frame error probability of 10^{-3} , it provides a power saving of about 13 dB improvement as compared to D-Alamouti-OFDM. Figure 5.11 demonstrates the comparison of the purposed scheme when combined with outer convolutive coding to exploit coding gain with the previous feedback scheme in distributed manner. It can be seen that at FER= 10^{-3} the one-bit/subcarrier feedback scheme, it can achieve a power saving approximately 2.5 dB as improvement as compered to the previous coded closed-loop scheme [30].



Figure 5.11. The coded FER performance as a function of total transmit power of the proposed scheme and with previous feedback scheme in two relay nodes with two antennas in each relay for a wireless network.

5.3 Summary

In this chapter, the M-D-QO-STBC scheme for asynchronous cooperative relay networks was proposed. An OFDM data structure was employed with cyclic prefix (CP) insertion at the source in the one-way cooperative relay network to combat the effects of time asynchronicity at the relay nodes. The simulation results indicated that the OFDMbased scheme can significantly mitigate the delay when $\tau < CP$ which enhances the performance of the system under imperfect synchronization. Furthermore, we presented a transmission scheme which utilizes closed-loop EO-STBC OFDM based on selection cyclic rotation of phase angle feedback for asynchronous cooperative relay networks to provide high data wireless communication over frequency selective fading channels. OFDM type transmission is used to mitigate the ISI from multipath fading and timing error between relay nodes at the destination node. Furthermore, the proposed feedback scheme is limited to only one-bit feedback per subcarrier based on selection cyclic phase rotation to i - th antenna of each relay node. The received signals at the destination node have the Alamouti code structure on each subcarrier, therefore the proposed scheme utilizes symbol-wise ML decoders and can achieve 4th-order diversity using two relay nodes each co-located with two antennas. From simulation results it is clear that the proposed feedback scheme outperform previous closed-loop method in cooperative wireless relay networks.

CLOSED-LOOP EXTENDED ORTHOGONAL SPACE-FREQUENCY CODING FOR COGNITIVE WIRELESS RELAY NETWORKS

In this chapter, closed-loop extended orthogonal space-frequency-block coding (CL-EO-SFBC) is proposed for use within cognitive wireless relay networks. To exploit the available spectrum opportunities within the EO-SFBC, a spectrum indicator matrix is used in the development of the relay transmission scheme. The approach has the flexibility to dynamically adapt the code matrix to the number of available relays. The performance of the scheme is evaluated in terms of end-to-end frame error rate. Furthermore, the probability density function of the multi-path links is modeled in the time domain with an Erlang distribution function. The analytical expressions for the probability density function and cumulative density function of the end-to-end signal-tonoise ratio are obtained for one and two cognitive relay nodes and multi-path channel lengths of two and three. Moreover, expressions for outage probability are determined for a perfect and imperfect spectrum acquisition. Finally, the theoretical results are compared with simulations to confirm the validity of the analysis

6.1 Introduction

The continuous growth in wireless technologies requires more spectrum resources to be exploited [23]. However, these resources are limited and valuable and controlled by specialized governmental bodies, by which licensed users (primary users) have the right to freely utilize licensed spectrum allocated to them. However, recent studies have shown that the licensed spectrum is most of the time underutilized by the licensed (primary) users [116]. The utilization of spectrum can be more improved by allowing unlicensed (secondary) users to share the spectrum with primary users whenever it is not utilized.

In Haykin's paper [39], it was stated that "cognitive radio is an intelligent wireless communication system that is aware of its surrounding environment (i.e., its outside world), and uses the methodology of understanding-by-building to learn from the environment and adapt its internal states to statistical variations in the incoming radio frequency (RF) stimuli by making corresponding changes in certain operating parameters". On the other hand, relaying is known to be an effective technique to improve signal quality and reliability at the destination [24], [117]. This improvement is called cooperative diversity due to the fact that it is achieved by having different users in the wireless network cooperate in some manner. The inclusion of cognitive radio technology in cooperative communications, by means of equipping the intermediate relay nodes with cognitive radios, will provide a capability of spectrum sharing among primary and secondary users in addition to the other benefits of relaying mentioned hereinabove. They may act as spectrum monitors (sensing) as well as spectrum access coordinators (sharing).

In this chapter, cognitive technology over the intermediate wireless relay nodes is deployed in order to achieve reliable communication and coordinate the spectrum sharing among primary and secondary users. The term "cognitive relays" will therefore denote such relay nodes hereinafter. In [23], the application of the cooperative communications approach was considered for cooperative spectrum sensing and sharing. The spectrum sensing side is considered to be practically possible and it is not the main concern of this chapter, however, the focus will be on fully utilizing the dynamic spectrum through a number of cognitive relay nodes in order to achieve reliable communication and seamless transmission without causing any interference to primary users when they are present.

In order to support a high rate full diversity transmission with simple detection, this chapter extends the proposed space-frequency (SF) coding scheme introduced in [23] through the use of the closed-loop extended orthogonal space-time block coding (CL-EO-STBC) technique in [30] with the assumption that the minimum number of cognitive relays present in the surrounding area is four so that a seamless cooperative transmission is guaranteed over two (Alamouti), three or four active cognitive relays regardless of the presence of PU_s during the transmission process. The CL-EO-SFBC ensures the achievement of full spatial diversity when all the four cognitive relays are in operation. Furthermore, In this work, the outage probability of the proposed cognitive relay network is considered when there are one and two active relays with perfect as well as imperfect spectrum acquisition.

6.2 Transmission Model

The cooperative relay nodes in cognitive radio networks are intended to coordinate sharing of the spectrum utilization among primary and secondary users and bridging the channel availability from the source to the destination node based on prior knowledge of the surrounding spectrum environment. The cognitive relays have the ability to adapt dynamically to this local spectrum environment where each of the cognitive relays first obtains the spectrum map of its local channel environment after every spectrum sensing operation. Following the approach in [23], let $\mathbf{b}_i = (b_{i,1}, b_{i,2}, \dots, b_{i,N})$ denote the spectrum indicator of the ith cognitive relay, where the entries $b_{i,k}$, k = 1, 2, ..., N, denote the availability for the frequency bands $f_1, f_2, ..., f_N$, respectively. The parameters $b_{i,k} \in \{0,1\}$, where a binary 1 indicates that the frequency band f_k is not being utilized by the PU and it is available for cognitive relay i, whereas a binary 0 indicates that access to that frequency band is not allowed to the cognitive relay i and the PU is utilizing this frequency band. Obviously, the spectrum sensed environment of the whole cognitive relay network can be characterised by the following matrix:

$$\mathbf{B} = \begin{pmatrix} b_{1,1} & b_{1,2} & \cdots & b_{1,N} \\ b_{2,1} & b_{2,2} & \cdots & b_{2,N} \\ \vdots & \vdots & \ddots & \vdots \\ b_{m,1} & b_{m,2} & \cdots & b_{m,N} \end{pmatrix}$$
(6.2.1)

$$\downarrow relay(space) \longrightarrow band(frequency)$$

The binary matrix **B** represents the spectrum availability in the cognitive relay network. The cognitive SF coding technique was utilized to fully exploit the spectrum opportunities in cognitive relay networks while supporting high-rate cooperative transmission. The cooperative transmission needs two stages to convey the data from the source node to the destination node. In the first stage, the source broadcasts to all the cognitive relay nodes a block of N symbols and then stops sending. During the second stage cognitive relays decode the received signal and obtain the spectrum map of their local spectrum environment before encoding the signal as per the SF coding structure. At the destination, the received signal block on the k^{th} band is given by

$$z_k = \sum_{i=1}^m b_{i,k}(x_{i,k}h_{R_{i,k},D} + n_{i,k})$$
(6.2.2)

where $x_{i,k}$ is the coded signal block sent from band k of cognitive relay i, and $n_{i,k}$ is the additive noise vector with zero-mean complex Gaussian random variable entries. This can be rewritten as

$$z_k = c_k h_k + n_k \tag{6.2.3}$$

where

$$c_k = \left[\begin{array}{ccc} b_{1,k} x_{1,k} & b_{2,k} x_{2,k} \cdots & b_{m,k} x_{m,k} \end{array} \right]$$
(6.2.4)

and

$$h_k = \begin{bmatrix} h_{R_{1,k},D} & h_{R_{2,k},D} \cdots & h_{R_{m,k},D} \end{bmatrix}^T$$
 (6.2.5)

where $(.)^T$ denotes the transpose operation. Combining the received signal over all channels will give

$$z = ch + n$$

The proposed EO-SF code structure is designed from the simple SF code introduced in [23] as follows:

$$\mathbf{C}_{eo} = \mathbf{I}_N \otimes \mathbf{A}_{eo} \tag{6.2.6}$$

where \mathbf{A}_{eo} is the conventional extended orthogonal SFBC matrix

$$\mathbf{A}_{eo} = \begin{bmatrix} x_{1,k} & x_{1,k} & x_{2,k} & x_{2,k} \\ -x_{2,k}^* & -x_{2,k}^* & x_{1,k}^* & x_{1,k}^* \end{bmatrix}$$
(6.2.7)

where $(.)^*$ is the complex conjugation operation.

This cognitive EO-SFC has a flexible code structure that can be distributed over the available cognitive relays after every sensing activity is performed such that when only two relays are active (Alamouti):

$$\mathbf{A}_{2} = \begin{bmatrix} x_{1,k} & 0 & x_{2,k} & 0 \\ \\ -x_{2,k}^{*} & 0 & x_{1,k}^{*} & 0 \end{bmatrix}$$

or

$$\mathbf{A}_{2} = \begin{bmatrix} 0 & x_{1,k} & 0 & x_{2,k} \\ 0 & -x_{2,k}^{*} & 0 & x_{1,k}^{*} \end{bmatrix}$$
(6.2.8)

are employed, whereas, in the case of three active cognitive relays then,

$$\mathbf{A}_{3} = \begin{bmatrix} x_{1,k} & x_{1,k} & x_{2,k} & 0 \\ -x_{2,k}^{*} & -x_{2,k}^{*} & x_{1,k}^{*} & 0 \end{bmatrix}$$

or

$$\mathbf{A}_{3} = \begin{bmatrix} 0 & x_{1,k} & x_{2,k} & x_{2,k} \\ 0 & -x_{2,k}^{*} & x_{1,k}^{*} & x_{1,k}^{*} \end{bmatrix}$$

or

$$\mathbf{A}_{3} = \begin{bmatrix} x_{1,k} & 0 & x_{2,k} & x_{2,k} \\ -x_{2,k}^{*} & 0 & x_{1,k}^{*} & x_{1,k}^{*} \end{bmatrix}$$
(6.2.9)

or

$$\mathbf{A}_{3} = \begin{bmatrix} x_{1,k} & x_{1,k} & 0 & x_{2,k} \\ -x_{2,k}^{*} & -x_{2,k}^{*} & 0 & x_{1,k}^{*} \end{bmatrix}$$
(6.2.10)

are employed.

The phase rotation angles are multiplied by the transmitted symbols from cognitive relays 1 and 3 or 2 and 4 when three or four cognitive relays are active in order to achieve full diversity and array gain in the case of four active cognitive relays, and enhance the diversity in the case of three active cognitive relays.

6.3 The Distributed Transmission Protocol

The cognitive wireless communication relay network is considered to be a half-duplex cognitive relay network consisting of a source node (S), a destination node (D) and a set of m intermediate relay nodes



Figure 6.1. Cognitive cooperative wireless relay network architecture (phase 1)

 $R_i, i \in \{1, 2, 3, 4\}$, acting as CR_s , and distributed over a large geographical area where there exists several PU_s in the vicinity operating over a wide-band spectrum. The source node, destination node as well as each cognitive relay node (R) are equipped with one antenna. It is assumed that each cognitive relay node is within the transmission range of one PU node. It is also considered that the communication channels are quasi-static frequency selective fading channels. In order to combat this frequency selective fading, orthogonal frequency division multiplexing (OFDM) with cyclic prefix (CP) is implemented at the source node. Also OFDM-based decode-and forward (DF) relaying scheme is adopted at the cognitive relays. It is assumed that there is no direct link between the source and destination nodes, and there is perfect channel information $g_{i,k}$ at the relays, moreover, the receiver has perfect channel information $h_{i,k}$, where $g_{i,k}$ and $h_{i,k}$ denote the channels from the kth path from the transmitter to the *i*th relay and from the *i*th relay to the kth path to the receiver respectively, note that all channels are frequency selective. In the first communication phase, Figure 6.1, the source transmits the signal vector $x = [x_{1,k}, ..., x_{m,k}]^T$ to the m number of cognitive relays. In the DF scheme, the cognitive relays are assumed to correctly decode the transmitted signal awaiting the retransmission phase to the destination.

Prior to commencing the second phase shown in Figure 6.2. and retransmitting the signal to the destination, each cognitive relay should first obtain the spectrum map of its channel environment, this can be a practical overhead [23], however, it is assumed that the channels are sufficiently static so that this map remains constant for each frame.

6.3.1 Source Node Implementation

In this phase, the source node transmits an OFDM block representing a set of m modulated complex symbols $\mathbf{x} = [x_{1,k}, x_{2,k}, ..., x_{m,k}]^T$. Before broadcasting the signal to the cognitive relays, the IFFT of the symbols is taken and a sufficiently long cyclic prefix is inserted to overcome the inter-block interference. In this scheme, the optimum power allocation is implemented as in [27], so that the source uses half the total power and the active cognitive relays share the remaining half. If the total power is denoted with P, then



Figure 6.2. Cognitive Cooperative wireless relay network architecture (phase 2)

$$P_1 = \frac{P}{2} \quad \text{and} \quad P_2 = \frac{P}{2R} \tag{6.3.1}$$

The received signal at the relays is to be decoded and encoded before being relayed to the destination.

6.3.2 Cognitive Relay and Destination Nodes Implementation

During the second phase, the decoded received signal at the relays is encoded as per the SF coding technique so that the code **A** is distributed over the relays as follows

$$\mathbf{A} = \begin{bmatrix} x_{1,k} & x_{1,k} & x_{2,k} & x_{2,k} \\ -x_{2,k}^* & -x_{2,k}^* & x_{1,k}^* & x_{1,k}^* \end{bmatrix}$$
(6.3.2)

The cognitive receiver, in turn, being aware of the spectrum map represented in the spectrum indicator matrix, in addition to its perfect knowledge of the channels between the cognitive relay and the destination, will accordingly calculate the channels available as expressed below:

$$\left[\begin{array}{cccc} b_{1,k}h_{1,k} & b_{2,k}h_{2,k} & b_{3,k}h_{3,k} & b_{4,k}h_{4,k} \end{array}\right]$$

From [30], it is known that the corresponding equivalent channel matrix for the open-loop-EO-STBC is given by:

$$\mathbf{H} = \begin{bmatrix} h_{1,k} + h_{2,k} & h_{3,k} + h_{4,k} \\ h_{3,k}^* + h_{4,k}^* & -h_{1,k}^* - h_{2,k}^* \end{bmatrix}$$
(6.3.3)

Assuming the full diversity case, that is when all the frequency bands in (1) are available for the cognitive relays, i.e. $\forall b_s = 1$, then the data symbols from the antenna of R_1 are rotated by phase angle $U_1 = e^{j\Theta_{1,k}}$ and the data symbols from the antenna of R_3 are multiplied by phase rotation angle $U_2 = e^{j\Theta_{2,k}}$ while the other two antennas related to R_2 and R_4 remain unchanged. Therefore, the closed loop code will be:

$$\mathbf{A}_{CL(\forall b_s=1)} = \begin{bmatrix} x_{1,k}e^{j\Theta_{1,k}} & x_{1,k} & x_{2,n}e^{j\Theta_{2,k}} & x_{2,k} \\ -x_{2,k}^*e^{j\Theta_{1,k}} & -x_{2,k}^* & x_{1,k}^*e^{j\Theta_{2,k}} & x_{1,n}^* \end{bmatrix}$$
(6.3.4)

Phase rotation performed on the transmitted symbols is similar to rotating phases of the corresponding channel coefficients [30]. Hence, the received signals z_1 and z_2 at the two independent time slots can be expressed as follows:

$$\begin{bmatrix} z_1 \\ z_2^* \end{bmatrix} = \begin{bmatrix} U_1(h_1b_1) + h_2b_2 & U_2(h_3b_3) + h_4b_4 \\ (U_3(h_3b_3)^* + h_4^*b_4) & -(U_1(h_1b_1))^* - h_2^*b_2 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}$$
(6.3.5)

where n_1 and n_2 both $\varepsilon CN(0, \sigma_n^2)$, are additive Gaussian noise. Since $|U_1|^2 = |U_2|^2 = 1$, then the corresponding channel gain G is given in [30] by:

$$G = \sum_{i=1}^{4} |h_i|^2 + 2Re\{(h_1h_2^*e^{j\theta_1} + h_3h_4^*e^{j\theta_2})\}$$
(6.3.6)

where the rotation angles θ_1 and θ_2 are obtained by maximizing the interference factor which leads to maximizing the SNR accordingly, and are expressed as follows:

$$\theta_1 = -angle(h_1h_2^*)$$
 and $\theta_2 = -angle(h_2h_3^*)$

In the next section, expressions for outage probability are determined for a perfect and imperfect spectrum acquisition.

6.4 The Outage Probability Analysis

The system model discussed above is considered. The channel impulse responses (CIRs) from the source node S to the relay nodes R_i with channel length N, are $g_{i,N} = [g_{i,1}, g_{i,2}, \ldots, g_{i,N}]$ and from R_i to the destination node D with the same channel length are $h_{i,N} =$ $[h_{i,1}, h_{i,2}, \ldots, h_{i,N}]$. These channel coefficients are assumed to represent quasi-static multi-path channels. Using the signal-to-noise ratio E_s/N_0 , then $\gamma_{(sr_i,N)} = E_s |g_{i,N}|^2/N_0$ and $\gamma_{(r_id,N)} = E_s |h_{i,N}|^2/N_0$ are the instantaneous received SNR at the i^{th} relay from the source and from the i^{th} relay to the destination, respectively, and $\bar{\gamma}_{(sr_i,N)} = \sigma_{(sr_i,N)}^2 E_s/N_0$ and $\bar{\gamma}_{(r_id,N)} = \sigma_{(r_id,N)}^2 E_s/N_0$ are the average received SNR at the i^{th} relay and from the i^{th} relay to the destination respectively. Then, the average end-to-end SNR at the destination is given by [118]

$$\bar{\gamma}_{D} = \frac{\bar{\gamma}_{sr_{i,N}}\bar{\gamma}_{r_{i,N}d}}{\bar{\gamma}_{sr_{i,N}} + \bar{\gamma}_{r_{i,N}d}}$$
(6.4.1)

The outage probability is defined as when the average end-to-end SNR falls below a certain predefined threshold value, i.e. $\bar{\gamma} < 2^{2B} - 1$, where B is the target rate [119], [34]. The outage probability can be expressed as

$$P_{out} = \int_0^{\gamma} f_{\gamma}(\gamma) d\gamma = F_{\gamma}(\gamma) \tag{6.4.2}$$

where $f_{\gamma}(\gamma)$ is the probability density function (PDF) and $F_{\gamma}(\gamma)$ is the cumulative distribution function (CDF) of the SNR. For frequency selective Rayleigh fading channels, the PDF and CDF of the sum of N independent paths of each channel can be modeled as an Erlang distribution, with $u \in SR_i, R_iD$ links, which is given by

$$f_{\gamma u}(\gamma) = \frac{\gamma^{N-1} e^{-\frac{\gamma}{\bar{\gamma}}}}{\Gamma(N)\bar{\gamma}^N}$$
(6.4.3)

where $\Gamma(N) = \int_0^\infty s^{N-1} e^{-s} ds = (N-1)!$ is termed the complete Gamma function, N is called the shape parameter which represents the number of paths in each channel, and $\bar{\gamma}$ is called the scale parameter and is

denoted as the average SNR. Equation (6.4.3) has the same distribution as the sum of N independent exponential distribution random variables with circular complex normal distribution $CN(\mu, \sigma^2)$ where $\mu = N\bar{\gamma}$ and $\sigma^2 = N\bar{\gamma}^2$ are the mean and the variance of the Erlang distribution, respectively. This shows that the PDF of the sum squared coefficients of the frequency selective channel is an Erlang distribution with different numbers of paths N so when N = 1 the PDF is exponentially distributed which is equivalent to the flat fading channel [120]. Thus, the CDF can be obtained by taking the integral of the PDF in (6.4.3) with respect to γ , yielding (6.4.4)

$$F_{\gamma u}(\gamma) = 1 - e^{-\frac{\gamma}{\bar{\gamma}}} \sum_{k=0}^{N-1} \frac{\gamma^k}{k! \, \bar{\gamma}^k} \tag{6.4.4}$$

where $\bar{\gamma}$ denotes the mean SNR for all links. In order to calculate the outage probability, similarly to [121], the following upper-bound on the SNR is employed

$$\gamma_i = \min(\gamma_{sr_{i,N}}, \gamma_{r_{i,N}d}) > \gamma_D \tag{6.4.5}$$

and therefore, the CDF of $min(\gamma_{sr_{i,N}}, \gamma_{r_{i,N}d})$ can be expressed as

$$F_{\gamma_i}(\gamma) = 1 - P_r(\gamma_{sr_{i,N}} > \gamma) P_r(\gamma_{r_{i,N}d} > \gamma)$$

$$= 1 - [1 - P_r(\gamma_{sr_{i,N}} \le \gamma)] [1 - P_r(\gamma_{r_{i,N}d} \le \gamma)]$$
(6.4.6)

By substituting (6.4.4) into (6.4.6), the CDF of the end-to-end SNR can be expressed as

$$F_{\gamma_i}^N(\gamma) = 1 - e^{-\frac{2\gamma}{\bar{\gamma}}} \left[\sum_{k=0}^{N-1} \frac{\gamma^k}{k! \, \bar{\gamma}^k} \right] \left[\sum_{\bar{k}=0}^{N-1} \frac{\gamma^{\bar{k}}}{\bar{k}! \bar{\gamma}^{\bar{k}}} \right]$$
(6.4.7)

a superscript N to the left hand side is added to denote the channel length of a link. For algebraic convenience, simplifying (6.4.7) by limiting the channel length to N = 2 and N = 3, yields

$$F_{\gamma_i}^2(\gamma) = 1 - e^{-\frac{2\gamma}{\bar{\gamma}}} - \frac{2e^{-\frac{2\gamma}{\bar{\gamma}}}\gamma}{\bar{\gamma}} - \frac{2e^{-\frac{2\gamma}{\bar{\gamma}}}\gamma^2}{\bar{\gamma}^2}$$
(6.4.8)

and

$$F_{\gamma_i}^3(\gamma) = 1 - e^{-\frac{2\gamma}{\bar{\gamma}}} - \frac{2e^{-\frac{2\gamma}{\bar{\gamma}}}\gamma}{\bar{\gamma}} - \frac{2e^{-\frac{2\gamma}{\bar{\gamma}}}\gamma^2}{\bar{\gamma}^2} - \frac{2e^{-\frac{2\gamma}{\bar{\gamma}}}\gamma^3}{\bar{\gamma}^3} - \frac{2e^{-\frac{2\gamma}{\bar{\gamma}}}\gamma^4}{\bar{\gamma}^4} \quad (6.4.9)$$

The outage probability analysis of the two relay scheme will be considered in the next section.

6.4.1 Outage Probability Analysis of the Two-Relay Scheme

In this case, the number of available cognitive relays is two. It is assumed that $\gamma_{sr_{i,N}} = x$ and is greater than $\gamma_{r_{i,N}d} = y$, therefore the joint distribution of these ordered values can be expressed [122] as

$$f_{X,Y}^N(x,y) = M(M-1)f_X^N(x)f_Y^N(y)\left[F_Y^N(y)\right]^{M-2}$$
(6.4.10)

Equation (6.4.10) represents the general PDF form of the multi-path environment transmission with N paths per link and M relays. To simplify (6.4.10), in this work, the channel length is limited to be either N = 2 or 3, and the number of relays (M) is two, then (6.4.10) can be rewritten as either:

$$\begin{split} f_{X,Y}^2(x,y) &= 2(2-1) \left[\frac{2e^{-\frac{2x}{\bar{\gamma}}}x}{\bar{\gamma}^2} + \frac{2e^{-\frac{2x}{\bar{\gamma}}}x^2}{\bar{\gamma}^3} \right] \left[\frac{2e^{-\frac{2y}{\bar{\gamma}}}y}{\bar{\gamma}^2} + \frac{2e^{-\frac{2y}{\bar{\gamma}}}y^2}{\bar{\gamma}^3} \right] \\ & \left[1 - e^{-\frac{2y}{\bar{\gamma}}} - \frac{2e^{-\frac{2y}{\bar{\gamma}}}y}{\bar{\gamma}} - \frac{2e^{-\frac{2y}{\bar{\gamma}}}y^2}{\bar{\gamma}^2} \right]^{2-2} \end{split}$$

leading to

$$f_{X,Y}^{2}(x,y) = 4 \left[\frac{e^{-\frac{2x}{\bar{\gamma}}}x}{\bar{\gamma}^{2}} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^{2}}{\bar{\gamma}^{3}} \right] \left[\frac{e^{-\frac{2y}{\bar{\gamma}}}y}{\bar{\gamma}^{2}} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^{2}}{\bar{\gamma}^{3}} \right] \quad (6.4.11)$$

or

$$\begin{split} f_{X,Y}^3(x,y) &= 2(2-1) \left[\frac{e^{-\frac{2x}{\bar{\gamma}}}x^2}{\bar{\gamma}^3} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^3}{\bar{\gamma}^4} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^4}{\bar{\gamma}^5} \right] \left[\frac{e^{-\frac{2y}{\bar{\gamma}}}y^2}{\bar{\gamma}^3} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^3}{\bar{\gamma}^4} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^4}{\bar{\gamma}^5} \right] \\ \left[1 - e^{-\frac{2y}{\bar{\gamma}}} - \frac{2e^{-\frac{2y}{\bar{\gamma}}}y}{\bar{\gamma}} - \frac{2e^{-\frac{2y}{\bar{\gamma}}}y^2}{\bar{\gamma}^2} - \frac{e^{-\frac{2y}{\bar{\gamma}}}y^3}{\bar{\gamma}^3} - \frac{e^{-\frac{2y}{\bar{\gamma}}}y^4}{\bar{\gamma}^4} \right]^{2-2} \end{split}$$

yielding

$$f_{X,Y}^{3}(x,y) = 2\left[\frac{e^{-\frac{2x}{\bar{\gamma}}}x^{2}}{\bar{\gamma}^{3}} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^{3}}{\bar{\gamma}^{4}} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^{4}}{\bar{\gamma}^{5}}\right]\left[\frac{e^{-\frac{2y}{\bar{\gamma}}}y^{2}}{\bar{\gamma}^{3}} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^{3}}{\bar{\gamma}^{4}} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^{4}}{\bar{\gamma}^{5}}\right]$$
(6.4.12)

where again the superscripts in the left-hand side correspond to N. Then, integrate (6.4.10) to find the CDF, yielding

$$F_{\gamma}^{N}(\gamma) = \int_{0}^{\gamma/2} \int_{0}^{\gamma-y} f_{X,Y}^{N}(x,y) dx dy$$
 (6.4.13)

The CDF for two relays when N=2 is calculated by substituting (6.4.11) into (6.4.13), yielding

$$F_{\gamma}^{2}(\gamma) = \int_{0}^{\gamma/2} \int_{0}^{\gamma-y} 4\left[\frac{e^{-\frac{2x}{\bar{\gamma}}}x}{\bar{\gamma}^{2}} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^{2}}{\bar{\gamma}^{3}}\right] \left[\frac{e^{-\frac{2y}{\bar{\gamma}}}y}{\bar{\gamma}^{2}} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^{2}}{\bar{\gamma}^{3}}\right] dxdy$$
(6.4.14)

Similarly when N=3

$$F_{\gamma}^{3}(\gamma) = \int_{0}^{\gamma/2} \int_{0}^{\gamma-y} 2\left[\frac{e^{-\frac{2x}{\bar{\gamma}}}x^{2}}{\bar{\gamma}^{3}} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^{3}}{\bar{\gamma}^{4}} + \frac{e^{-\frac{2x}{\bar{\gamma}}}x^{4}}{\bar{\gamma}^{5}}\right] \left[\frac{e^{-\frac{2y}{\bar{\gamma}}}y^{2}}{\bar{\gamma}^{3}} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^{3}}{\bar{\gamma}^{4}} + \frac{e^{-\frac{2y}{\bar{\gamma}}}y^{4}}{\bar{\gamma}^{5}}\right] dxdy$$
(6.4.15)

Therefore the outage probability of the two-relay scheme can be evaluated using the CDF's form in (6.4.14)when N=2 and (6.4.15) when N=3 as

$$P_{out} = \int_0^{\gamma_{th}} f_{\gamma}^N(\gamma) d\gamma = F_{\gamma}^N(\gamma_{th}) \tag{6.4.16}$$

The probability of the cognitive relays acquiring the available spectrum is discussed in the next section.

6.4.2 The Outage Probability Analysis when Spectrum Acquisi-

tion is Imperfect

In cognitive relay networks the relay nodes need to acquire the spectrum before they transmit; however, in a realistic condition, the spectrum may not always be available for transmission. Hence, the probability of sensing the available spectrum must be considered. If i is the number of cognitive relays which successfully acquire spectrum opportunistically among M cognitive relay nodes, the probability of acquiring available spectrum is given by [123]

$$P_r(i) = \binom{M}{i} P_d^i (1 - P_d)^{M-i}$$
(6.4.17)
where $P^{i}{}_{d}$ denotes the probability of detection for every relay *i*. Therefore, the outage probability of the cognitive relay network when considering the probability of detection is given as

$$P^{Cog}_{out} = \sum_{i=0}^{M} P_d^i P_r(i)$$
 (6.4.18)

Next, the simulation results of the outage probability analysis, and analysis of the end-to-end frame error rate (FER) as a function of E_b/N_0 are presented.

6.5 Simulation results

6.5.1 Analysis of the FER vs E_b/N_0

In this section, the error performance of the proposed scheme was evaluated in quasi-static frequency selective channels. The fading was assumed constant within a frame and changes independently from one frame to another. The cognitive wireless communication relay network was considered to be a half-duplex cognitive relay network consisting of a source node, a destination node, and a minimum of four intermediate relay nodes acting as CR_s . It was assumed that each cognitive relay node is within the transmission range of one PU node. In order to combat this frequency selective fading, orthogonal frequency division multiplexing (OFDM) with cyclic prefix (CP) was implemented at the source node. Also OFDM-based decode-and forward (DF) relaying scheme was adopted at the cognitive relays. It was assumed that there is no direct link between the source and destination nodes, and there is perfect channel information at the relays moreover, the receiver has perfect channel information. The end-to-end FER is sim-



Figure 6.3. The FER performance of the EO-SFBC-OFDM for cognitive wireless relay network

ulated against E_b/N_0 . In Figure 6.3, a simulation result is provided showing the performance of the proposed scheme in which the spectrum map of the local channel between the source node and cognitive relays was changed randomly after every frame. The result shows that a seamless transmission was realized due to the existence of at least one available band in the cognitive relays that was utilized as a relay channel to ensure continuous data transmission. The results in Figure 6.4 demonstrate the comparison between end-to-end FER performance at three different situations, that is: when two, three, and four cognitive relays are in operation. The results show a significant improvement in performance when the number of available bands (cognitive relays) increases maintaining at the same time non-stoppable transmission. At a frame error probability of 10^{-3} , the four-relay cognitive system pro-



Figure 6.4. The FER performance of the EO-SFBC-OFDM for cognitive wireless relay network

vides about three dB improvement in the E_b/N_o than the three-relay system. The four-relay system performance benefits from the utilization of closed loop EO-SFBC which was proved to enhance the diversity gain as well as the array gain leading to the overall improvement of the system performance. Furthermore, Figure 6.5 shows the spectrum map of the simulated system for a 100 frames indicating the availability of spectrum during which period and the number of active cognitive relays at each frame during the transmission from the cognitive relays to the destination node. It can be noticed from the figure that a seamless data service for cognitive users is supported while causing no interference to primary users.



Figure 6.5. Spectrum map of the closed loop EO-SFBC cognitive relay system for 100 frames

6.5.2 Simulation results of outage probability analysis

In order to verify the results obtained from (6.4.16) and (6.4.18), all the relay node links have the same average SNR ($\bar{\gamma}$), there is no direct link between the transmitter and the receiver due to shadowing, or distance, and all nodes are equipped with a single antenna are assumed. Simulations are obtained when N = 2 and N = 3 for a fixed number of one and two relays, and SNR=20 dB. Figure 6.6 shows that the outage performance of the system improves as the number of cognitive relays increases. It can be seen that when the number of available relays M increases the outage probability decreases and hence when the number of the cognitive relays is large the outage event (no transmission) becomes less likely. For instance, at SNR value of 8 dB and N=2 paths, the outage performance improves from approximately 1.507×10^{-2} to 1.262×10^{-3} when the number of cognitive relays increase



from one to two relays respectively. Similarly, at SNR value of 8 dB

Figure 6.6. Comparison of the outage probability of one relay and two relay schemes of two-hop wireless transmission with two paths (N=2) and three paths (N=3), with $\bar{\gamma} = 20 dB$. The theoretical results are shown in line style and the simulation results as points.

and when N=3 paths, the outage performance improves from approximately 2.032×10^{-4} to 5.644×10^{-5} when the number of cognitive relays increases from one to two relays respectively. This result confirms that the two relay pair provides more robust transmission than a single relay. Furthermore, there is approximately a 6dB improvement in outage performance at 10^{-4} as the number of paths are increased from two to three paths when one as well as two cognitive relays are in operation. Figures 6.7 and 6.8 presents the comparison of outage probability performance for the proposed cognitive relay network at different values of spectrum acquisition probability when there are two active cognitive relays and



Figure 6.7. The theoretical and the simulation results of the outage probability performance when spectrum acquisition (P_d) is imperfect for two cognitive relays with two paths (N=2).

when N=2 and N=3 respectively. These results confirm that increasing the channel length potentially provides more robust transmission. For instance, in Figure 6.7, when the number of paths N=2 at SNR value of $\gamma = 4dB$ when the spectrum acquisition probability changes from $P_d = 0.95$ to Pd = 0.99 the outage probability decreases significantly from approximately 9.243×10^{-3} to 1.681×10^{-3} and it reaches the minimum value of 2.216×10^{-4} when $P_d = 1$ (perfect spectrum acquisition). Similarly, in Figure 6.8, when the number of paths N=3 at SNR value of $\gamma = 8dB$ when the spectrum acquisition probability changes from $P_d = 0.95$ to Pd = 0.99 the outage probability decreases significantly from approximately 4.073×10^{-3} to 4.312×10^{-4} and it reaches the minimum value of 4.178×10^{-6} when $P_d = 1$ (perfect spectrum acquisition).



Figure 6.8. The theoretical and the simulation results of the outage probability performance when spectrum acquisition (P_d) is imperfect for three cognitive relays with three paths (N=3).

In general, the higher outage probability is the indication of performance degradation due to the unavailability of the spectrum; as P_d decreases, the gain also decreases resulting in a higher outage probability. The outage probability improves if the spectrum can be acquired with a higher probability. Hence, the performance of the cognitive relay network depends entirely on the acquisition of the spectrum.

6.6 Summary

In this chapter a closed loop EO-SFBC for wireless cognitive relays has been proposed simulated and the performance evaluation of this system has been demonstrated. It is shown that the coding scheme together with the closed loop EO-SFBC technique do not only guarantee a seamless and continuous transmission for the cognitive users without causing any interference to the primary users, but also improve the diversity gain as well as the array gain of the system. Moreover, the outage probability performance of a proposed cooperative cognitive relay network was evaluated. A closed form expression for the outage probability for cooperation over frequency selective fading channels was derived for perfect and imperfect spectrum acquisitions. The derived formulas are simple, applicable to arbitrary number of relays and SNR values. A straightforward approach based on a Gamma distribution with positive integer scale parameter which has the same distribution as the sum of N independent exponential distribution random variables was used. The results indicated that the theoretical calculation and simulations are identical at different channel lengths (N = 2 and 3). In addition, the robustness of the two operational cognitive relay scheme over the single relay scheme was confirmed. Moreover, increasing the channel length and/or the number of relays improve the outage probability. Furthermore, The results showed the advantage in probability performance of the cooperative cognitive systems when the spectrum acquisition is perfect. In contrast, when the spectrum acquisition is

In the next chapter, the summary and conclusion to the thesis and future work will be provided.

imperfect the performance of the outage degrades significantly.

CONCLUSIONS AND FUTURE WORK

In this chapter, the novel results of this thesis and the conclusions that can be drawn from them are summarized. A discussion of possible future work is then included.

7.1 Conclusions

In this thesis, the issue of asynchronism among relay nodes was taken into consideration in cooperative communication systems. With the existence of asynchronism, the STB code structure at the destination node is not orthogonal and hence the transmitted signals are prevented from being successfully decoded at the destination node due to the resulting interference with a conventional STBC decoder. Different interference cancellation methods were therefore proposed in this thesis to mitigate such interference at the destination node and deliver a good performance with full cooperative diversity.

In Chapter 1, a general introduction of conventional MIMO systems was provided. Moreover, a brief introduction to the design of spacetime codes and their enhancement for cooperative relay networks was introduced. Then, the impairments of cooperative relay channels were briefly discussed. Finally, a quick introduction to cooperative cognitive networks was included.

In Chapter 2, a concise background for STCs was given and the milestone works in this field were described. This was followed by discussion of the concept of maximum likelihood (ML) decoding and pairwise error probability (PEP). In addition, the definition of uncoded and coded transmission was introduced thereafter. Then, the principle of parallel interference cancelation (PIC) was presented, followed by the principle of orthogonal frequency division multiplexing (OFDM) type transmission and its advantages and disadvantages. A review of synchronous cooperative relay networks and cooperative operation was next provided, then, the literature relevant to asynchronous cooperative relay networks was briefly discussed.

In Chapter 3, a parallel interference cancelation (PIC) detection scheme was used with distributed closed-loop extended orthogonal STBC (CL-EO-STBC) design for four asynchronous cooperative relay nodes and distributed closed-loop quasi orthogonal STBC (CL-QO-STBC) design for two dual-antenna asynchronous cooperative relay nodes to cancel the ISI caused by imperfect synchronization and achieve full diversity order with coding gain and full data transmission rate in each hop for both proposed schemes.

It was observed that the PIC detection is an effective approach to improve the performance of distributed STBC and mitigates the ISI and also improves its efficiency over large amounts of time misalignments. For example as shown in Table 3.1, the value of 10^{-2} BER cannot be achieved by ML detection in the proposed scheme even under small time misalignment $\beta_k = -6$ dB, whereas with the PIC detector involved, it just required approximately 14 dB of gain to achieve the value of 10^{-2} BER when $\beta_k = -6$ dB and q = 3. Furthermore, as it was observed in simulations, even under large time misalignment $\beta_k = 3$ dB and q=3, it required approximately 18 dB of gain to achieve 10^{-2} as compared to DT link performance, which required approximately 22 dB to achieve 10^{-2} BER. Similarly, simulation results for the CL-QO-STBC scheme under imperfect synchronization were depicted and as shown in Table 3.2, for example, the value of 10^{-2} BER cannot be achieved by ML detection in the proposed scheme even under small time misalignment $\tau_k = 0.125T$ dB, whereas with the PIC detector involved, it just required approximately 14 dB of gain to achieve the value of 10^{-2} BER when $\tau_k = 0.5T$ dB and q = 3.

Furthermore, the PEP utilizing distributed CL-EO-STBC and CL-QO-STBC for an asynchronous cooperative relay network was analyzed to confirm that the cooperative diversity gain is equal to the number of transmitting antennas on the relay nodes in the case of distributed CL-QO-STBC and that a smaller PEP is attained due to array gain in the case of distributed CL-EO-STBC. This result was based on the destination node having full knowledge of all fading channels.

These approaches ensured diversity gain of the order of the number of transmitting antennas within the cooperative relay nodes in addition to achieving array gain with full data transmission rate between the relay nodes and the destination node. Moreover, the results of the PIC approach with both proposed schemes showed that there was significant performance improvement over ML detection under synchronization error and performance close to the perfect synchronized case was achieved in just three iterations; whereas ML detection fails to mitigate the impact of imperfect synchronization even under small time misalignment. Although both distributed STBC schemes showed similar results, in the case of CL-QO-STBC a significant complexity reduction was achieved at the relay nodes in terms of time misalignment calculations due to the dual-antenna system model incorporated.

In Chapter 4, a near-optimum detection approach was proposed and analyzed employing half-duplex DF and AF types of relaying transmission based on distributed closed-loop EO-STBC with phase rotation [30] [111] and outer convolutive coding for wireless relay networks over frequency flat fading under imperfect synchronization. A system with two dual-antenna relay nodes was demonstrated in particular.

It was shown that in both cases (DF and AF), this near-optimum approach is effective at removing ISI at the destination node caused by time misalignment among relay nodes with computational detection complexity at the destination node dependant only on the constellation size and without the need for the direct transmission link. Furthermore, fourth order diversity with coding gain and unity data transmission rate between the relay nodes and the destination node was exploited by the proposed approach that employed two dual-antenna relay nodes with the equivalent decoding complexity as in [48]. In the DF scheme, the results showed that the smaller the time misalignment, the better the performance in terms of BER. For example, at a BER of 10^{-3} , the proposed near-optimum scheme with RS and minimum time misalignment (in the range of [-6 ... 6] dB) required approximately 4.7 dB of SNR and in the range of [-3 ... 6] dB required approximately 5 dB of SNR, while the conventional perfectly synchronized distributed CL-

EO-STBC scheme required approximately 4 dB of SNR for achieving the same value of BER.

On the other hand, when the AF protocol was used, the complexity at the relay nodes was reduced and the feedback resolution was restricted to one-bit which is practically achievable in the bi-directional control channels present in many communication systems; therefor reducing the high cost and high complexity due to the increased feedback information utilized in the case of the DF transmission protocol. It was noticed that the one-bit feedback scheme outperforms the open-loop and the two-bit feedback schemes. For example it required approximately 13.5 dB of total power to achieve a BER of 10^{-3} for the one-bit feedback scheme without RS whereas, approximately 11 dB was required for the same value of BER when RS was applied. Moreover, when the time misalignment, for instance, $\beta = 0$ dB and 3 dB, the one-bit feedback EO-STBC detector could not deliver good performance comparing with the PS case. However, when the proposed RS scheme was used to randomly select the minimum time misalignment from a uniformly distributed values of $\beta = [0, ..., 10]$ dB among the set of candidate relay nodes without prior knowledge of the values of timing errors, the performance of the one-bit feedback scheme was effectively enhanced achieving a BER of 10^{-3} with approximately 14.5 dB of total transmission power.

In Chapter 5, a narrowband system, where the two-hop channels are assumed to be flat Rayleigh fading channels, was initially considered with the M-QO-STBC transmission technique. Next, a new low complexity one-bit feedback scheme based on the selection of the cyclic phase rotation scheme was proposed for broadband systems, where the two-hop channels are assumed to be frequency-selective Rayleigh fading channels, and the D-EO-STBC transmission technique was employed. This approach attained unity rate over each hop in the network and full cooperative spatial diversity. Moreover, in both approaches, OFDM was implemented at the source node and a cyclic prefix (CP) was added at the source and relay nodes to combat the effects of random delays at the relay nodes. The relays operated in a simple amplify-and-forward (AF) mode. The obtained results showed that the OFDM-based scheme can significantly mitigate the delay when $\tau < CP$ which enhanced the performance of the system under imperfect synchronization. In addition the simulations showed that the OFDM-based M-D-QO-STBC scheme provides a better performance over the OFDM-based conventional D-QO-STBC [28] due to achieving full spatial diversity and the improvement in the code gain distance (CGD) which was obtained from the maximization of the distance properties of the original codebooks of the modified code. On the other hand, results demonstrated that the proposed D-EO-STBC-OFDM scheme with one-bit feedback is better than the closed-loop D-EO-STBC-OFDM in [30]. In particular it was clear that the un-coded proposed scheme, when FER was 10^{-3} , provided a power saving of approximately 8 dB as compared to the un-coded open-loop scheme and approximately 4 dB as compared to the previous feedback scheme [30] in wireless relay networks respectively. It was shown that the proposed closed-loop D-EO-SFBC-OFDM with onebit/subcarrier feedback can get the conventional diversity gain equal to the number of transmit antennas co-located in each relay and more additional array gain (feedback gain) at the same time. Furthermore, the comparison of the purposed scheme when combined with outer convolutive coding to exploit coding gain with the previous feedback scheme in a distributed manner was performed and it showed that at $FER = 10^{-3}$ the one-bit/subcarrier feedback scheme can achieve a power saving of approximately 2.5 dB as compered to the previous coded closed-loop scheme [30].

Finally, in Chapter 6, a closed loop EO-STFBC for wireless cognitive relays was proposed and the performance evaluation of this system was demonstrated. It was shown that the coding scheme together with the closed loop EO-STFBC technique do not only guarantee a seamless and continuous transmission for the cognitive users without causing any interference to the primary users, but also improve the diversity gain as well as the array gain of the system. Moreover, the outage probability performance of a proposed cooperative cognitive relay network was evaluated. The results demonstrated the comparison between end-toend FER performance at three different situations, that is: when two, three, and four cognitive relays are in operation. The results showed a significant improvement in performance when the number of available bands (cognitive relays) increased maintaining at the same time nonstoppable transmission. For example, at a frame error probability of 10^{-3} , the four-relay cognitive system provides about 3 dB improvement in the E_b/N_o than the three-relay system. The four-relay system performance benefits from the utilization of closed loop EO-SFBC which was proved to enhance the diversity gain as well as the array gain leading to the overall improvement of the system performance.

A closed form expression for the outage probability for cooperation over frequency selective fading channels was derived for perfect and imperfect spectrum acquisitions. The derived formulas are simple, applicable to arbitrary number of relays and SNR values. A straightforward approach based on a Gamma distribution with positive integer scale parameter which has the same distribution as the sum of N independent exponential distribution random variables was used. The results indicated that the theoretical calculation and simulations are identical at different channel lengths (N = 2 and 3). In addition, the robustness of the two operational cognitive relay scheme over the single relay scheme was confirmed. Moreover, increasing the channel length and/or the number of relays improved the outage probability. Furthermore, The results showed the advantage in probability performance of the cooperative cognitive systems when the spectrum acquisition is perfect. In contract, when the spectrum acquisition is imperfect the performance of the outage degrades significantly. The results showed that the outage performance of the system improved as the number of cognitive relays increased. It was seen that when the number of available relays M increased the outage probability decreased and when the number of the cognitive relays was large the outage event (no transmission) became less likely. For instance, at SNR value of 8 dB and N=2 paths,

the outage performance improved from approximately 1.507×10^{-2} to 1.262×10^{-3} when the number of cognitive relays increased from one to two relays respectively.

7.2 Future Work

This thesis opens up a number of research problems that can be considered. A few of the possible extensions of this work are as follows:

• Throughout the thesis, channel state information (CSI) at the

destination node is assumed to be known perfectly. However, in a real world application, CSI can only be estimated which obviously will introduce an error in CSI. An interesting study would be examining the proposed schemes in this thesis with channel estimate imperfections, and designing robust algorithms to mitigate such uncertainty.

- The near-optimum detection method in Chapter 4 can be applied to distributed CL EO-STBC in Chapter 3.
- Studying of successive interference cancelation (SIC) detection in [124] and [125] with the proposed closed-loop STBC schemes and comparisons with PIC detection schemes in frequency flat fading channels is an important open problem.
- Consideration of overloading in asynchronous multi-user cooperative multiple-input multiple-output (MIMO) [126] environments, possibly exploiting more sophisticated optimization approaches, such as genetic algorithms (GAs) in [127] and [128] is an interesting research area.
- In Chapter 6, an interesting extension of this work is to analyse cognitive relaying in an underlay environment dealing with power distribution to avoid or minimize the interference created to the primary user. Furthermore, the proposed work in this chapter can be extended to be considered in two-way cooperative cognitive relay networks.
- Finally, wireless communication security is becoming a topical area of research and engineering [129] and [130]. In particular,

physical layer security becomes an important research area that explores the possibility of achieving prefect secrecy data transmission among intended network nodes, whereas possible malicious node that eavesdrop the communication can not obtain useful information [131], therefore, how to use multi-relay selection scheme to improve secrecy outage probability which is a challenge problem should be studied in the future.

CLOSED-LOOP BASED METHODS FOR EXTENDED SPACE TIME BLOCK CODE FOR ENHANCEMENT OF DIVERSITY

The closed-loop based methods of STBC for the enhancement of a diversity system with four transmit and one receive antennas are proposed in [30] and [31]. In [31], for a four transmit antenna scheme, it was proposed in order to achieve full diversity gain, that the transmitted signal from the first and third transmit antenna are per-multiplied by $U_1 = (-1)^a$ and $U_2 = (-1)^b$, where a, b = 0, 1, respectively. Therefore, the Grammian matrix can be calculated as in (2.2.31), where

$$\lambda_c = |U_1|^2 |h_1|^2 + |h_2|^2 + |U_2|^2 |h_3|^2 + |h_4|^2$$
(A.1)

and

$$\lambda_f = 2\Re\{U_1h_1h_2^* + U_2h_3h_4^*\}$$
(A.2)

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Here, λ_c is the conventional channel gain for four transmit antennas, the feedback performance gain λ_f in (A.2) can also be represented as follows

$$\lambda_f = 2\Re(h_1h_2^*)(-1)^a + 2\Re(h_3h_4^*)(-1)^b \tag{A.3}$$

This changes with the defined values $U_1 = (-1)^a$ and $U_2 = (-1)^b$, which are determined by the feedback information bits a and b. According to the above analysis, the design criterion of the two-bit feedback scheme was proposed to ensure that each element of the feedback performance gain λ_f in (A.3) should be nonnegative as follows

$$(a,b) = \begin{cases} (0,0), & \text{if } \Re(h_1h_2^*) \ge 0 & \text{and } \Re(h_3h_4^*) \ge 0 \\ (0,1), & \text{if } \Re(h_1h_2^*) \ge 0 & \text{and } \Re(h_3h_4^*) < 0 \\ (1,0), & \text{if } \Re(h_1h_2^*) < 0 & \text{and } \Re(h_3h_4^*) \ge 0 \\ (1,1), & \text{if } \Re(h_1h_2^*) < 0 & \text{and } \Re(h_3h_4^*) < 0 \end{cases}$$
(A.4)

It can be noted that from (A.4), the feedback bits are chosen to maximize the value of λ_f in (A.3). This can lead to a larger received signal-tonoise ratio (SNR) at the receiver and achieve full diversity order.

However, in [30] the signal transmitted from the first and the third antennas are instead rotated by phase angles (phase shifted) $U_1 = e^{j\theta_1}$ and $U_2 = e^{j\theta_2}$ respectively while the other two antennas are kept unchanged. The phase rotation on transmitted symbols is importantly effectively equivalent to rotating the phase of the corresponding channel coefficients. Therefore, the conventional channel gain for four transmit antennas can be represented as in (A.1) and the feedback performance gain λ_f in (A.3) can be represented as follows

$$\lambda_f = 2\Re(h_1 h_2^*) e^{j\theta_1} + 2\Re(h_3 h_4^*) e^{j\theta_2} \tag{A.5}$$

where θ_1 and θ_2 can be obtained as follows

$$\theta_1 = -\angle (h_1 h_2^*)$$

$$\theta_2 = -\angle (h_3 h_3^*)$$
(A.6)

This design criterion of two-bit feedback was designed to ensure that the feedback performance gain should be real and positive, which leads to an improved SNR at the receiver and achieve full diversity order.

END-TO-END BIT ERROR RATE (BER)

If the l^{th} stage experiences independent probability of error BER, which is denoted here as $P_{b,l\in(1,K)(e)}$, where K is the number of relaying stages, the probability of error free transmission at stage l can be expressed as

$$1 - P_{b,l \in (1,K)(e)}$$
 (B.1)

hence the average probability of correct end-to-end transmission $P_{c,e2e}(e)$ can be expressed as the joint probability of correct transmission at each stage, i.e.

$$P_{c,e2e}(e) = \prod_{l=1}^{K} (1 - P_{b,l(e)})$$
(B.2)

A bit transmitted from a source terminal is received correctly at the destination only when at all the stages the bits have been transmitted correctly. Thus, the end-to-end BER $P_{b,e2e}(e)$ can therefore be expressed as

$$P_{b,e2e}(e) = 1 - P_{c,e2e} \tag{B.3}$$

which at low BER at each stage can be approximated as

$$P_{b,e2e}(e) \approx \sum_{l=1}^{K} P_{b,l(e)}$$
$$\approx \sum_{l=1}^{K} \frac{P_{s,l(e)}}{\log_2(M_l)}$$
(B.4)

where M_l is the modulating index employed at the l^{th} stage. From (B.4), it is clear that the end-to-end BER will be dominated by the worst stage [47].

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