



Faculty of Engineering

**Multiple Symbol Double Differential Transmission for
Cooperative Wireless Communication Networks with Different
Mobility**

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DECLARATION

The thesis has not been accepted for any degree and is not concurrently submitted in candidature for any other degree.

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ABSTRACT

The cooperative communications schemes are emerging towards the next generations of various wireless communication applications. With its distributed nature, it is challenging for the coherent detection to acquire fading channels knowledge than in the conventional single point communications. In order to avoid the complexity of channel and frequency offset estimation, Double Differential (DD) modulation transmission with non-coherent detection are introduced in this thesis. Specifically, this thesis studies the behaviour of non-coherent detection with different mobility scenarios (i.e. time-varying channels). The channel variation can be related to the normalized Doppler shift which indicates the user's mobility. The normalized Doppler shift is utilized to differentiate the slow time-varying (slow fading) and fast time-varying (fast fading) channels. In order to characterize the time-varying channel, a time series model is developed in the Amplify-and-Forward (AF) cooperative network. Firstly, the performance of two-symbol DD detection under Rayleigh Fading channels with time-varying is examined for cooperative network. It is observed that the error performance degrades especially in fast fading channels. In order to mitigate the degradation problem, a multiple symbol detection is developed. The proposed scheme adopting a direct combining scheme improve the performance of the system. In the second part of the thesis, a new combining weight for Maximal Ratio Combining (MRC) based on statistical channel knowledge is proposed. In order to further improve the system performance without requiring the channel and frequency offset estimation in MRC, a simpler combiner, Selection Combiner (SC) is developed and analysed in time-varying channels. The performance results shown that the SC outperform the proposed MRC. The final part the thesis studies a multi-branch relayed network with a direct link in double differential distributed space-time coding

environment. It is observed that by using the two-symbol detection, the system fails to perform as the diversity is affected by the channel variation. Thus, a multiple symbol double differential detection is developed in the distributed space-time coding environment to improve the network performance.

Keywords: Frequency offsets, double differential, distributed space-time coding, multiple symbol double differential detection

***Penghantaran Multiple Symbol Double Differential untuk Rangkaian Komunikasi
Kerjasama Tanpa Wayar dengan Perbezaan Mobiliti***

ABSTRAK

Skim Komunikasi kerjasama berkembang ke arah generasi akan datang dalam pelbagai aplikasi untuk komunikasi tanpa wayar. Berdasarkan sifat penyebarannya, amat mencabar bagi pengesanan koheren untuk memperoleh informasi saluran pudar daripada konvensional komunikasi satu poin. Bagi mengelakkan kerumitan anggaran saluran dan frekuensi offset, penghantaran permodulatan Double Differential (DD) bersama dengan pengesanan bukan koheren diperkenalkan dalam tesis ini. Tesis ini, khususnya, mengkaji sifat untuk pengesanan bukan koheren dengan senario mobiliti yang berbeza (iaitu saluran variasi masa). Variasi saluran boleh dikaitkan dengan Doppler anjakan yang merujuk kepada mobiliti pengguna. Doppler anjakan digunakan untuk membezakan di antara masa saluran variasi perlahan (pudar perlahan) dan masa variasi pantas (pudar pantas). Bagi mencirikan saluran variasi masa, satu model siri masa telah dibangunkan dalam rangkaian kerjasama Amplify-and-Forward (AF). Pertama sekali, prestasi untuk dua-simbol pengesanan DD di bawah saluran pudar Rayleigh dengan variasi masa dianalisa untuk rangkaian kerjasama. Diperhatikan bahawa prestasi ralat merosot terutamanya dalam saluran pudar pantas. Bagi mengatasi masalah kemerosotan tersebut, satu pengesanan pelbagai simbol telah dibangunkan. Skim yang dicadangkan menggunakan skim penggabungan langsung meningkatkan prestasi sistem. Dalam bahagian kedua tesis, satu pemberat penggabungan yang baru untuk Maximal Ratio Combining (MRC) berdasarkan informasi saluran statistik telah dicadangkan. Untuk meningkatkan lagi prestasi sistem tanpa memerlukan anggaran saluran dan frekuensi offset dalam MRC, satu gabungan yang lebih mudah, Selection Combiner (SC) dibangunkan dan dianalisa dalam saluran variasi masa. Keputusan prestasi

menunjukkan bahawa SC menghasilkan keputusan yang lebih baik berbanding dengan MRC yang telah dicadangkan. Bahagian terakhir tesis ini mengkaji satu rangkaian kerjasama multi saluran dan satu saluran langsung dalam persekitaran pembezaan berganda pengekodan ruang-masa. Diperhatikan bahawa dengan menggunakan pengesanan dua-simbol, sistem gagal untuk mencapai kepelbagaian disebabkan oleh variasi saluran. Maka, satu pengesanan pembezaan berganda pelbagai simbol telah dibangunkan dalam persekitaran pembezaan berganda pengekodan ruang-masa untuk memperbaiki prestasi rangkaian.

Kata kunci: *Frekuensi anjakan, pembezaan berganda, ruang-masa kod, pengesanan pelbagai simbol pembezaan berganda*

TABLE OF CONTENTS

	Page
DECLARATION	i
ACKNOWLEDGEMENT	ii
ABSTRACT	iii
<i>ABSTRAK</i>	v
TABLE OF CONTENTS	vii
LIST OF TABLES	xi
LIST OF FIGURES	xii
LIST OF ABBREVIATIONS	xvii
LIST OF MATHEMATICAL NOTATIONS	xix
CHAPTER 1 : INTRODUCTION	1
1.1 Problem Statement	8
1.2 Research Objectives	10
1.3 Structure of Thesis	11
1.4 Contribution of Research	12
CHAPTER 2 : LITERATURE REVIEW	15
2.1 Introduction	15
2.2 Single Link Wireless Communication	18
2.3 Wireless Channel Model	20
2.3.1 Rayleigh Statistical Model	21
2.3.2 Autoregressive Models	22
2.4 Cooperative Relaying Protocols	23
2.4.1 Decode-and-Forward (DF)	23

2.4.2	Amplify-and-Forward (AF)	24
2.5	Wireless Cooperative Networks	25
2.5.1	Signalling Schemes in Cooperative Communication Networks	27
2.5.1.1	Repetition-Based Transmission	27
2.5.1.2	Space-Time Coding	27
2.5.1.3	Distributed Space-Time Block Code (DSTBC)	30
2.6	Diversity Combining Techniques	31
2.6.1	Direct Combining	32
2.6.2	Maximal Ratio Combining	33
2.6.3	Selection Combining	34
2.7	Cooperative Diversity State-of-the-Art	36
2.8	Detection Techniques for Cooperative Diversity under Repetition-Based Transmission	38
2.8.1	Coherent Detection	38
2.8.2	Non-Coherent Detection	39
2.8.3	Differential Transmission to Bypass Channel Estimation	40
2.8.4	Detection Schemes in Highly Mobile Environment	45
2.8.5	Multiple Symbol Differential Detection	53
2.9	Cooperative Differential Distributed Space-Time Coding Transmission	62
2.10	Conclusion of Analysis	68
2.11	Summary	69
CHAPTER 3 : MULTIPLE SYMBOL DOUBLE DIFFERENTIAL SPHERE DECODING		70

3.1	Introduction	70
3.2	System and Channel Model	71
3.2.1	Channel Simulation Model	74
3.2.2	Channel Model	75
3.3	Detection Schemes at the Receiver for Cooperative Diversity	82
3.3.1	Conventional Double Differential Detection	82
3.3.2	Performance Analysis of Double Differential Detection	85
3.3.3	Multiple Symbol Detection	88
3.3.4	Performance Analysis of Multiple Symbol Detection	94
3.4	Discussion of Simulation Results for Multiple Symbol Based Detection	95
3.4.1	Power Allocation	100
3.4.2	Multiple Symbol Double Differential Scheme Under Various Mobility Effect	105
3.5	Summary	112
CHAPTER 4 : DIVERSITY COMBINING SCHEMES		113
4.1	Introduction	113
4.2	System and Channel Model	114
4.3	Maximal Ratio Combiner	115
4.4	Performance Analysis of Maximal Ratio Combiner	117
4.4.1	Error Probability Performance Analysis	117
4.4.2	Outage Probability	123
4.5	Discussion of Simulation Results for Maximal Ratio Combining Scheme	126
4.5.1	Error Probability Performance	127

4.5.2	Outage Probability	129
4.6	Selection Combiner	133
4.7	Performance Analysis of Selection Combiner	134
4.8	Discussion of Simulation Results for Selection Combiner	142
4.8.1	Performance of Error Probability	143
4.9	Summary	151
CHAPTER 5 : MULTIPLE CODEWORD DOUBLE DIFFERENTIAL SPHERE DECODING OVER DISTRIBUTED SPACE- TIME BLOCK CODING		152
5.1	Introduction	152
5.2	System Model	153
5.2.1	Two-Codeword Double Differential Detection	157
5.2.2	Multiple-Codeword Double Differential Detection	159
5.3	Discussion of Simulation Results for Multiple-Codeword Double Differential Sphere Decoding under Distributed Space-Time Block Coding (DSTBC)	163
5.3.1	Performance of Two-Codeword and Multiple-Codeword in Dual-Hop Transmission Only and Dual-Hop Transmission with Direct Link	166
5.3.2	Performance of MCDDSD Under Time-Varying Channels	170
5.4	Conclusion	172
CHAPTER 6 : CONCLUSION AND RECOMMENDATIONS FOR FUTURE WORKS		174
6.1	Conclusion	174
6.2	Future Recommendations	179
REFERENCES		181

LIST OF TABLES

	Page
Table 2.1 Contribution of detection scheme with frequency offset estimation	48
Table 2.2 Contribution of detection scheme that eliminate frequency offsets Estimation	48
Table 2.3 Major contributions of MSD scheme under fast fading channel	54
Table 2.4 Summary of receiver designs in the survey	61
Table 2.5 Major contributions of differential detection schemes for STC	66
Table 3.1 The channel coefficients and corresponding autocorrelation values over a Rayleigh flat fading channel under various fading rate	77
Table 3.2 Simulation parameters	83
Table 3.3 Power allocation for various channel variances	102
Table 3.4 Various channel qualities based on channel variances	104
Table 3.5 Simulation case based on normalized Doppler frequency	106
Table 4.1 Power allocation for different channel variances	147
Table 5.1 Different fading channels	170

LIST OF FIGURES

	Page
Figure 1.1 Cooperative communication in mobile networks	3
Figure 1.2 Cooperative communication for a single source, relay and destination	4
Figure 2.1 Flowchart of literature review	16
Figure 2.2 A simple block diagram of single link wireless communication	18
Figure 2.3 a) BPSK; b) QPSK, I denote the in-phase channel and Q is the quadrature channel	19
Figure 2.4 One-Way Relay Network (OWRN) and (b) Two-Way Relay Network (TWRN) [36]	26
Figure 2.5 Illustration of multi-relay cooperative communication system	26
Figure 2.6 A block diagram of the Alamouti two-branch transmit diversity scheme	28
Figure 2.7 Block diagram of STBC decoder for the point-to-point communication network	30
Figure 2.8 DSTBC transmission (a) phase I; (b) phase II	31
Figure 2.9 General block diagram of direct combining scheme for cooperative communication network	33
Figure 2.10 General block diagram of MRC scheme for cooperative communication network	34
Figure 2.11 General block diagram of SC for cooperative communication network	35
Figure 2.12 A block diagram of a single differential encoder and decoder [50]	41
Figure 2.13 Geometrical representation of the sphere decoding algorithm	45
Figure 2.14 Block diagram of DD encoder and decoder for single link wireless communication network [66]	50
Figure 2.15 Block diagram of the DD transmission [66]	51

Figure 2.16	Proposed DD transmission system model [24]	52
Figure 2.17	(a) Symbol-by-symbol based single differential detection (b) Multiple Symbol Double Differential Detection	59
Figure 3.1	Proposed system model	72
Figure 3.2	Direct link channel coefficients and its corresponding autocorrelation values of different cases	77
Figure 3.3	Relayed channel and its corresponding autocorrelations of different cases	78
Figure 3.4	Block diagram of DD transmission for single relay cooperative communication system	83
Figure 3.5	Block diagram of proposed MSD detection	88
Figure 3.6	Comparison of BER versus SNR performance of MSDD scheme, conventional DD scheme and coherent scheme when $\sigma_{s,r}^2 = 1$; $\sigma_{s,d}^2 = 1$; $\sigma_{s,d}^2 = 1$ and $P_S = P_R = 0.5P$ with BPSK constellation	96
Figure 3.7	Performance of BER for the proposed MSDD scheme with different length received symbol sequence N when $\sigma_{s,r}^2 = 1$; $\sigma_{s,d}^2 = 10$; $\sigma_{s,d}^2 = 1$ and $P_S = P_R = 0.5P$	99
Figure 3.8	Power Allocation Factor (PAF) at $P/N_0 = 20$ dB for different channel variances scenarios	101
Figure 3.9	BER performance comparison of the proposed MSDD scheme with symmetric power allocation and optimized power allocation under various channel quality link	105
Figure 3.10	Comparison of BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1$; $\sigma_{s,d}^2 = 1$; $\sigma_{s,d}^2 = 1$ for DDBPSK	107
Figure 3.11	Comparison of BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1$; $\sigma_{s,d}^2 = 1$; $\sigma_{s,d}^2 = 10$ for DDBPSK	108
Figure 3.12	Comparison of BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 10$; $\sigma_{s,d}^2 = 1$; $\sigma_{s,d}^2 = 1$ for DDBPSK	108

Figure 3.13	Comparison of BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1; \sigma_{s,d}^2 = 1; \sigma_{s,d}^2 = 1$ for DDQPSK	109
Figure 3.14	Comparison of BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1; \sigma_{s,d}^2 = 1; \sigma_{s,d}^2 = 10$ for DDQPSK	110
Figure 3.15	Comparison of BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 10; \sigma_{s,d}^2 = 1; \sigma_{s,d}^2 = 1$ for DDQPSK	110
Figure 4.1	Comparison of the BER performance for conventional MRC and proposed MRC utilizing DDBPSK for $f_{D_{s,d}}T_s = f_{D_{s,r}}T_s = f_{D_{r,d}}T_s = 0.05$	127
Figure 4.2	Simulation results of proposed multiple symbol AF relaying with different numbers of relays $N = 2,3,4$ employing DDBPSK when $f_{D_{s,d}}T_s = f_{D_{s,r}}T_s = f_{D_{r,d}}T_s = 0.05$	128
Figure 4.3	Outage probabilities of the MSDDSD-AF network for BPSK with semi MRC employing different numbers of relay terminals	130
Figure 4.4	Outage probabilities of the MSDDSD-AF network for QPSK with semi MRC employing different numbers of relay terminals	131
Figure 4.5	Outage probability performance of proposed MRC under different faded channels for MSDDSD-AF system for BPSK constellation	132
Figure 4.6	Outage probability performance of proposed semi MRC under different faded channels for MSDDSD-AF system for QPSK constellation	132
Figure 4.7	Block diagram of the SC at the destination of a DDAF network	133
Figure 4.8	Comparison of BER performance for the conventional MRC, semi MRC, direct combiner and proposed SC scheme based on MSDD when $\sigma_{s,r}^2 = 1; \sigma_{s,d}^2 = 1; \sigma_{s,d}^2 = 1$ for DDBPSK	144
Figure 4.9	Comparison of BER performance for the conventional MRC, semi MRC, direct combiner and proposed SC scheme based on MSDD when $\sigma_{s,r}^2 = 1; \sigma_{s,d}^2 = 1; \sigma_{s,d}^2 = 1$ for DDQPSK	145
Figure 4.10	BER as a function of PAF for $\frac{P}{N_o} = 20 \text{ dB}$	147

Figure 4.11	Theoretical numerical analysis and simulation BER of the DD-AF rates system with semi MRC and SC using BPSK under different fading and symmetrical channels $\sigma_{s,r}^2 = 1; \sigma_{s,d}^2 = 1; \sigma_{r,d}^2 = 1$	148
Figure 4.12	Theoretical numerical analysis and simulation BER of the DD-AF rates system with semi MRC and SC using BPSK under different fading and symmetrical channels $\sigma_{s,r}^2 = 1; \sigma_{s,d}^2 = 10; \sigma_{r,d}^2 = 1$	149
Figure 5.1	System model for distributed DD STC using AF relaying protocol	153
Figure 5.2	Comparison of the BER performance for the conventional DD and proposed MCDDSD scheme when $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1$	167
Figure 5.3	Comparison of the BER performance for the conventional DD and proposed MCDDSD scheme when $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 10; \sigma_{r,d}^2 = 1$	169
Figure 5.4	Comparison of the BER performance for the conventional DD and proposed MCDDSD scheme when $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 10$	169
Figure 5.5	Performance comparison of two-codeword and MCDDSD under different fading channels and $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1$ for BPSK constellation	171
Figure 5.6	Performance comparison of two-codeword and MCDDSD under different fading channels and $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 10; \sigma_{r,d}^2 = 1$ for QPSK constellation	172

LIST OF ABBREVIATIONS

AF	Amplify-and-Forward
AR	Autoregressive
ASK	Amplitude-Shift Keying
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase-Shift Keying
CCI	Co-Channel Interference
CDMA	Code Division Multiple Access
CDF	Cumulative Density Function
CF	Compress-and-Forward
CRC	Cyclic Redundancy Check
CSI	Channel State Information
DAF	Differential Amplify-and-Forward
DD	Double Differential
DDAF	Double Differential Amplify-and-Forward
DDF	Differential Decode-and-Forward
DDPSK	Double Differential Phase-Shift Keying
DDSTBC	Double Differential Space-Time Block Coding
DE	Differential Evolution
DF	Decode-and-Forward
DSTBC	Distributed Space-Time Block Coding
DSTM	Differential Space-Time Modulation

EMRC	Emulated Maximal Ratio Combining
EXIT	Extrinsic Information Transfer
FSK	Frequency-Shift Keying
GLRT	Generalized Likelihood Ratio Test
i.i.d	independent, identically distribution
ISI	Inter-Symbol Interference
LC	Linear Combiner
LOS	Line-of-Sight
MCDDSD	Multiple Codeword Double Differential Sphere Decoding
MGF	Moment Generating Function
MIMO	Multiple-Input Multiple-Output
ML	Maximum Likelihood
MMSE	Minimum Mean Square Error
MRC	Maximal Ratio Combiner
MSD	Multiple Symbol Detection
MSDD	Multiple Symbol Differential Decoding
MSDDSD	Multiple Symbol Double Differential Sphere Decoding
MSDSD	Multiple Symbol Differential Sphere Decoding
MSE	Mean Square Error
NCFSK	Non-Coherent Frequency-Shift Keying
OFDM	Orthogonal Frequency Division Multiplexing
OOK	On-Off Keying
OWRN	One-Way Relay Network
PDF	Probability Density Function

PEP	Pairwise Error Probability
PL	Piecewise Linear
PSK	Phase-Shift Keying
PSAM	Pilot-Symbol-Assisted Modulation
RF	Radio Frequency
SC	Selection Combiner
SD	Sphere Decoding
SDMA	Space-Division Multiple Access
SER	Symbol Error Rate
SISO	Single-Input Single-Output
SNR	Signal-to-Noise Ratio
SoS	Sum-of-Sinusoids
ST	Space-Time
STBC	Space-Time Block Code
STC	Space-Time Coding
STTC	Space-Time Trellis Codes
TDD	Time Division Duplex
TWRN	Two-Way Relay Network
QPSK	Quadrature Pulse-Shift Keying
WSN	Wireless Sensor Network
ZF	Zero-Forcing

LIST OF MATHEMATICAL NOTATIONS

x	vector x
X	matrix X
X^T	transpose of matrix X
X^H	Hermitian transpose of matrix X
X^*	complex conjugate of matrix X
$Tr\{X\}$	trace of matrix X
$\det\{X\}$	determinant of matrix X
$ X $	Frobenius norm of matrix X
$diag\{x\}$	square diagonal matrix with x as diagonal
$diag(X)$	column vector built with the main diagonal of X
$x[k]$	k th element of the vector x
I_n	$n \times n$ identity matrix
$Re\{x\}$	real part of x
$I\{x\}$	imaginary part of x
$\hat{x}$	estimate of x
$ x $	absolute value of x
$E\{x\}$	expectation of random variable x
$\mathcal{N}(\mu, \sigma^2)$	normal distribution with mean μ and variance σ^2
$N\{X\}$	Cholesky decomposition of X
$J_0(\cdot)$	0 th order Bessel function of the first kind
P	error probability
$\otimes$	linear convolution

f_D	Doppler frequency or fading bandwidth
M	Number of signals in modulation format
$\text{sign}\{x\}$	Signum function of a real number x

CHAPTER 1

INTRODUCTION

Wireless communications have become part of our daily life that connects the users wirelessly around the world. The applications vary widely from the tiny and small sensors, mobile phones to vast uses such as ad hoc wireless networks, radar system, satellites communications as well as a future cellular communication system. Specifically, the next-generation wireless communication system indicates the next major paradigm shift beyond the present technologies to become one of the attractive and unique domains in ubiquitous wireless communication fields that provide untethered connectivity, higher spectral efficiency to support the high data rate services.

The data transmission over the wireless digital communication channels is greatly affected by the natural phenomenon, referred as fading. Fading channel causes variation of random fluctuations or amplitude and phase attenuation in the received signals resulted by the arriving of different multiple signal components via different propagation paths at the receiver. In fact, it is a challenging task to cater reliable as well as high-speed information transfer in such environment due to the effect of fading. Fading can be categorized into two, namely large-scale and small-scale fading. For large-scale fading, the detected signals at the receiver suffer from performance loss or degradation of signal strength. This is due to the long period of time and shadow effect caused by large-sized objects (i.e., buildings, mountains and hills). On the contrary, small-scale fading arises owing to the destructive and constructive interfering components of the multipath signal propagation as a consequence of scattering, reflection and diffraction from the obstructing objects over a short distance. These particularly severe forms of interference result in multiple replicas of received signals and

ultimately experience attenuation that leads to performance degradation. In certain practical scenarios such as wireless mobile communications and vehicular networks, small-scale fading constitutes larger impact on the received signal as the small changes in the transmitter and receiver mobility causes rapid frequency fluctuations due to Doppler effects. The fading channel is further described as slow and fast fading. Hence, with the intention to overcome the deleterious fading, fading mitigation techniques are employed to enhance the quality of the wireless communications.

In wireless communication, diversity is one of the efficient mechanisms that has the ability to mitigate the deleterious effects associated with multipath propagation using multiple independent communication channels [1]. Amongst numerous diversity-oriented technologies, Multiple-Input Multiple-Output (MIMO) system utilizes multiple antennas at the source as well as the destination whereby the information signals are carried over the similar frequency through a variety of communication paths. The system is often desirable to enhance the link reliability, spectral as well as energy efficiency. Yet owing to the size, cost and hardware constraints, it is often impractical to install a numerous number of antennas within the pocket-sized transmitter and receiver nodes, for instance, in Wireless Sensor Network (WSN) and wireless mobile ad hoc network [2, 3]. Moreover, the messages broadcasted via multiple antennas may likely to experience deep fade simultaneously which leads to diversity gain¹ loss [4].

Hence, cooperative diversity appears to be one of the promising approaches to encounter the challenging requirement for the next-generation of broadband systems such as

¹ The diversity gain denotes the overall reliability performance of wireless link due to the diversity approach wherein numerous copies of information signal is obtained through independent fading channels.

WiMAX and 3GPP LTE-Advanced concerning fading mitigation, coverage, capacity, reliability, diversity and to overcome the unpredictable characteristics of the channel [2, 5, 6]. Multipath fading is mitigated via the provision of multiple independent transmission paths. For instance, the information signal is propagated via relays and at the same time along the source-destination path. In this way, cooperative relaying exploit spatial diversity even when the relays choose to listen rather than to transmit. This, consequently enhancing the system performance concerning outage and probability of error. It is shown in Figure 1.1 that the relay technologies are greatly and effectively improved in terms of service coverage as well as system throughput [7].

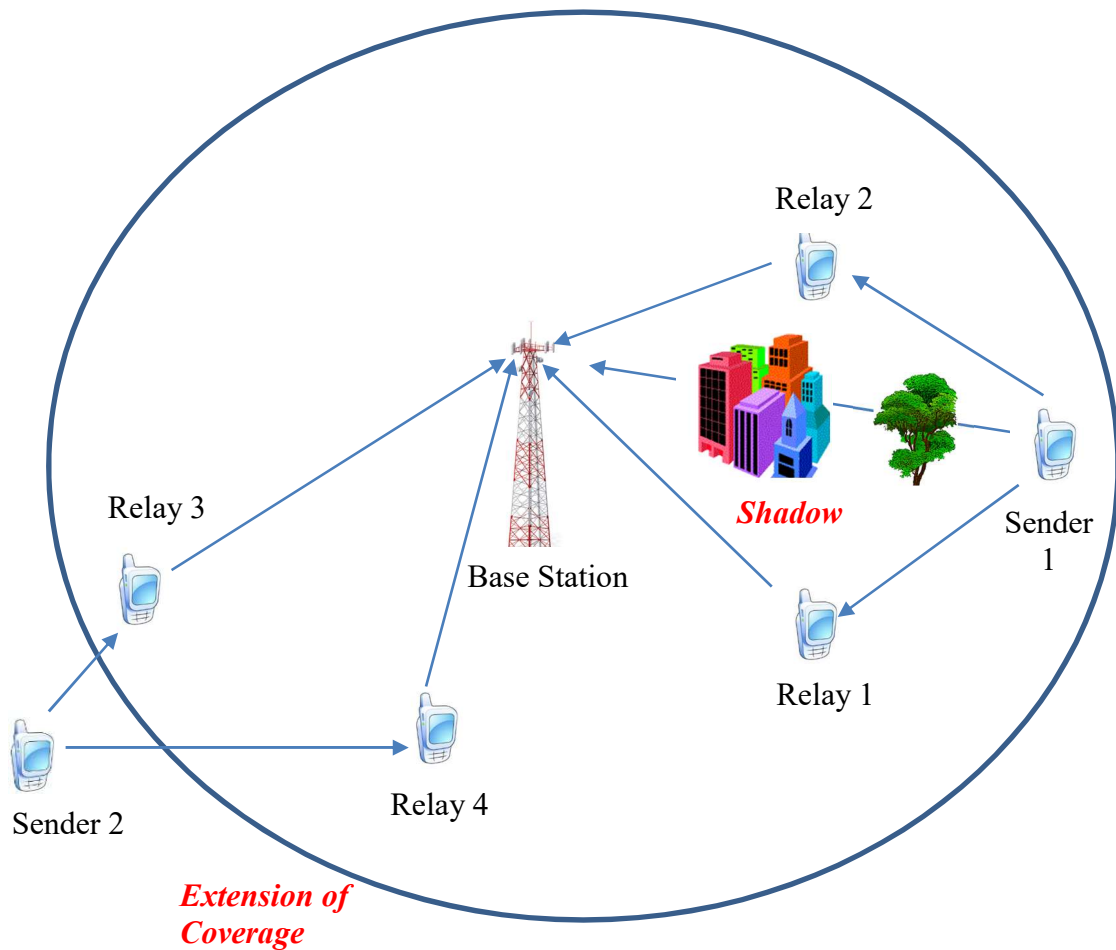


Figure 1.1: Cooperative communication in mobile networks

In general, wireless cooperative diversity exploits the broadcast nature and is capable to realize diversity via multiple independent propagation paths. By taking the benefit of exploiting the wireless medium transmission features, adjacent relay-aided nodes are able to ‘hear’ the broadcasted messages. The cooperation formed among multiple adjacent nodes can be realized by helping a source to send its messages to the desired destination, thus creating a virtual antenna array i.e. in wireless ad hoc mobile and vehicular networks. In these networks, the terminals communicate within a network based on the mobile-to-mobile communication via multi statistically independent paths. In the scenario when unreliable communication link occurs, a nearby node that acts as a relay can help to forward the messages to the desired destination. An illustration of a single relay cooperative network with three nodes is depicted in Figure 1.2.

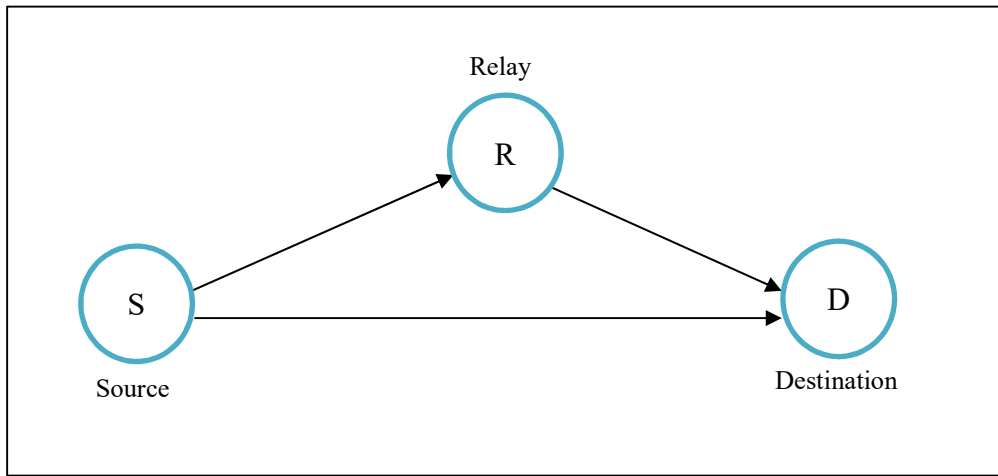


Figure 1.2: Cooperative communication for a single source, relay and destination [5]

As portrayed in Figure 1.2, the primary concept of a cooperative wireless communications network can be formed with a single source, relay and destination. Due to the interest of non-directional electromagnetic waves propagation, the source broadcasts its messages via two independent paths to the desired destination. The first path is alike to the

traditional point-to-point communication (i.e. the channel from source to destination). The second path is established since the relay can “listen” to the source via source-relay channel and at the same time processed the received signals based on the relaying protocols namely, Decode-and-Forward (DF), Amplify-and-Forward (AF) as well as Compress-and-Forward (CF). It is noted that AF relaying protocol is extensively emphasis in most studies due to its simplicity process at the relay.

Cooperative transmission can be divided based on repetition and Space-Time Block Coding (STBC) distribution. Based on repetition approach, the relay(s) transmit the processed signals towards the destination in Time Division Duplex (TDD) manner. In contrast, by using Distributed Space-Time Block Coding (DSTBC) approach, the relay(s) simultaneously transmit the processed signals whereby space-time code is devised at the destination. DSTBC approach can achieve superior bandwidth efficiency in comparison to repetition-based approach but it involves more sophisticated techniques.

At the destination, the detection schemes play a vital role to detect the incoming signals correctly. The detection schemes can be categorized as coherent and non-coherent. Coherent detection requires the Channel State Information (CSI) at the receiver and applied matched filters to detect and decide the transmitted data. On the contrary, non-coherent detection do not require CSI in order to recover the data. Many studies of cooperative diversity have considered coherent detection, assuming either partial or full CSI knowledge in the context of perfect fading coefficients estimation at both relay(s) and destination [8–10].

By using the conventional channel estimation scheme known as Pilot-Symbol-Assisted Modulation (PSAM) [1], the CSI can be obtained at the destination with the aid of the known pilot symbols. However, such an assumption implies system’s complexity, costly,

undesirable wastage of both the transmission power as well as the bandwidth and subjected to communication overhead especially in fast fading environment [11, 12]. Moreover, the inaccurate channel estimation investigated in [13, 14] may degrade the performance of the cooperative systems.

Alternatively, to alleviate the necessities of estimating the CSI knowledge at the destination, low-complexity differential decoding or non-coherent detection schemes is considered as one of the alternative solutions and has attracted research interest in recent years. Differential modulation incorporates the symbol-by-symbol based detection has been highlighted in [10, 15–18] based on Differential Amplify-and-Forward (DAF) scheme. The data symbols are encoded differentially utilizing Phase-Shift Keying (PSK) modulation at the source before being transmitted to the intended destination with the assumption that the channel is stationary for at least two continuous time period. This is because rather than analyzing the phase corresponds to the coherent signal, differential decoder determine the phase difference based on the previously and presently received signal and efficient decoding relies on constant response of the channel from one time sample to the next. Then, the decision variable is computed to obtain the optimized gain for Maximum Ratio Combiner (MRC) requiring the instantaneous channel knowledge of the relayed channel, which is unknown. Therefore, a group of fixed gains has been determined in [10, 15–17, 19] so as to obtain the statistical knowledge of the communication channel.

However, previous studies have primarily concentrated on omitting channel knowledge without considering the effect of frequency offsets [2, 4, 6, 8]. In wireless mobile communications, frequency offsets arise as a result of the relative motion between the source and receiver whereby the component is directly proportional to the Doppler effect. In addition, the velocity of the mobile terminal v and the carrier frequency f_c , determine the

fading rate or fading bandwidth f_D , that is, $f_D = \frac{v}{c} f_c$ where c is the speed of light. Hence faster motion leads to more rapid fading. For example, a mobile user travels in a fast moving train over 300 km/h causes Doppler effect up to 880 Hz in case that the carrier frequency is 2GHz [20]. Another cause of frequency offset is the frequency difference in the transmitter and receiver local oscillators.

With the frequency offsets effect, the fading channel varies over three consecutive time periods, which violates the basic assumption of differential transmission scheme – leads to performance degradation. So far, however, there has been little discussion about this practical problem in cooperative diversity system. There are two main methods to alleviate the frequency offsets effect. The first method involves the estimation of frequency offsets by introducing frequency acquisition and tracing circuits [21]. Another method is the Double Differential (DD) modulation which is insensitive to frequency offsets [22, 23]. Nevertheless, all the above schemes are executed for direct communication (i.e. point-to-point) system and is not suitable for One-Way Relay Network (OWRN) using AF with the presence of frequency offsets, and unavailable channel knowledge at all communication terminals. Therefore the aim of this research works is to develop a simple, low complexity, reliable as well as efficient DD based and diversity combining schemes for OWRN utilizing AF relaying protocol in a cooperative communication network with the presence of frequency offsets. The complexity problem is solved by finding a closed-form expression as the previous research solutions turn out to be polynomial.

Contrary to [24], the suggested scheme employing the DD modulation perform well for Single-Input Single-Output (SISO) system [25–27] wherein the communication channels are fixed over at least three time intervals. In order to further achieve better DD approach performance, a signal detection based on multiple-symbol is recommended. An extensive

literature review revealed that the other decoding techniques at the destination include Zero-Forcing (ZF), Minimum Mean Square Error (MMSE) and Maximum Likelihood (ML) decoder [28, 29]. By all of these techniques, ML decoding offers the best outcomes in view of other detectors. ML can be estimated by maximizing the function $\ln p(\bar{y}; h)$ where $\bar{y}$ represents the observation vector and h is the channels in this thesis. However, the ML decoder is frequently assumed to be realistically impractical because of its high computational complexity in the communication systems influenced by the number of antennas and high-order constellations applications. One such scheme to decrease the complexity, which is referred to as Sphere Decoding (SD) can be employed in this work. The results demonstrated in [30] showed a near performance between the SD algorithm and the ML decoding scheme while keeping the complexity moderately low.

1.1 Problem Statement

In the last decade, the wireless communication technologies have progressed exceptionally and the demand for the technologies will grow drastically in near future. As a way to meet the uprising demand, many studies had been made in the previous research by implementing various schemes, as highlighted at the beginning of this chapter. The concern of the research is to reduce and minimize the error performance of Bit Error Rate (BER) and probability of network outage to achieve full information capacity and more reliable communication transmission particularly in rapidly time varying Rayleigh channels. Stated below are the important problems faced in a wireless cooperative communication system.

- (a) Performance degradation due to high mobility environment

Highly mobility, limited transmission range as well as bandwidth, and unreliable noisy channels create a harsh environment for communication. Most of the existing literature on differential transmission networks assume a flat-fading environment. Though, this consideration is unjustified as cooperative communications are mainly applied in the wireless mobile system, whereby the users are mobile. In a practical scenario such that whenever a user moves with high-velocity, channels eventually changes rapidly in time. Needless to say, this will substantially degrade the performance and the wireless link may totally get disconnected. To overcome the challenges, it is vital to investigate the impact of the deployed differential AF-based cooperative diversity networks under time-varying channel. Hence a random Gauss-Markov mobility model employing Sum-of-Sinusoids (SoS) statistical simulation is adopted to model practical propagation channel.

- (b) In the presence of frequency offsets, the differential schemes experience significant performance loss whereby the flat fading channel does not stay fixed over two consecutive time durations.

Most existing works consider the challenges of developing cooperative diversity system which eliminate the channel knowledge without considering the effect of frequency offsets over any link. In a classical differential system, the channels stay static for at least two symbol intervals. But the effect of frequency offsets cause the block fading to act as a time-varying channel, that violates the concept of a classical differential system, thus degrades the network performance. Hence, to mitigate the problem of channel estimation and frequency offsets, DD modulation and Multiple Symbol Detection (MSD) scheme in AF cooperative diversity system are proposed. MSD is proposed as to enhance the classical differential detector performance which

fails in fast fading channels. By applying the MSD, multiple consecutive symbols are jointly detected. However, far too little attention has been paid to consider both the channel knowledge and frequency offsets in the time-varying channel, particularly in a fast fading environment.

- (c) Imperfect statistical channel knowledge for the combining schemes at the destination over an unknown channel environment.

The MRC scheme acquires the extended or more complex second-order statistic knowledge for all wireless channels at the destination. However, obtaining the second-order statistic is a challenging task in certain applications such as in an unpredictable channel knowledge scenario in the presence of frequency offsets. Therefore it is the interest herewith to develop Selection Combiner (SC) for DDAF networks due to the benefit of SC that eliminates the channel knowledge.

1.2 Research Objectives

The research emphasizes on the differential transmission scheme that bypass the channel and frequency offsets knowledge for cooperative diversity systems. The main objectives are outlined as follows:

- (a) To study and analyze the cooperative diversity schemes in a time-varying channel.
- (b) To develop an improved performance of an Amplify-Forward MSDD transmission for cooperative wireless communication scheme with different mobility.
- (c) To simulate and verify the performance of the proposed schemes in terms of BER and outage probability.
- (d) To compare the proposed schemes with the existing conventional schemes.

1.3 Structure of Thesis

The thesis structure is outlined as follows after a brief introduction to the primary concept of cooperative communications and techniques of diversity.

Chapter 2: This chapter reviews works by other researchers. The basic philosophy of point-to-point wireless communication is reviewed. In order to create the time-varying channel environment, Rayleigh statistical model that focuses on Autoregressive (AR) models are also presented. Then, cooperative communication and relaying protocols, for instance AF and DF are presented. Both one-way relay network as well as two-way relay network schemes are also briefed. The idea of signaling schemes focuses on repetition-based, Space-Time Coding (STC) and DSTBC for wireless communication are elaborated. Then the detection techniques, as well as diversity combining techniques, are reviewed. Coherent detections and non-coherent detection are reviewed but emphases are given to the differential detection for repetition-based transmission and DSTBC.

Chapter 3: The system and channel model of the cooperative communication system is explained. The idea of Multiple Symbol Double Differential (MSDD) employing SD for repetition-based transmission and its benefit in a OWRN is demonstrated. Theoretical BER is analyzed and solved. The network performance optimization corresponding to its power allocation is further investigated. This chapter also presents the results of the proposed schemes and its theoretical numerical analysis. Conventional symbol-based DD detection is compared with the Multiple Symbol Double Differential Sphere Decoding (MSDDSD) over repetition-based cooperative communication system.

Chapter 4: Several existing and proposed combining schemes are investigated. Four combining schemes which include the Direct Combiner (DC), conventional MRC, proposed

semi MRC and SC are devised and elaborated. Then, four combiner schemes are compared for the MSDDSD and it shows that the SC scheme is more robust in fast fading channels.

Chapter 5: In this chapter, the principles of distributed Double Differential Space-Time Block Coding (DDSTBC) scheme for unidirectional relaying scenarios under time-varying channels are further explained and analyzed. The process of encoding and decoding for the suggested DSTBC are also presented. The concept of multiple-symbol transmission is further analyzed for the DSTBC environment under two scenarios : (a) relayed link only without a direct link and (b) relayed link with a direct link. Then, a proposed multiple codeword DD algorithm is devised and studied for OWRN without channel knowledge and frequency offset estimation. The emphasis of the proposed schemes in this chapter is specified to improve the error performance especially in fast fading environment.

Chapter 6: This chapter summarizes the research main contributions as a thesis conclusion. Also, recommendations are presented for future research works.

1.4 Contribution of Research

Numerous contribution are presented in this research. Some of the key contributions can be summarized as follows:

- In this research, a MSDDSD with M -PSK constellation is developed in wireless cooperative networks under time-varying channels in [31], which is used for symbol-by-symbol based cooperative wireless networks in the past literature. The presented detection scheme offers a simple mathematical algorithm that reduces the computational complexity by the previous research and the infeasibility for practical implementation in cooperative communication system, especially in rapidly time-varying channels. Basically, MSDDSD detector using AF relaying protocols is

developed and analyzed bypassing channel knowledge and frequency offsets estimation improve the performance of repetition-based over independent, identically distribution (i.i.d) Rayleigh time-varying channels.

In order to form a time-varying Rayleigh channel, an AR(II) model based on auto-correlation is modeled. The performance metric BER, for the proposed scheme and conventional symbol-by-symbol based DD encoding and decoding M -PSK constellation, are investigated in the context of fast fading channels. In addition, to gain more insight into the system parameters, for instance fading coefficient and the quantity of relay terminals, BER performance is presented. The simulation results revealed that the recommended scheme is able to perform better in terms of error performance with lower computational complexity and significantly reduce the error floor exhibited by the conventional symbol-based transmission scheme.

Approximate BER expressions are further formed for MSDDSD cooperative systems which provide efficient means to assess the performance of wireless mobile cooperative networks with the proposed scheme. Based on the derived numerical expression, the optimization problem is solved by minimizing the BER of the network. Specifically, power allocation between the source terminal and relay terminal(s) are taken into account, in which a closed-form expression is obtained. The results provide a significant improvement by reducing the error floor in comparison to existing conventional scheme. The obtained results and analysis is useful and can be applied for the next generation (4G and 5G) wireless networks.

- Secondly, SC is devised at the destination with repetition based transmission for a fully DD half-duplex cooperative communication network with the effect of frequency offsets. The scheme is able to perform based on Pairwise Error Probability

(PEP) analysis without the need of second order knowledge of the channel, which is required in the MRC scheme. It is demonstrated that the channel gain is significantly achieved by implementing the SC with the additional advantage of simpler implementation in comparison to other diversity combining schemes in the realistic wireless networks. Whereby, MRC is unable to deliver the required performance, because of the channel noise variance that presents at the destination. Theoretical numerical analysis is derived for the SC scheme and its performance are verified through computer simulation.

- Thirdly, multiple codeword SD employing DD transmission with DSTBC is developed under time-varying Rayleigh channels in [32]. Specifically, the signals are double differentially modulated based on PSK are transmitted and DSTBC is constructed during the relay-destination transmission. The proposed DDPSK with DSTBC scheme has the ability to expedite communication simultaneously in a non-coherent approach when channel knowledge and frequency offset estimation is difficult to execute during both transmission periods. Recovery of differentially-encoded information utilizing block decoding algorithm could produce significant improvement in terms of error rate.

CHAPTER 2

LITERATURE REVIEW

2.1 Introduction

In the previous chapter, the significance of cooperative communication has been briefly explain for the future wireless communication system, especially in mobile networks. This chapter further presents the concept of a wireless communication system which includes the single link and cooperative communication networks and its relaying protocols. Channel modeling such as the statistic equation and the related parameters are also studied and investigated in order to model time-varying channel environment for simulation and analysis purposes. The literature on the existing coherent, non-coherent and differential detection techniques in repetition based and DSTBC cooperative communication networks are also presented. Emphases are given to the differential detection, in which channel knowledge and frequency offset estimation are bypassed for the above-mentioned environment. In addition, this chapter also detailed the diversity combining techniques such as direct combiner, MRC and SC for multiple relays cooperative networks. The flowchart of the literature review is shown in Figure 2.1.

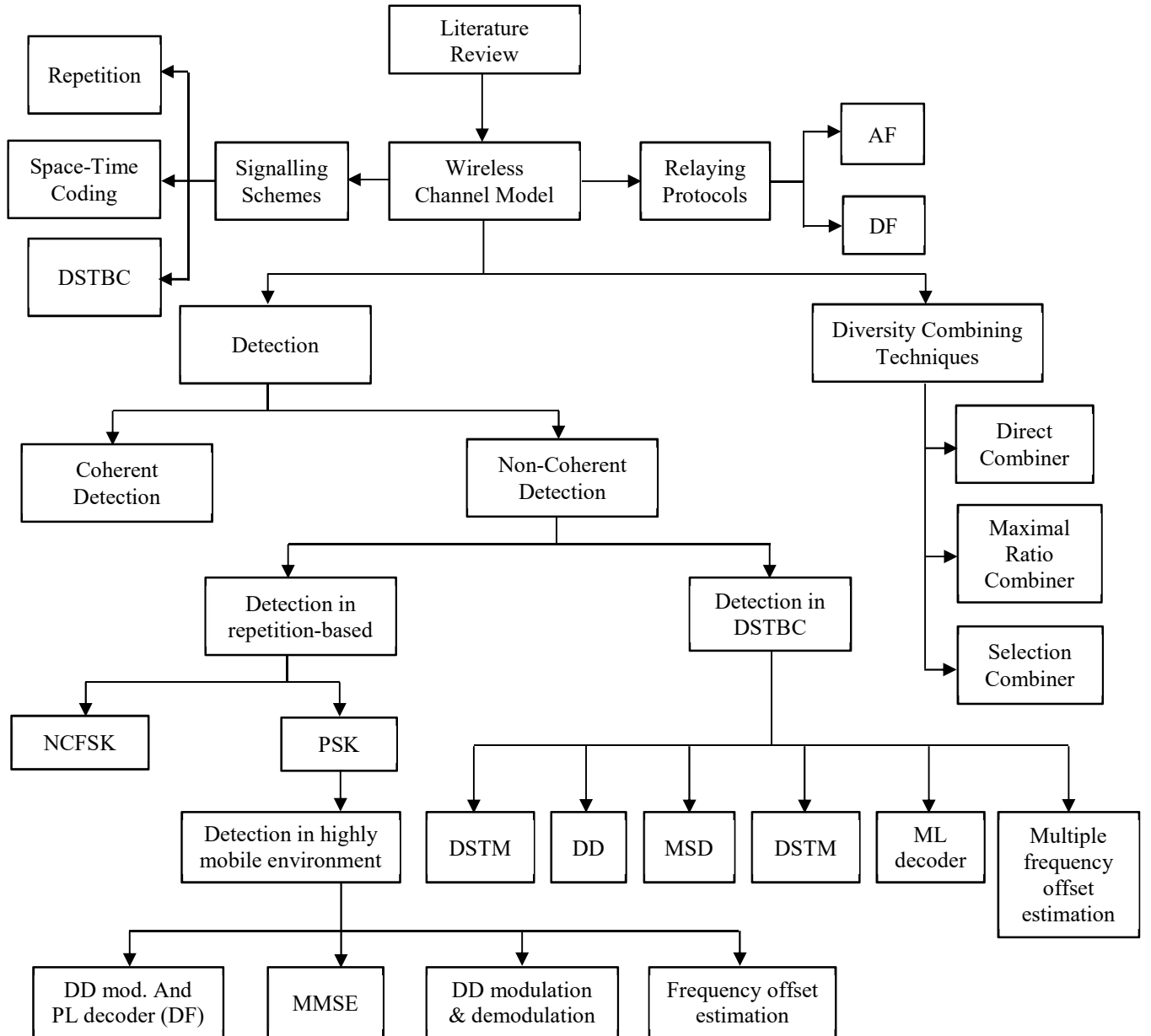


Figure 2.1: Flowchart of Literature Review

AF relaying approach are extensively analyzed in the perspective of the coherent and non-coherent detection system. In the coherent systems, basically instantaneous CSI knowledge from the channel estimation scheme is considered perfect in a quasi-static or slow fading environment. Yet, for the fast fading scenarios, non-coherent detection

techniques is preferable with the assumption that only the statistics of channel and fading distribution is known, whereas instantaneous CSI is unavailable. Therefore Non-Coherent Frequency-Shift Keying (NCFSK) and differential transmission technique are the effective techniques that do not require the channel knowledge of all the communication links in the networks. Although the differential modulation and demodulation provides better performance in terms of bandwidth and the power management, the system suffers a performance loss as compared to the perfectly estimated channel knowledge in coherent detection system. On the other hand, in practical scenario which involves mobility, it is a challenging job to acquire the perfectly known channel information [11]. Thus, the reliability of the wireless communications possibly is not achievable because of the inaccurately estimated channel coefficient.

The chapter outline is presented as follows. Section 2.2 describes the single link wireless communication and wireless channel model is outlined in Section 2.3 in order to better define the basic aspects of wireless communication networks in this thesis. Section 2.4 discusses on various cooperative relaying protocols to be incorporated into the wireless cooperative networks which is explained in Section 2.5. The next part of the chapter, Section 2.6 provides an overview of available diversity combining techniques. Section 2.7 outlines the state-of-art of cooperative diversity and the existing research in terms of detection techniques in repetition and STC strategy are further discussed in Section 2.9 and Section 2.10. The analysis of the review are concluded in Section 2.10. Finally, Section 2.11 summarizes this chapter.

2.2 Single Link Wireless Communication

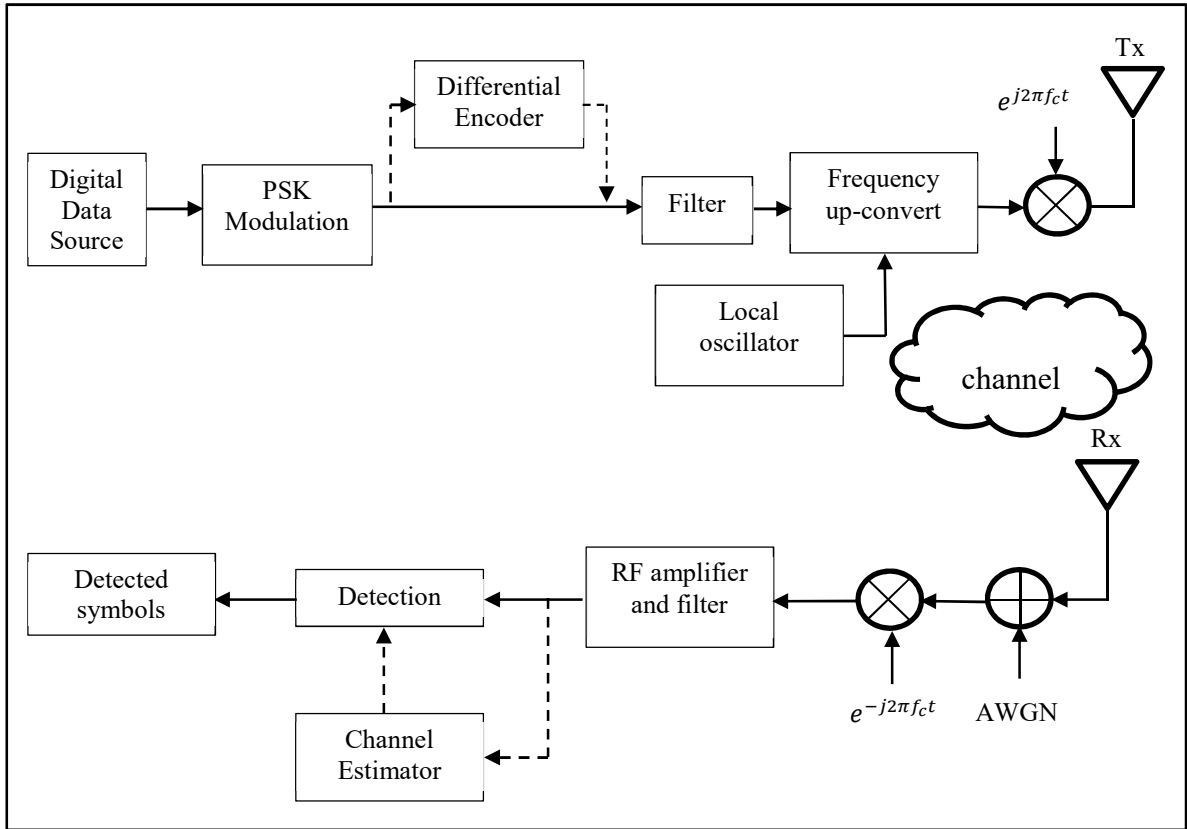


Figure 2.2: A simple single link wireless communication block diagram

Figure 2.2 shows a point-to-point wireless communication path in which the source transmits signals towards the desired destination via a channel. In general, there can be various data information such as the analog (i.e. video or audio), and digital (i.e. words or multimedia) signals. Information bits are then transformed into symbols using the modulation schemes, for example the Amplitude-Shift Keying (ASK), Frequency-Shift Keying (FSK) or Phase-Shift Keying (PSK) for transmission over a physical channel of interest. Among the modulation schemes, PSK is a preferred modulation. PSK is broadly applied in the wireless technology such as satellite communications, Bluetooth, wireless LANs, RFID and etc. This is because PSK exhibit persistence envelope property and high

bandwidth performance. A number of data bits, $\log_2 M$ is modulated into symbols that is represented by the particular symbol phase $\mathcal{V} = e^{j\frac{2\pi m}{M}}$. M represents the number of signal points equally spaced on the circle. Figure 2.3 shows the space plot of Binary Phase-Shift Keying (BPSK) and Quadrature Phase-Shift Keying (QPSK).

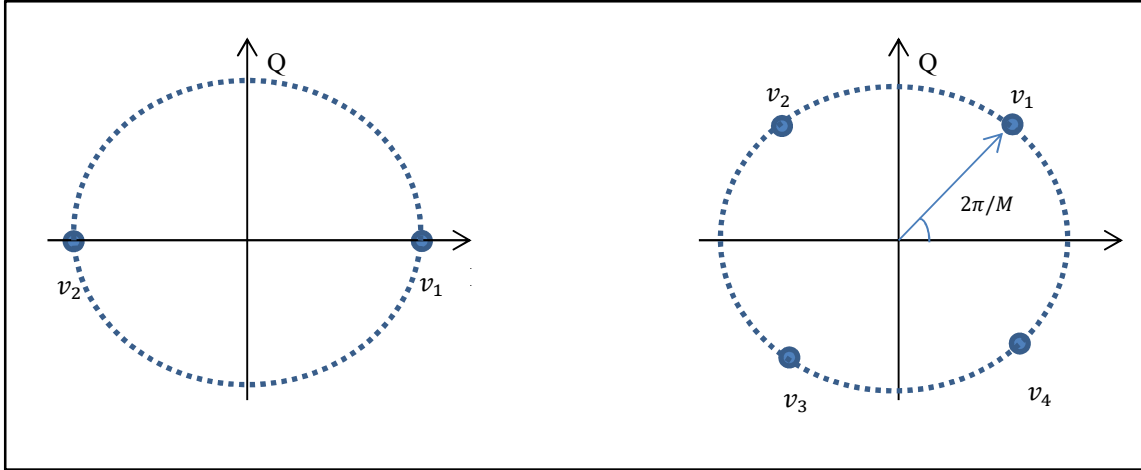


Figure 2.3: a) BPSK; b) QPSK constellation diagram

Referring to Figure 2.2, for coherent detection the differential encoder block can be eliminated. But, the performance of the coherent detection becomes infeasible due to the complexity channel estimation for practical applications. Thus, differential M -DPSK is introduced to eliminate the estimation of channel coefficients. Then, the discrete symbol is converted into Radio Frequency (RF) signals using the local oscillator and propagated via the wireless channel. During the propagation, the signal will be affected by both small scale as well as large scale fading. Therefore, it is important in providing a numerical model for a fading channel.

2.3 Model of Wireless Channel

In wireless communications, its channels may be affected by the propagation environment encountered such as fading and the mobility effects of the mobile terminals. As the results, the propagating signals may experience rapid fluctuation. Due to that reason, the wirelessly transmitted signals that propagate through the unfavorable random channels does not allow for simple and ideal Additive White Gaussian Noise (AWGN) assumption. Thus, it is important to study and model the fading channel for simulation purposes. The wireless channel model $h(t)$ is based on linear time-varying systems as in [33] to avoid the complexity of dealing with electromagnetic equations.

$$h(t) = \sum_i a_i(t)x(t - \tau_i(t)) + w(t) \quad (2.1)$$

where $a_i(t)$ denotes the weight, $\tau_i(t)$ represents the i th channel delay and $w(t)$ is the noise. Referring to the equation model in (2.1), the channels can be described with regards of two dissimilar degradation groups, frequency-selective fading or flat fading dependent on the propagation delay and symbol time relationship. A channel is categorized as frequency-selective if the propagation delay is larger than the symbol time. On the contrary, a channel is said to exhibit flat fading if the propagation delay exhibit lesser value than the symbol time. In the flat fading case, all the frequency components of the received signals vary in the same proportion simultaneously as the channel has only one filter tap. This research focuses on the flat fading channels.

Flat fading channels can be divided into two, namely slow fading and fast fading. In fast fading, the bandwidth W or symbol rate $\frac{1}{T_s}$ is less than the normalized frequency f_d or fading rate $\frac{1}{T_0}$ (i.e.), which can be expressed with:

$$W < f_d \quad (2.2)$$

or

$$T_S > T_0 \quad (2.3)$$

Since $T_S > T_0$, it is known that the channel experiences fading while a symbol is propagating, causing distortion or attenuation to the received signals. On the other hand, in slow fading channel the bandwidth is larger than the normalized frequency, that is:

$$W > f_d \quad (2.4)$$

or

$$T_S < T_0 \quad (2.5)$$

Thus, it can be described that the channel condition stays static throughout the period of a symbol transmission. The transmitted symbol is unlikely to suffer from the pulse distortion. In general, slow fading means that each one symbol time or more having the same multipath channel effect whereas fast fading means that one symbol will have a different effect over the time. Since the received wireless signals vary in a flat fading channel, it is vital to determine the channel coefficient distribution accurately in designing the wireless communication networks. The commonly used statistic models of signal amplitude distributions for the channel coefficient are Rayleigh, Rician and Nakagami. In this research, Rayleigh flat-fading distribution is adopted due its ability to generate arrival of multiples reflections at the receiver. In addition it is applicable for radio signals propagation, in the heavily built-up urban environments whereby Line-of-Sight (LOS) path between both transmitter and receiver antenna is not applicable.

2.3.1 Rayleigh Statistical Model

Let $\{h_i(t)\}_{i=1}^N$ represents the i.i.d Rayleigh random process, each distributed by:

$$f_{h_i}(h) = \frac{2h}{\Omega_i} \exp\left(-\frac{h^2}{\Omega_i}\right), \quad h \geq 0 \quad (2.6)$$

where $\Omega_i = E\{h_i^2(t)\}$ represents the i -th random process average power. $f_{h_i}(t)$ denotes the distribution of the signal envelopes over the scattering radio channel in the presence of Doppler shift - $f_c v/c$ with f_c represents the carrier frequency while v and c denote the velocity of terminals and speed of light, respectively. The process is considered to be a time-correlated random process that is dependent on some Doppler spectrum. The Doppler spectrum varies according with two different channels, stationary to mobile channel or mobile to mobile channel. Nevertheless, it was noted that the i -th envelope time derivative is independent with its envelope in both cases. Thus, the Gaussian Probability Distribution Function (PDF) $f_{x_i}(x)$ can be expressed as [34]:

$$f_{x_i}(x) = \frac{1}{\sqrt{2\pi}\sigma_{x_i}} \exp\left(-\frac{x^2}{2\sigma_{x_i}^2}\right), \quad x^2 \geq 0 \quad (2.7)$$

where $\sigma_{x_i}^2 = \pi^2 \Omega_i f_i^2$ is the variance.

For a stationary to mobile channel,

$$f_i = f_{dmax} \quad (2.8)$$

where f_{dmax} denotes the maximum Doppler frequency at the mobile terminal. On the other hand, for a mobile to mobile channel environment,

$$f_i = \sqrt{f_{dmax}'^2 + f_{dmax}''^2} \quad (2.9)$$

where f_{dmax}' and f_{dmax}'' represent the maximum Doppler frequency for both transmitting and receiving mobile terminals, respectively.

2.3.2 Autoregressive models

AR models are important and commonly used to mathematically model the channel coefficients of the discrete-time random processes. The model is useful for simulation and performance analysis purposes, particularly in time-varying wireless communication

systems due to its simplicity yet effective approach in determining the channel correlation property. A p -th order AR (AR(p)) in time domain can be expressed as:

$$x[n] = - \sum_{k=1}^p a_k h[n-k] + w[n] \quad (2.10)$$

where $w[n]$ denotes the complex white Gaussian noise. For Rayleigh faded environment, $w[n]$ has zero mean with $|h[n]|$ as output, variance of σ_p^2 and independent of $h[n-k]$. a_k is the filter coefficients and also known as the auto-correlation value of the channel. In practical environment, it is desirable to attain a solution that is uncorrelated with time and at the same time act as a function for the previous error terms. In cooperative communication network, uncorrelated channel is required in order to avoid Inter-Symbol Interference (ISI) and Co-Channel Interference (CCI). Since the solution is uncorrelated with time, initial condition value is not required. However, previous error terms are necessary for forecasting reasons. In this research, the second-order AR model, AR (II) is implemented.

2.4 Cooperative Relaying Protocols

In a cooperative diversity system, the transmitted signals from the transmitter are processed by the relay(s) and the processed signals are re-forward to the intended destination. The relaying protocols are mainly categorized into two classes, DF and AF depending on the processing styles.

2.4.1 Decode-and-Forward (DF)

DF is also known as a regenerative relaying protocol that modifies the information (symbols) and requires more complex baseband operation [5]. In DF relaying protocol, the relay firstly demodulates/decodes the receiving source signals and then re-modulates/re-

encode the messages before forwarding towards the desired destination. The DF relaying protocol is ideally used when the communication channels are in Gaussian, which is suitable for urban cellular mobile and wireless LOS communication. In other communication channel networks due to the demodulation/decoding process at the relay, it is undeniable that DF relaying protocol introduces error propagation and time delay limitation because of the transmission of erroneous signal that weaken the whole system. In addition, the relays may require perfect regeneration and retransmission process from the source. Thus, this protocol is unfitted for time-delay wireless networks.

2.4.2 Amplify-and-Forward (AF)

Generally, the idea behind the AF protocol is simple as the decoding and encoding process is circumvented. AF is a transparent relaying protocol that allows for simple power scaling and phase rotations without modifying the information by utilizing advanced signal processing techniques. By implementing the amplification process, the noise in the signal is also being amplified at the same time, which is the main downside for the AF relaying protocol. Yet, this relaying protocol is still preferable and has the potential advantages over DF relaying protocol. Firstly, AF relaying protocol is simple and suitable for application that requires less signal processing burdens as well as low power consumptions. Secondly, the application of a AF relay does not affect the operations at the source and the destination due to the reason that no handshaking process is required for each information signals going through the relays. Thirdly, an AF relay does not expose to security information threat even if eavesdropping occur because virtually the relay does not contain any information for source decoding process. Fourth, an AF relay can process and relay the signals faster in comparison to a DF relay. Last but not least, the attractiveness of this relaying protocol is the error free

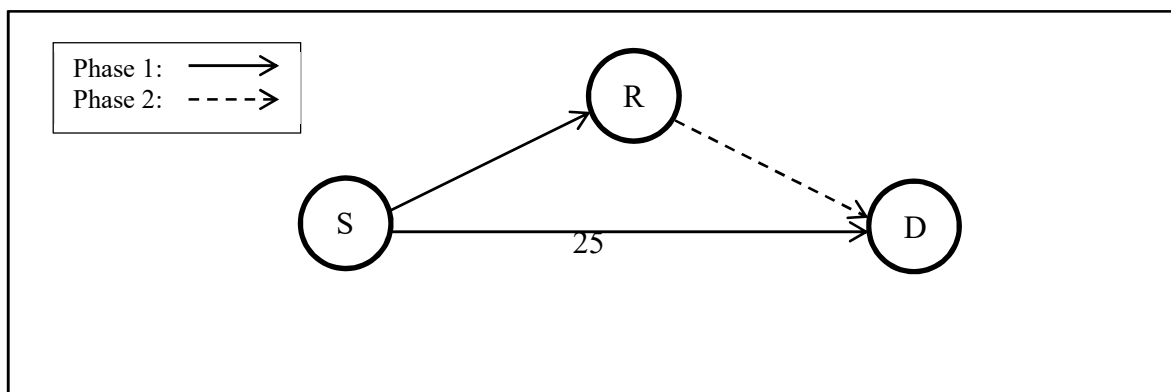
from the signal propagation at relays as compared to the DF relaying protocol wherein decoding process is not required. These promising benefits could make AF relays useful in mobile ad hoc networks and/or wireless networks. Thus, AF relaying protocol is the focus of this research.

2.5 Wireless Cooperative Networks

The main concept of the wireless cooperative networks is to allow the source(s) to transmit information signals via multiple independent wireless path towards its desired destination with the aid of other terminal(s) that perform as relay(s). A cooperative network involving a single relay and multiple relays are depicted in Figure 2.4 and Figure 2.5. The transmission schemes can be categorized as OWRN or Two-Way Relay Network (TWRN) [27,33] as shown in Figure 2.4(a) and (b), respectively.

In OWRN, the transmission process is typically distinguished into two phases, namely the relayed and broadcast phase. During the broadcast phase, the source transmit the data signal towards a relay and at the same time directly towards the destination. Then, in the second (i.e. relayed) phase, the source stays idle. Concurrently, the relay processed the received signals with the amplification process and reforwards the processed signals to the desired destination.

On the other hand, for TWRN transmission, it is often considered that there is no direct communication between the two users (S_1 and S_2). The two users transfer information with one another through a relay. The relay then superimposed (summed) the received signal from the users and convey the processed signal back to its intended destination [29, 34].



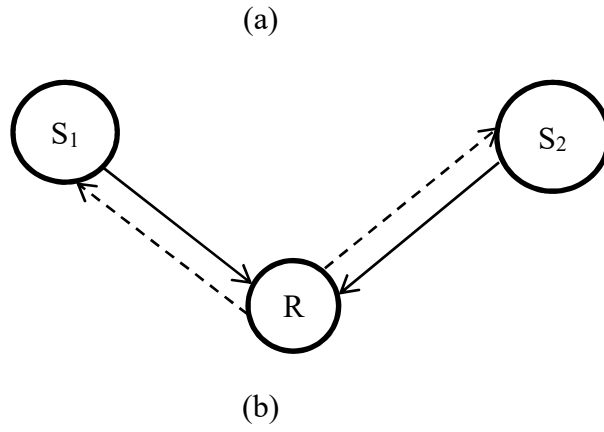


Figure 2.4: a) One-Way Relay Network (OWRN) and (b) Two-Way Relay Network (TWRN) [36]

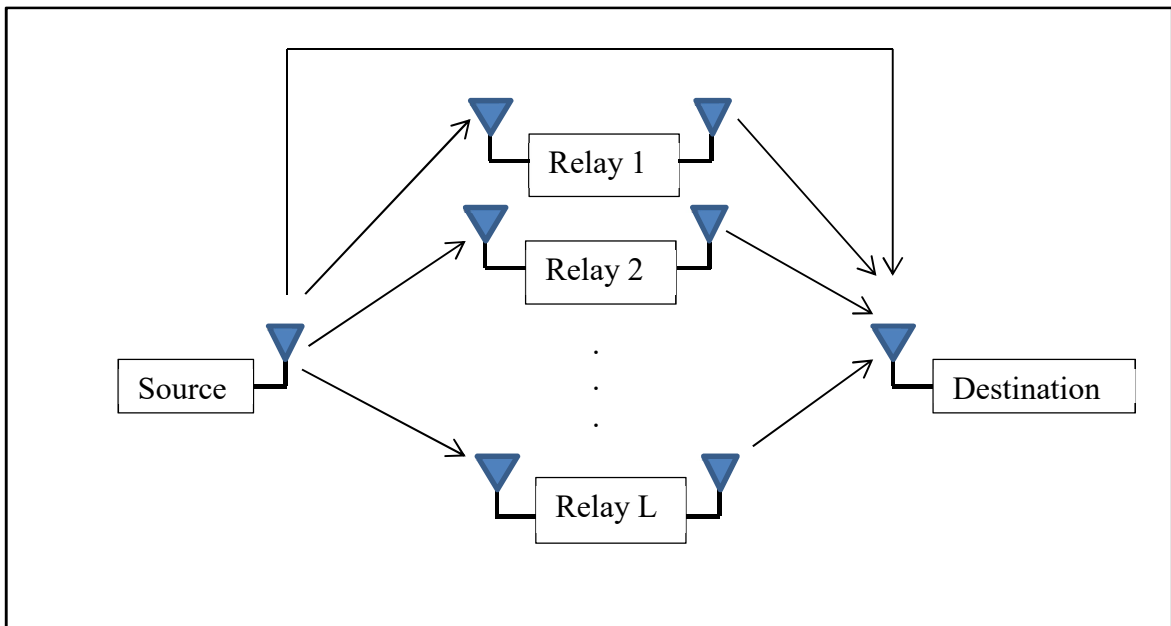


Figure 2.5: Illustration of multi-relay cooperative communication system

2.5.1 Signalling Schemes in Cooperative Communication Networks

The signalling schemes for cooperative communication networks can be divided into two main category, namely repetition-based and STC as refer to the next sub-sections.

2.5.1.1 Repetition-Based Transmission

In order to transmit data signals that may be corrupted over a fading channel, the main idea of repetition-based transmission is where antennas transmit similar signals in different time slots to the destination following the TDD approach.

2.5.1.2 Space-Time Coding

STC-based transmission has been commonly investigated and integrated into massive MIMO systems in early 90s pioneering research on transmit diversity. STC is a method employed time and space diversity to enhance the data transmission reliability in wireless communication systems. The main attraction of this technique is to send multiple copies of information simultaneously through numerous antennas and exploit the multiple received data versions. Furthermore, the diversity gains and higher spectral efficiency is achieved whereby each cooperative node transmits a column of encoded signal across both time and space domains. The STC is categorized into two main groups that are the trellis coding and block coding. The basic concept of STC was introduced by [37] based on trellis code and is further developed by Alamouti in [38] with two transmit antennas using the simplest block codes to achieve full diversity.

Space-Time Trellis Codes (STTC) work based on a single input symbol distributed over time that produces a vector symbols sequence whereby the length signifies antennas. Similar to traditional Trellis Coded Modulation (TCM), space-time trellis codes offer

excellent coding gain. Yet, both space-time trellis codes and TCM are extremely hard to design where computationally in-depth encoder and decoder are required.

Space-Time Block Code (STBC) is designed based on a block basis such that the matrix output is represented by the columns (time) and rows (antennas). Unlike conventional single-antenna schemes, most STBC offers only diversity gain and not coding gain. One particularly attractive feature of STBC is that it can achieve full diversity with exceptionally low encoder/decoder complexity, for example Alamouti code. A simple baseband encoding and transmission principle of Alamouti is demonstrated in Figure 2.6.

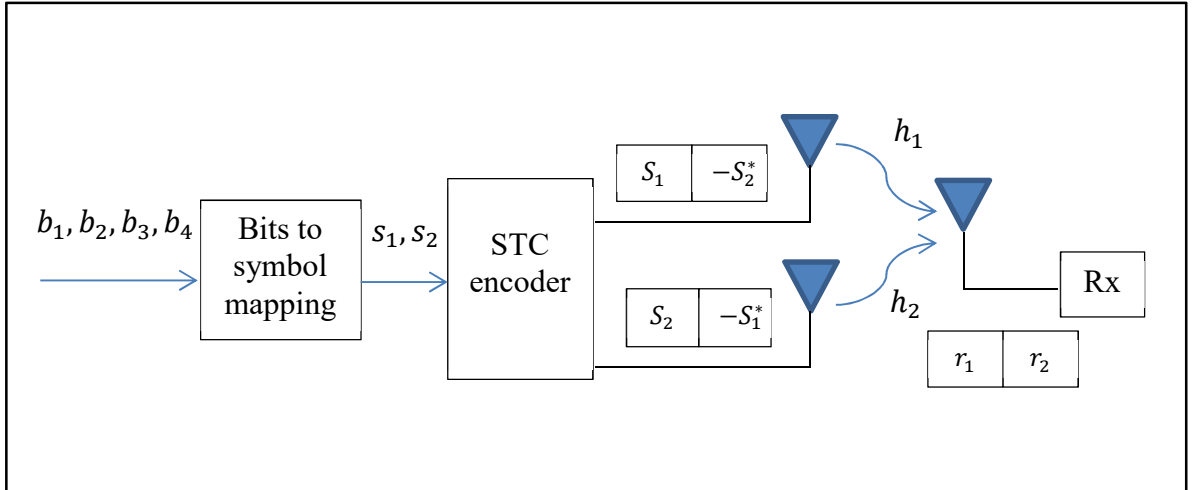


Figure 2.6: A block diagram of the Alamouti two-branch transmit diversity scheme

Referring to Figure 2.6, m number of information bits (i.e. b_1, b_2, b_3, b_4) are mapped employing an M -ary modulation approach whereby $m = \log_2 M$. The encoded M -ary symbols are then grouped by STC encoder into set of pairs that represents a valid space-time code matrix expressed as:

$$\mathcal{S} = \begin{bmatrix} S_0 & -S_1^* \\ S_1 & S_0^* \end{bmatrix} \quad (2.11)$$

where s_i^* indicates the complex conjugate of s_i for $i = 1, 2$. After the STC matrix is generated, it is delivered to the source antennas. Then the coded signals are transmitted towards the desired destination in two time slot. At time slot 1, both antennas, Antenna 1 and 2 transmit S_1 and S_2 simultaneously. In time slot 2, both Antenna 1 and 2 transmit $-S_2^*$ and S_1^* at the same time. It is generalized that the codeword matrix row denotes that the encoding process happens at various antennas in the space domain whilst the column signifies the transmission interval in time domain. With this, the encoder considers two continuous block of modulated symbols at a respectively period. The two-antennas transmit diversity scheme is known as STBC. The Alamouti-based STBC is orthogonal due to the property of $s_0 s_1^* = s_1^* s_0$ by obtaining the product of the antennas transmit sequences.

At the destination, the Alamouti receiver consists of three main components: channel estimator, combiner and decoder. Channel estimator works as an estimator to find the channel coefficient gain and the combiner combines all the available copies of the transmitted information signal so as to mitigate the fading. The combined signals are then decoded to determine the transmitted symbol applying minimum distance. For the decoding process of Alamouti code at the destination, ML gives the best performance at the cost of complexity. A well-known simple two-branch transmit antennas over the fading channels h_1 and h_2 by Alamouti [38] is depicted in Figure 2.7.

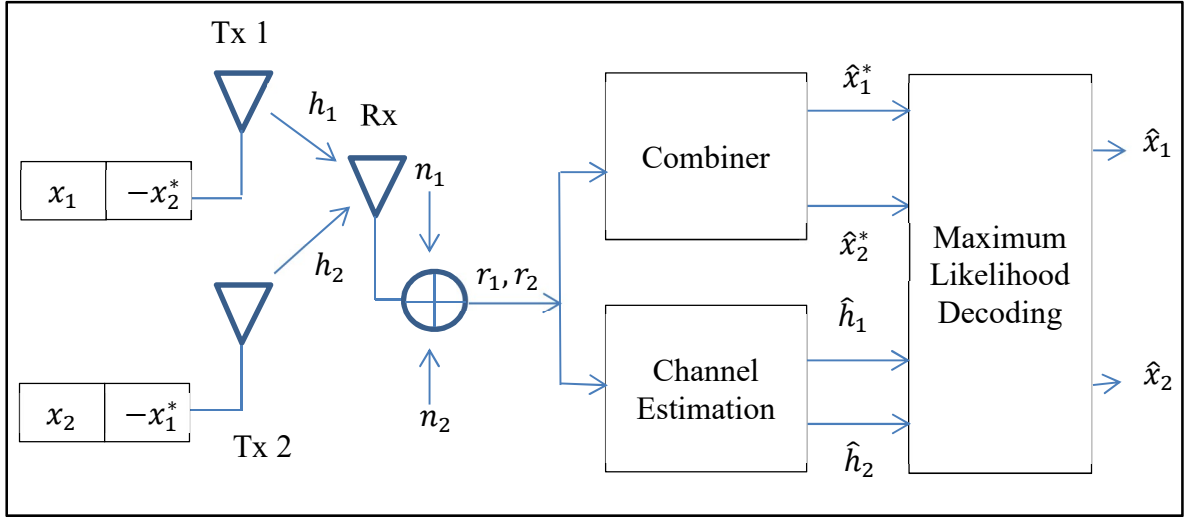


Figure 2.7: Block diagram of STBC decoder for the point-to-point communication network

At the destination, the consecutive two received sequence signals can be expressed as:

$$r_1 = r(T_1) = h_1 x_1 + h_2 x_2 + n_1 \quad (2.12)$$

$$r_2 = r(T_1 + T_2) = -h_1 x_2^* + h_2 x_1^* + n_2 \quad (2.13)$$

The received signals are combined and can be written as:

$$\hat{x}_1 = h_1^* r_1 + h_2 r_2^* \quad (2.14)$$

and

$$\hat{x}_2 = h_2^* r_2 + h_1 r_1^* \quad (2.15)$$

Then the combined signal process through ML detector to form decision variables.

2.5.1.3 Distributed Space-Time Block Code (DSTBC)

When STBC is applied to wireless cooperative communication networks, it is known as DSTBC. The cooperative network forms a virtual MIMO system instead of real physical MIMO system. The cooperative mechanism in DSTBC is completed via two transmission phases as shown in Figure 2.8.

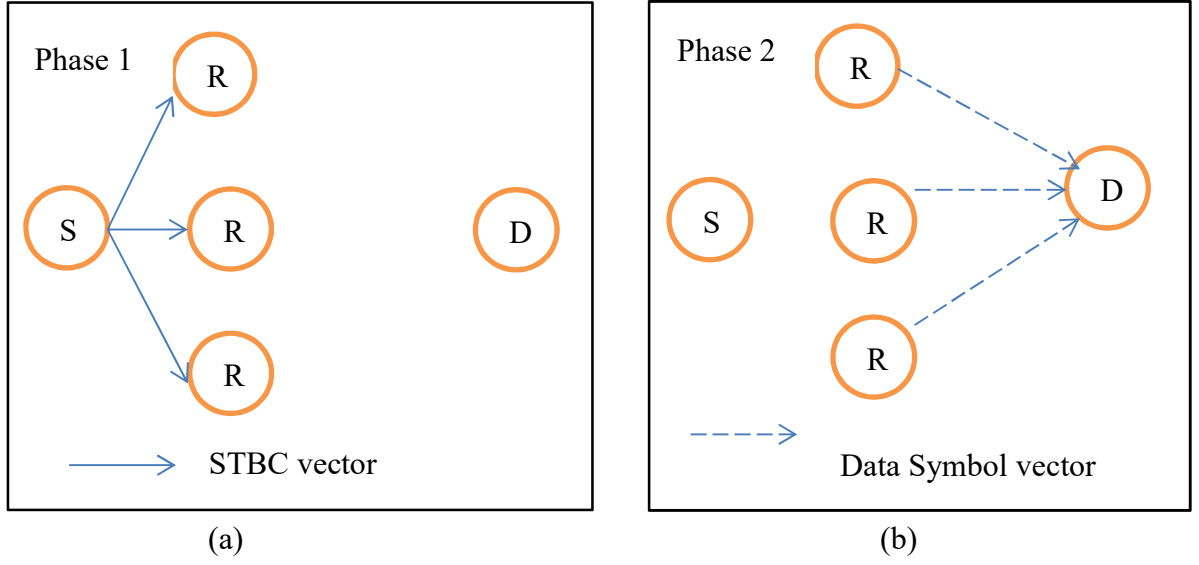


Figure 2.8: DSTBC transmission (a) phase I; (b) phase II

In the DSTBC scheme, cooperative terminals encode the signals based on STBC during phase I transmission shown in Figure 2.8(a). Each terminal transmits a column of the block code and DSTBC coding can be used during the second phase transmission as depicted in Figure 2.8(b). This scheme is beneficial for making effective usage in terms of available bandwidth and resources of power [8, 37] by transmitting signals in sequences of specified codes (i.e. STBC).

2.6 Diversity Combining Techniques

As mentioned earlier, the principle of diversity is to improve the reliability of the signal transmission (i.e. to increase the overall received SNR) in a wireless networks by sending multiple copies of the similar information via multiple independent fading paths. In this thesis, space diversity is considered wherein the distance between the receiving antenna is large enough to ensure independent fading. With the assumption that the fading stays independent from one component to the other and identically distributed, different versions

of the same transmitted signals are received at the destination. In other words, numerous number of received signal sequences copies are subjected to different level of statistical corruption that may be additive and/or multiplicative because of fading severity. Thus, the statistical probability of all the signal copies being in deep fade, apparently turn out to be lower. Diversity is attractive because it is unlikely all the M copies of the same signal experienced deep fade. At least one copy will have adequate power so that the receiver at the destination is able to further process the signal. Due to the multiple independent samples received, diversity combining is a very important aspect in order to increase the SNR and simultaneously, reduce the error performance.

Diversity combining refers to the method of combining the multiple received signals carried over different paths at the destination. There are several ways of categorizing diversity combining methods. The diversity combining state-of-art can be referred in [40]. Amongst the diversity combining techniques, direct combiner, MRC and SC are categorized as the common techniques in practical used. The mutual goal of the three diversity combining techniques is to determine a set of weighting factor, ω for each receiving signals of different paths. The goal for choosing the suitable weighting factor is to minimize the consequence of channel fading. The basic principle of the combining schemes is discussed in the following subsections.

2.6.1 Direct Combining

Figure 2.9 shows a basic three-channel direct-gain system. The received multiple signals at the destination are combined with precisely the same multiplication factor, $w_1 = w_2 = w_3 = 1$ that compensates the phase variation of the channels.

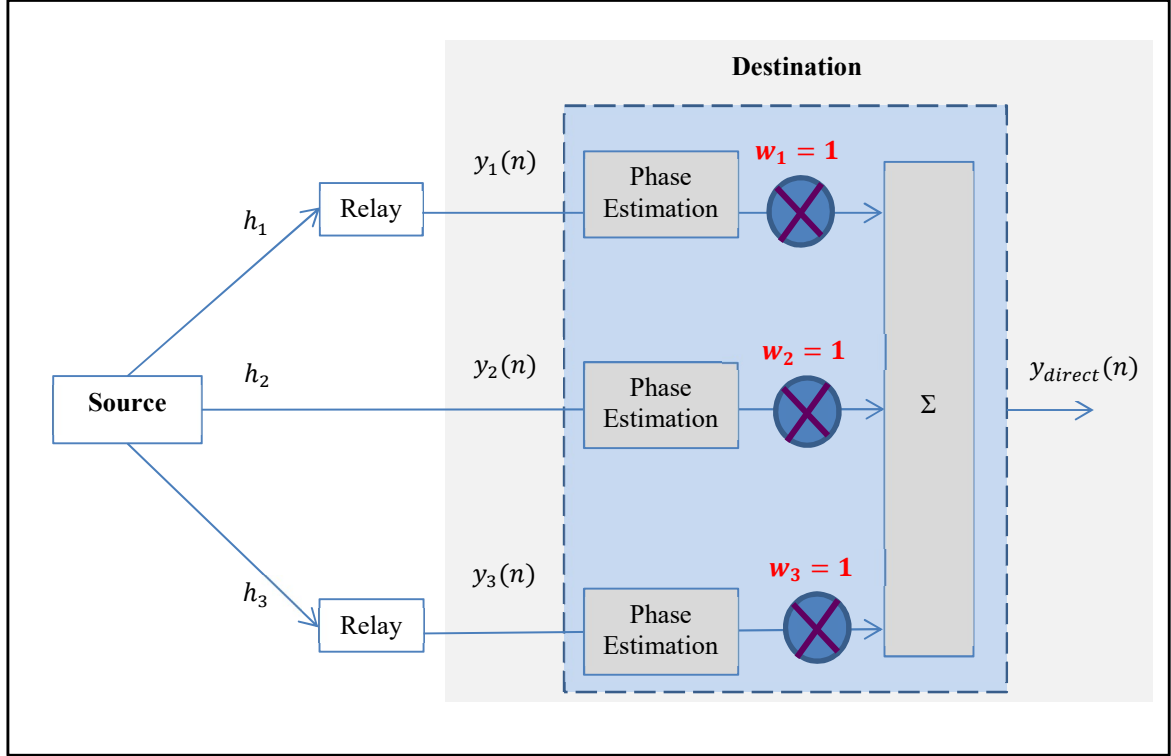


Figure 2.9: General block diagram of direct combining scheme for cooperative communication network

2.6.2 Maximal Ratio Combining

MRC requires complete knowledge of channel branch gains. In MRC technique, the receiving individual signal path is first co-phased and weighted proportionately to their channel gain. The weighting factor of each path can be obtained by estimating the complex conjugate of the channel gain. Then, each weighted signals are summed or combined. Figure 2.10 depicts the general block diagram of the 3-branch MRC. Each gain w_a , w_b and w_c is proportionate to the phase estimation.

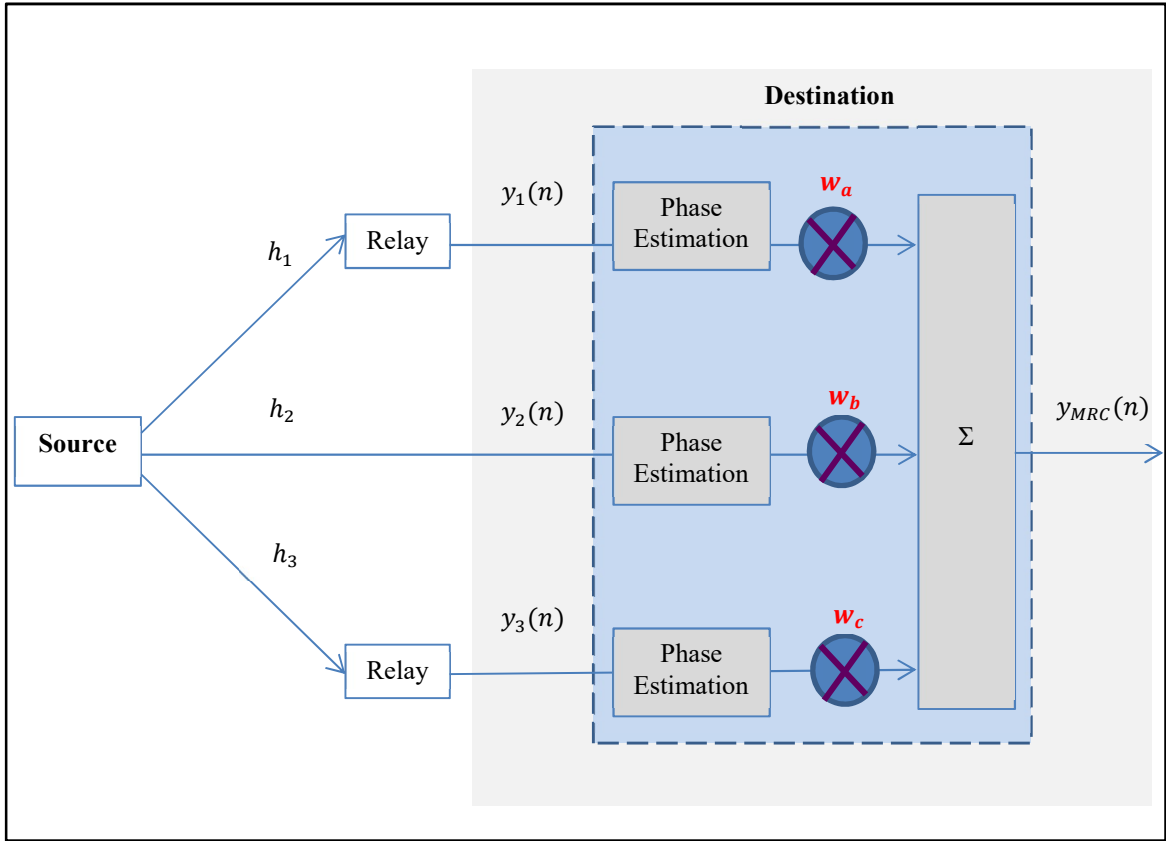


Figure 2.10: General block diagram of MRC scheme for cooperative communication network

2.6.3 Selection Combining

Generally, SC scheme is the simplest as compared to MRC and direct combiner schemes. Basically, the idea of SC is to select the path with the largest received SNR for further processing at the destination as depicted in Figure 2.11.

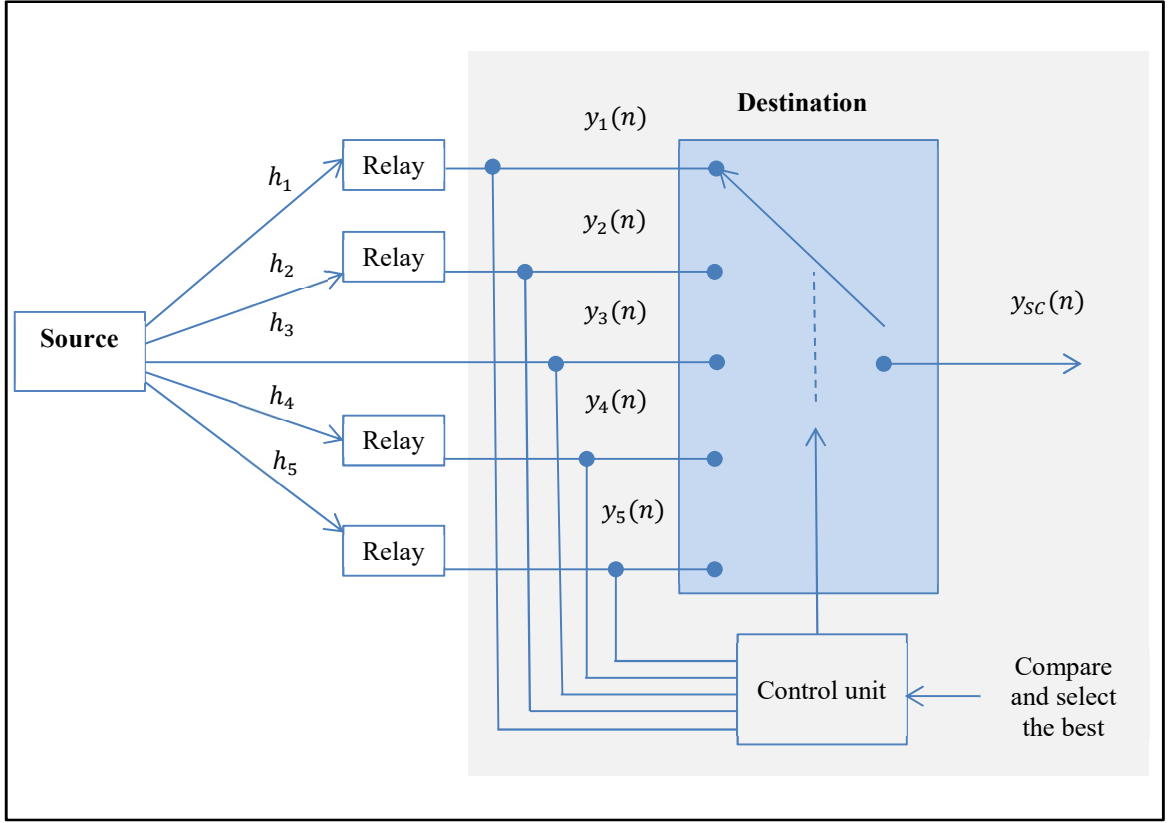


Figure 2.11: General block diagram of SC Scheme for cooperative communication network

Referring to Figure 2.11, the decision rule in SC scheme can be expressed as:

$$\omega = \begin{cases} 1 & \gamma_n = \max_i \{\gamma_i\} \\ 0 & \text{otherwise} \end{cases} \quad (2.20)$$

As the scheme chooses the path with the maximum SNR, the output of the SC scheme can be expressed with $\gamma = \max_i \{\gamma_i\}$. Thus, the scheme only requires the measurement of the signal power for each path received. Unlike the MRC and direct combining schemes, the estimations of phase shift or variable gains are not required. Hence, the SC scheme is the most appropriate scheme to reduce the probability of error at the destination, specifically in the fast fading environment since it is a difficult task to acquire the perfect channel estimation and frequency offset estimation in such scenario.

2.7 Cooperative Diversity Chronology

In general, the literature on cooperative relaying has seen a tremendous rise in contributions in recent years. Through cooperative relaying, coverage extension, increased reliability, and diversity can be harvested. The early thought of cooperative diversity derived in 1971 from the work of [41]. His work introduced the concept of three terminals channel model, comprised of the source, destination and relay deviating from the direct point-to-point data transferring (i.e. the conventional communication) in wireless communication environment. It was exposed that the system with the aided relay improved the data rate of the receiver. Due to the promising performance of the relayed network, Cover and El Gamal in [42] involved the relay channel capacity theorem within a single transmission system with AWGN channel in obtaining the relay channel capacity. Three theorems were contributed to determining the relay channel capacity based on the (i) degraded relay channel, (ii) reverse degraded relay channel and (iii) feedback relay channel. It is noted that the three channel models as well as the applications, mainly assumed single-hop single user wireless communication. The studies were therefore further considered in the multi-users environment and have been studied in [43] for the real-discrete-time Gaussian parallel relay network. The other multiple relays works are included in [42, 43]. Sendonaris et. al. [44, 45] generalized the multiple sources-relay networks and proven that two users cooperation gained beneficial focusing in cellular networks environment as the results performed higher data rate and were unaffected by the channel variation. All the related works mentioned above discussed on the great achievement of single and multi-users channel cooperation which became the root for the development of cooperative diversity systems.

Generally, the concept of cooperative communication incorporates the cooperation, relaying and combining methods among the nodes to enhance the performance of the

information transferred. Various relaying strategies such as AF, DF, and CF were studied. The root underlying of the relaying cooperative protocol that is widely referred to, namely DF and CF were investigated in the seminal work of [42]. The protocol is to encourage the cooperative and sharing of virtual antenna arrays without extra hardware costs. Another approach of relay cooperative methods, which includes AF and adaptive DF were further outlined in [9] pertaining to the measurement of the channel between the terminals and the feedback from the destination terminal. The study has taken account the outage behavior and the SNR performances to achieve full diversity with energy saving. Many studies focus on AF because of its simplicity. The relay protocol amplifies the received source signal with multiplication gain and reforwards the processed signals to the intended receiver.

In the past decade, wireless relay networks based on coding approach has progressed greatly. A Code Division Multiple Access (CDMA) cooperative transmission has been first presented in [2]. It is demonstrated that the protocol is enable to achieve full diversity as well as full data rate at the expense of relatively high complexity relay node. The coded approach cooperative transmission is further studied and analyzed based on DSTC. Laneman and Wornell in [8] proposed DSTC incorporate the DF relay network while Nabar, Bolcskei, and Kneubuhler in [48] studied the DSTC for an AF relay network. However, both studies resulted in high decoding complexity. The discovery DSTBC approach was first suggested by Tarokh and Jafarkhani [15] employing two transmit antennas. Due to the excellent performance of the DSTBC transmission, it has motivated researcher to further investigate the said schemes.

2.8 Detection Techniques for Cooperative Diversity under Repetition-Based Transmission

The discussion in this section starts with the review and comparison of various detection techniques. The literature is reviewed on the general coherent and non-coherent detection process in wireless cooperative system and in particular with emphasis on differential detection under repetition-based transmission.

2.8.1 Coherent Detection

In the coherent detection circumstances, full CSI knowledge (i.e. the carrier phase information) of all the wireless paths in the network is required and must be known at the desired destination. Additionally, matched filters are required to detect and decide on what the original data was sent. In order to obtain the CSI knowledge, channel estimation is necessary by transmitting training signaling or pilot symbols from each node to the intended node(s) beforehand to the original data transmission. Even though the coherent detection scheme which includes training/pilot signals offers excellent outcome for the immobile and slowly faded channels environment, the scheme fails to perform in high mobility applications. Moreover, the scheme becomes less effective by yielding an irreducible error floor.

In other words, it is challenging to estimate the channel perfectly due to the rapid changes of the surroundings as the previously estimated channel knowledge become out-of-date after a number of symbol transmission period. The estimation needs to be refreshed frequently and it causes inaccurate estimation. Therefore, in order to deliver reliable channel knowledge estimates under fast fading channels, an extra trainings are required which in turn reducing the achievable data rate. Moreover, the difficulty of the channel estimation

increases in multiple relay systems that consume extra power and introduce processing delay. Since the mobile communication systems impose stringent limitations on the spectrum resources as well as available power, it is imprudent to offer transmit power and valuable bandwidth to transmit extra training/pilot symbols. Thus, this phenomenon makes coherent detection scheme becomes less attractive in practice. Next subsections commences from a brief introduction of differential detection schemes and followed by the review of previous works.

2.8.2 Non-Coherent Detection

Non-coherent detection scheme does not require CSI and can be categorized into two approaches that involve FSK modulation and differential modulation. By employing FSK modulation at the source, information is sent via multiple carrier frequencies. At the destination, a simple formulation of square-law detector was pointed out in [1] for the single-hop point-to-point communication to recover the transmitted data.

However, when FSK is considered in the cooperative diversity networks involving AF relaying protocol, the non-coherent based ML detector becomes very complicated and complex due to its integral calculation. For that reason, numerous low-complexity NCFSK detectors have been carried out in [10, 17, 18, 47] to eliminate the instantaneous channel estimation. Nevertheless, owing to the multiple carrier frequencies practice, these detectors require significant numbers of spectral resource – resulting in the significant performance loss in comparison with the non-coherent ML detector.

In [29] an optimal non-coherent detector for FSK modulated and On-Off Keying (OOK) transmitted information is proposed to enhance the OWRN performance under a fast Rayleigh fading channels. At the detection, ML based rule is employed for the optimal

detector. Yet, the proposed optimal detector is not in closed form and furthermore, it involves the numerical integration operation, which makes the detection computationally complex to simplify. Hence a suboptimal detector is further proposed to reduce the receiver complexity (i.e. the real components which encompasses the interference over the frequency is canceled) at the cost of performance degradation. Due to the higher complexity as well as the expenses of the performance degradation using FSK, differential transmission scheme have been studied extensively because of the less spectral resources consumption than the NCFSK scheme, but importantly to bypass the need of instantaneous CSI.

2.8.3 Differential Transmission to Bypass Channel Estimation

DPSK has attracted much attention due to the merit of omitting channel estimation at the destination. The modulation scheme depending on the consecutive phase responses of the channel which in turn, does not need accurate phase reference of the signal received. Figure 2.9 depicts the block diagram of a single differential modulation and demodulation [50]. Firstly, the information signal in bits form is mapped into symbols $x[n]$ using PSK modulation. Then the symbols are differentially modulated. As depicted in Figure 2.12, the differentially modulated symbol $y[n]$ at time n , can be obtained as $x[n] \times y[n-1]$, whereby $y[n-1]$ represents the delayed form of the differentially modulated symbol at time $n-1$. At the destination, the received signal is represented by $r[n]$. It is assumed that $h[n]$ denotes the source-destination channel while $e[n]$ represent to the AWGN of the network. $r[n]$ can be obtained as $h[n]y[n] + e[n]$.

By using the complex conjugate² for differential detection process, the received symbols attained can be written as $r[n] \times r[n-1]^*$ where $[\]^*$ signifies the complex conjugate operation. Finally the decoder computes and searches for the closest constellation of the received symbols.

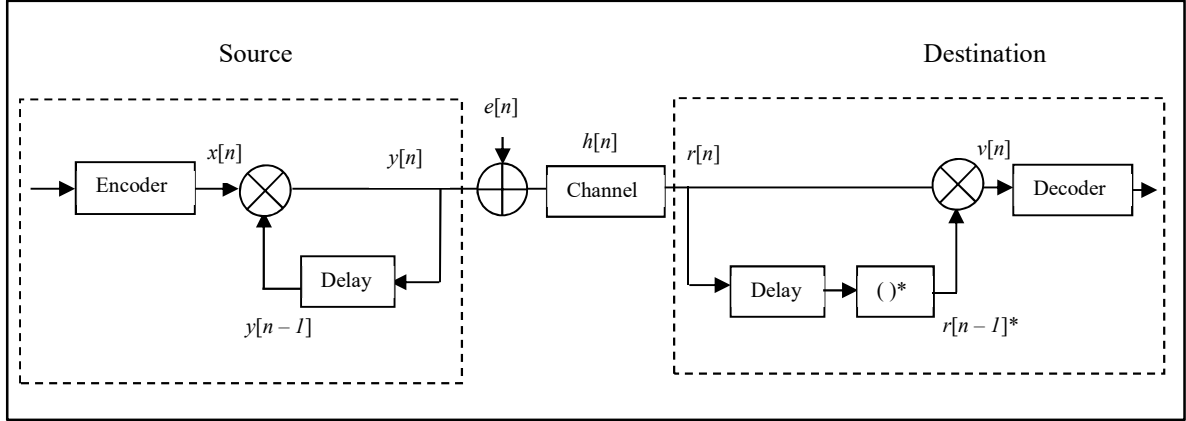


Figure 2.12: A block diagram of a single differential encoder and decoder [50]

Contrary to coherent detection that provide reference signal for phase relative comparison, the differential decoder obtains the phase changes by referring to the previous and currently detected signal. Differentially detected PSK symbol is more attractive in comparison to PSK due to the fact that exact phase reference estimation is not required during the transmission process. On the contrary, the scheme yields irreducible errors because of the difference between the noisy signals at the destination. This noisy signals is caused by the rapidly changes of amplitude over a duration of two continuous time intervals [15]. Albeit to the expenses of the poor performance, differential detection is still preferred

² The complex conjugate consists of real and imaginary value whereby the real value represents for a given complex number while the imaginary value represents the magnitude in opposite sign.

in wireless cooperative diversity system wherein the channel estimation faces challenges in the fading channels.

The related works on differential transmission scheme were studied in [12, 15, 16, 49, 50]. In [12, 16, 49] a two-user cooperation systems using differential scheme of AF protocol has been considered. At the destination, the directly transmitted signal and relayed signal are combined in accordance to the MRC concept, which needs the statistic knowledge of second-order for all transmission channels. A simple error rate performance, BER is derived in regard to Moment Generating Function (MGF) method [12]. This formula is derived with the aim to optimize power allocation among nodes for further system performance improvement. Although the formulation is simple, yet the power allocation optimization scheme is achieved after an exhaustive numerical search. Thus, Zhao *et al.* in [16] derived the PDF of the SNR as well as average BER in closed-form expression for the proposed scheme deploying two differential modulation: DAF and Differential Decode-and-Forward (DDF). However, this scheme requires average gains of the overall transmission links.

In [51], the users that acted as DF aided-relay implemented differential decoding and encoding process of the received partner's information. Before the information is forwarded to the destination, the encoded messages undertake a negative and conjugate operation based on Alamouti's scheme [38]. According to Alamouti's matrix scheme, the directly transmitted symbols and relayed transmitted symbols are devised in the form of matrix. It is stated that the first-row matrix components represent the directly transmitted symbols whilst the second-row matrix components represent the relayed symbols. Since the channel knowledge is unknown at the destination, Cyclic Redundancy Check (CRC) bits are included in both users before the process of encoding. Each user then act as relay to perform CRC check after

receiving the encoded source signal and retransmit the correct re-encoded signal towards the desired destination. However, the received signals at the destination are differentially decoded with the addition of noise term, symbol error and delay which degrades the system performance although it can achieve second-order diversity efficiently. The scheme has an acceptable performance if the noise is small. A Linear Combiner (LC) employing channel gains as well as the energy of the transmission signals are applied at the destination. But at the same time, the combined signals may experience transmission delay and noise due to the possibility of obtaining symbol errors at the destination. In [52], the PDF from the instantaneous SNR is expressed from the performance analysis in terms of average BER, diversity order as well as outage probability under Rayleigh fading channels. The signal received is decoded with the differential LC employing the average SNR instead of instantaneous SNR. The detection and combining processes follow an ML approach for DAF and Piecewise-Linear (PL) approach for DDF. Although the DAF and DDF analytical results outperform the traditional non-cooperative techniques and achieve cooperative diversity, the PL performed worse than the nonlinear ML detectors.

The existing differential cooperative schemes (i.e. single relay and two-user cooperative networks) can be expanded to multiple terminal networks for example, the ad hoc networks and WSN. In [53], differential schemes with DF and AF cooperative networks were investigated for multi-cooperative relays cooperation systems by assuming that the channel variances are obtainable at the destination. The combined signals are differentially decoded based on the detection rule in [54]. For DDF scheme, the relay only retransmit the correctly decoded symbol to the destination. A decision based on threshold is applied to obtain for the relayed signals to be combined with the directly transmitted signal for joint decoding purposes. The results showed that the scheme achieves diversity orders as a

number of cooperative relay N increases. However, the DAF and DDF schemes yield performance gap as compared to the coherent transmission scheme. It is due to the needs of additional SNR as compared to the coherent transmission scheme to realize the same average BER [23]. A highly effective approach to mitigating the SNR loss is referred to as MSD [40–42].

Generally without the channel knowledge, the traditional single differential detector scheme is subjected to a 3dB loss in contrast to the coherent detection. Additionally, irreducible error floor occurred when the Doppler frequency increases [4]. So as to solve the possible performance degradation at high Doppler frequencies, Multiple-Symbol Differential Decoding (MSDD) was proposed for non-coherent schemes in [53, 54]. For decoding process, one of the well-known decoder employed by previous studies is the ML decoder. However, the optimal decoder experiences computational complexity due to the exhaustive search over all the possible received symbols, especially when the symbols length increases. An effective algorithm, SD is suggested to mitigate the computational complexity whereby the basic idea of the algorithm is to search the nearest lattice point to the actual received signal within a hypersphere radius R .

In a two-dimension figure illustrated in Figure 2.12 each symbol is represented by a lattice point in a lattice field. The search area can be easily concentrated within a circle (hypersphere) surround the received vector to search only a lattice point (represents a hypothesis). Thus, at the same time disregard the search of all the points beyond the circle. Due to its ability to reduce the exhaustive search over the lattice point, the application of SD was proposed in [30, 55] for point-to-point wireless communication to decrease the MSD computational complexity when the detection block length increased.

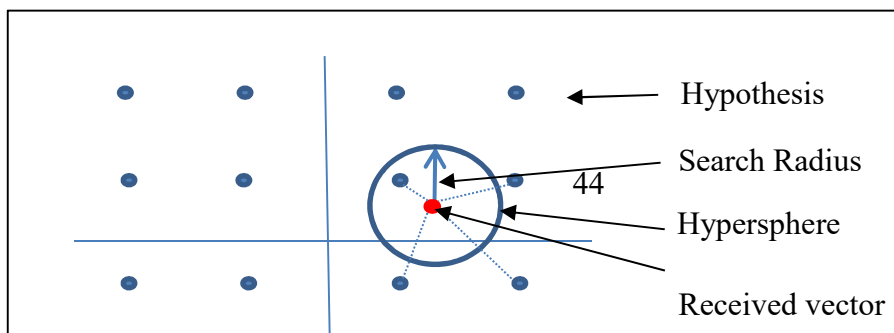


Figure 2.13: Geometrical representation of the SD algorithm

2.8.4 Detection Schemes in Highly Mobile Environment

Although cooperative diversity systems can achieve diversity gain, the systems are susceptible to frequency offsets due to Doppler effects caused by highly mobile nodes. Typically in the real-world scenario such as mobile network and mobile WSNs (i.e. environment monitoring, mining, meteorology, seismic monitoring and inventory tracking) especially when high Doppler frequencies involved, the differential detection may perform poorly [58]. Moreover, due to the nodes rapid movement, the oscillators of all the are facing challenges to synchronize accurately [50]. Primarily, two approaches are carried out with the aim to deal with the effect of frequency offsets. The first approach involves the frequency offsets estimation and compensation whereas the second approach focuses on alternatives to design a powerful detector that eliminates the effect of frequency offsets. Table 2.1 tabulated the current research that employed frequency offsets estimation.

Differential modulation for a half-duplex, forward one-way relaying communication and a two-way relaying in the presence of frequency offsets have been studied in [27, 33, 57] employing the frequency offsets estimator as well as DD based modulator that mitigates the frequency offsets effect. In [35] the author proposed a symbol based and multiple symbol based detection to regain the transmitted information upon the removal of the self-

information from the received information. On the other hand, in [27] the transmitted pass-band signals are required to be converted to baseband to perform carrier frequency synchronization.

It is demonstrated that the proposed estimator in [27, 33] requires additional pilot symbols prior to the transmission process. Due to the additional pilot symbols, complicated frequency offset tracking circuitry and acquisition is congruently needed whereby complex hardware implementation is acquired. On top of that, the adding of extra symbols decreases the data rate of the transmitted symbols [21].

The second approach is DD modulation. The modulation is able to effectively eliminate the frequency offsets under fast fading. The DD modulation is used when both channel knowledge h and frequency offsets w are unknown, while single differential modulation consider that only channel knowledge is unknown whereas frequency offsets is equal to 0 [22]. Thus, it is preferable to design DD based detector due to its insensitivity to frequency offsets especially in a fast fading environment. Table 2.1 shows the contributions of detection employing frequency offset estimation while Table 2.2 demonstrates the contribution of DD detection that eliminate frequency offset estimation throughout the previous years.

Table 2.1: Contribution of detection scheme with frequency offset estimation

Year	Author(s)	Contribution
2011	Mehrpouyan et. al. [60]	Proposed algorithm to mitigate frequency offsets effects utilizing extra training sequences assuming perfect time synchronization.

Table 2.1 continued

2012	Liu and Zhu [61]	Presented the joint ML frequency offsets and channel estimation of asynchronous cooperative communication systems. A low-complexity approximate SAGE-IE (A-SAGE-IE) is proposed. The proposed algorithms formed numerous one-dimensional optimization problems from the multi-dimensional optimization problem whereby the channel coefficients and frequency offsets of each relay-destination channel can be estimated independently.
2013	Liu et. al. [62]	A simple yet effective estimation method based on the devised training signals is presented to achieve synchronization. An iterative algorithm is also derived to enhance the performance associated with the synchronization impairments estimation. The proposed algorithm converts the difficult multiple parameter estimation problems into more tractable sub-problems of estimating numerous independent impairments pairs for the relays. Simulations indicate that the proposed estimator can asymptotically achieve the Mean Square Error (MSE) for the perfect frequency synchronized case.

Table 2.2: Contribution of detection scheme that eliminate frequency offsets estimation

Year	Author(s)	Contribution
2008	Yi et. al. [59]	Presented a DD transmission scheme in the DF relayed networks. The average SER performance is investigated for the slow Rayleigh-fading channel that successfully compensates the SNR loss exhibited in the doubly differential transmission.
2011	Rabiei & Beaulieu [23]	Proposed DDPSK and MSDD detection scheme for point-to-point communication that is insensitive to constellation rotation due to frequency offsets in the case when the frequency offsets are not large.
2013	Bhatnagar, Manav & Olav [63]	DD modulation and PL decoder are suggested in the DF protocol cooperative communication. The simulation results show that the DD modulation was superior than the non-coherent FSK at high SNR regime regardless of frequency offsets. However, the proposed scheme requires the average SNR knowledge for source-relay channel at the destination.
2014	Wu et. al. [35]	Proposed differential modulation schemes (i.e. the single differential and DD) for AF two-way relaying networks. Additionally, multiple symbol based signal detection is suggested to enhance the differential modulation schemes performance. Due to the lack of the CSI, the frequency offset is estimated and compensated.

Table 2.2 continued

2014	Gao et. al. [27]	Proposed a DD scheme for two-way AF relaying scheme under Nakagami-m fading channels with overhead aided. The average Symbol Error Rate (SER) performance is analyzed and the designed scheme works well with the existence of frequency offsets.
2013	Nasir et. al. [64]	A MMSE receiver is proposed for decoding process at the destination whereby the frequency offsets were compensated. A low complexity MMSE is also presented to minimize the overhead in regards with synchronization in the networks.
2015	Cano et. al. [65]	A simple distributed DD TDM scheme employing diagonal ST based on unitary codes is formed to eliminate multiple frequency offsets among the relays under Orthogonal Frequency Division Multiplexing (OFDM) transmissions. Specific modulations are designed at the relays considering multipath effects.

A general block diagram of DD encoder and decoder for a single link wireless communication network is depicted in Figure 2.14 with D representing delay and $x[n]$ is the encoded source signal at time instant n .

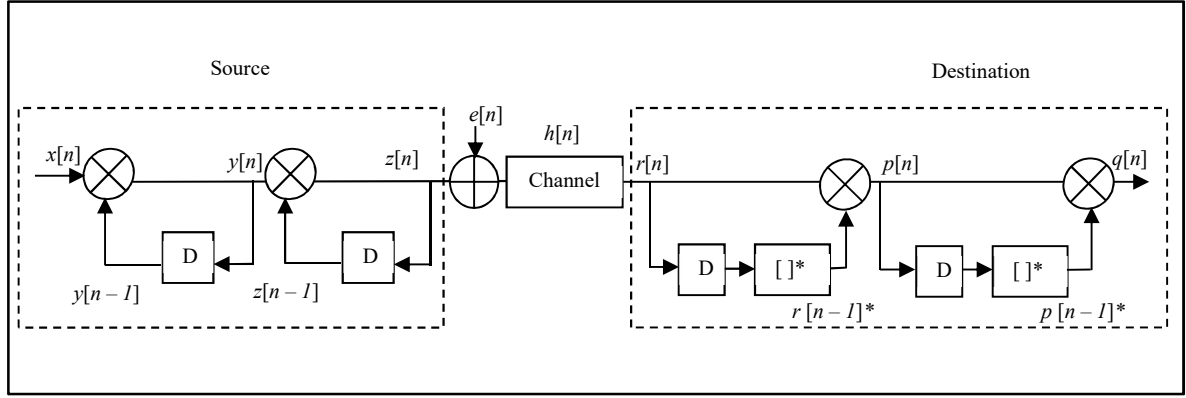


Figure 2.14: Block diagram of DD encoder and decoder for single link wireless communication network [66]

After DD demodulation process, the transmitted signal $z[n]$ can be expressed as $x[n] \times y[n-1] \times z[n-1]$, whereby an extra delay stage is added in the conventional single differential modulation. By assuming a point-to-point wireless transmission in the existence of frequency offsets, the signal received at the destination is described by $r[n] = he^{w[n]}r[n] + e[n]$. From the expression h is the channel gain, w is the frequency offset represented by $2\pi f$ and $e[n]$ is the AWGN. $q[n]$ represents the decision variable and is formulated as $p[n]p[n-1]^*$, where $p[n] = r[n]r[n-1]^*$. Despite the achievement for diversity gain, the systems are sensitive to frequency offsets. Stoica *et al.* in [22] proposed the DD demodulation in regards to the simple heuristic detector favoured over the complex ML decoder. It is shown that the result of the heuristic detector matches well with the ML approach.

Bhatnagar *et al.* in [48, 65] further implemented and analyzed the DD modulation and demodulation for the cooperative communication. The study utilized AF and DF relaying protocol with the effect of random frequency offsets over a flat fading Nakagami-m channels as depicted in Figure 2.15.

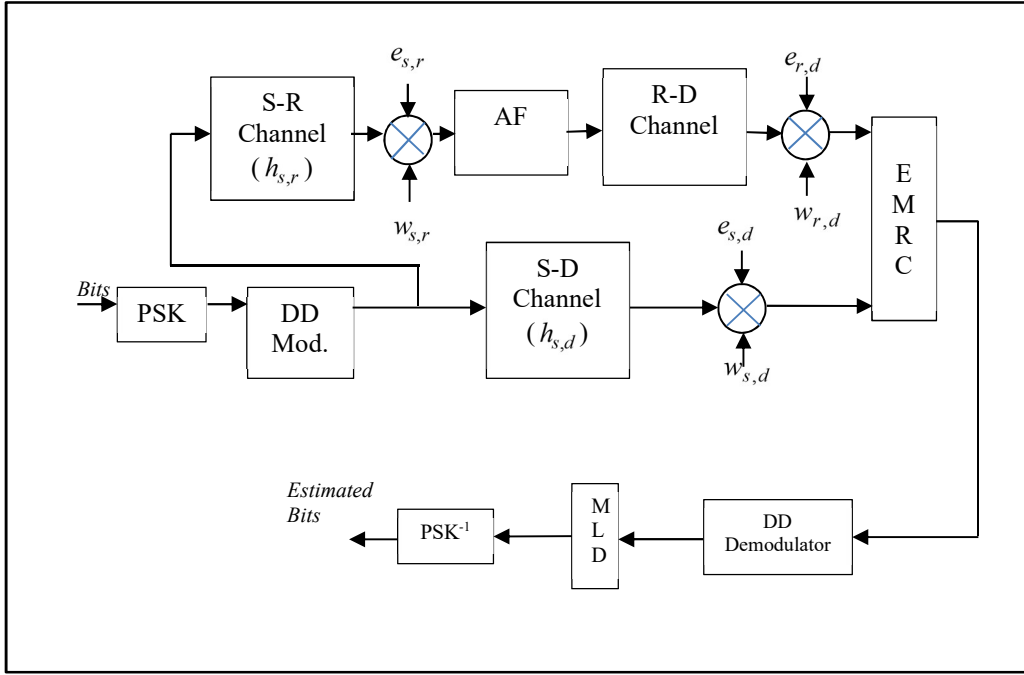


Figure 2.15: Block diagram of the DD transmission [66]

The signal was first encoded using double-differential modulation before being transmitted towards the relay and intended destination respectively. At the destination, the decision can be made by employing the MRC [40] based scheme. Since the channel knowledge and frequency offsets were unknown, the Emulated Maximum Ratio Combining (EMRC) scheme exploited the channel variances to replace the channel coefficients. However, the BER analysis showed that the EMRC produced poorer result as compare to the ideal MRC in [40], therefore instantaneous SNR was suggested which perform near to the MRC. The proposed scheme outperformed the conventional differential detection in certain scenarios.

DD modulation was further proposed by Cano *et al.* in [24] at the relays instead of the destination. The block diagram of proposed modulation and simple heuristic detector is depicted in Figure 2.16 for a distributive network.

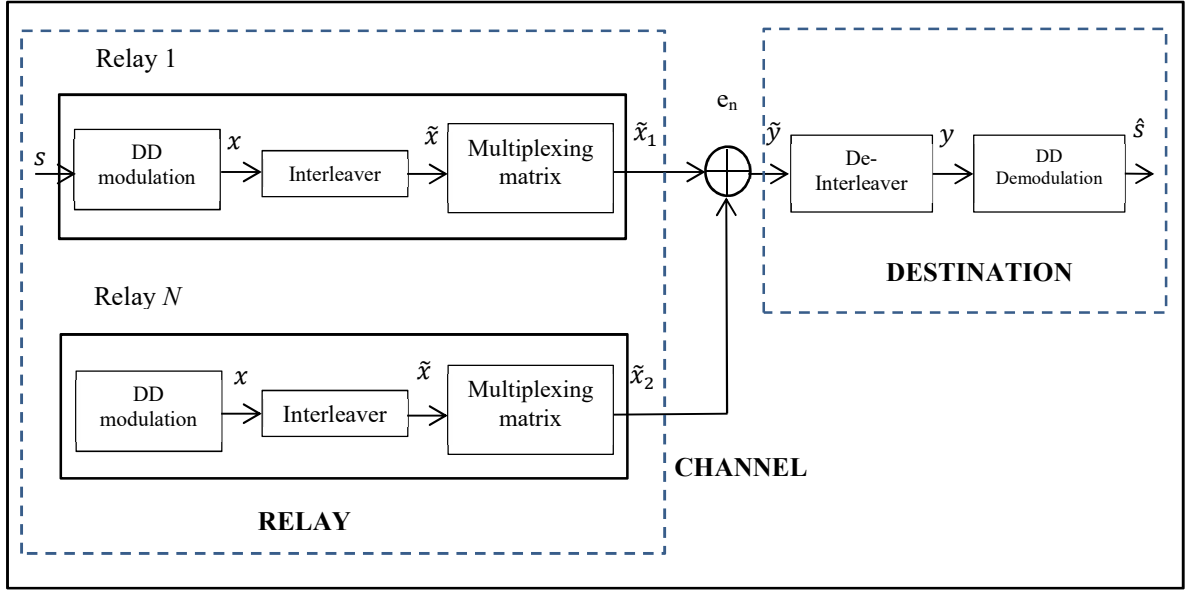


Figure 2.16: Proposed DD transmission system model [24]

Refer to Figure 2.16, a block of $K - 2$ information symbols s is received by the relays. Then the component of s is converted to a unitary based diagonal matrix v_k . In order to acquire the double-differentially modulated block of information symbols, an approach in regards of recursions is applied to v_k . The detected symbols x_k are then being fed into the interleaver block to form $\tilde{x}$ and analysed using a multiplexing matrix.

Thus, the heuristic detector at the destination can be expressed as:

$$v_k = \arg \min_{v \in \mathcal{E}} \left\| \sum_k^{-1/2} (y_k y_{k-1}^* - v y_{k-1} y_{k-2}^*) \right\|^2 \quad (2.16)$$

However, the proposed scheme requires the prior source information and transmission ordering protocol³ so that the source recognizes the cooperating relay. Due to that reason,

³Transmission order protocol determines the order for which terminals that acts as online transmitter with the source terminal representing the first node and the destination as the ending terminal in the $N+1$ network nodes order [111].

the proposed scheme is ineffective particularly in a wireless ad hoc network as each communicating terminal may connect or terminate from the network at any time [24].

Although the DD approach enhances the wireless network's performance in the presence of frequency offsets, higher SNR power ratio is desired in order to attain similar average BER performance of the coherent detection. With the intention of dealing with the SNR loss, MSDD becomes an attractive detection mechanism [35].

It is observed that most of the previous existing non-coherent and differential transmission schemes eliminate either the instantaneous CSI or the statistical knowledge of the channels and noises (i.e. the channel statistics, channel distribution and noise variances). Specifically, the Direct Combiner (DC) combines weights corresponding to the second-order channel statistics. But, the statistical channel/noise information may vary with time under time-varying channels and noise environments. Thus, it is a challenging task to estimate the statistical knowledge under such environment. In the literature, the presence of frequency offsets in the time-varying scenarios have not been considered for M -DPSK.

2.8.5 Multiple Symbol Differential Detection

Multiple-symbol based differential detection has been proposed in [53, 54] to overcome the limitation of symbol-based differential detection. MSDD scheme is referred to the implementation of numerous consecutive symbols (i.e. more than two symbols) at the destination for detection purposes.

(a) Multiple Symbol Detection Based Differential Scheme for Fast Fading Environment

Table 2.3: Major contributions of MSD scheme under fast fading channel

Year	Author(s)	Contribution
1989	Divsalar and Simon [55]	The pioneer work on non-coherent detection with DPSK signal over AWGN channel was presented. The research derived and studied the optimal detection algorithm based on ML sequence estimation and DPSK signalling. N data decisions is applied to solve the poor performance experienced in the system due to DPSK signalling.
1989	Wilson et. al. [68]	Proposed two empirical sub-optimal resolutions using MSDD with far less computational complexity. The scheme offered trade-off between complexity and performance.
1990	Makrakis and Feher [69]	A multiple-symbol interval aided non-coherent sequential receiver was proposed. The theoretical results demonstrated that the scheme had performance that was near approaching to the performance of coherently detected DPSK over fixed and stationary channels.
1994	Divsalar and Simon [70]	Extended the MSD in [55] to trellis based coding employing amplitude modulation under Rician as well as Rayleigh fading channels.

Table 2.3 continued

1994	Mackenthun [71]	Presented a simple and fast algorithm of MSD algorithm. However, the scheme only accommodate for AWGN channel as
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		well as quasi-static Rayleigh fading channel that is time-invariant over the decision period.
1997	Peleg and Shamai [72]	The concept of MSD was further enhanced for the implementation of soft input and soft output system. An iterative detection was proposed for convolutionally-coded, interleaved and DPSK encoded system.
2005	Lampe et. al. [30]	Presented a sub-optimal yet lower complexity version of MSD. In this new method, the ML estimation search space was reduced to a certain radius of a circle.
2006	Pauli et. al. [73]	The idea of MSDSD pertaining to MAP approach was further broaden to allocate iteration process with some other soft-in-soft-out detector. The Extrinsic Information Transfer (EXIT) characteristic of the proposed SISO-MSDSD module was thoroughly analyzed. Through the EXIT function analysis, maximum coding gain was significantly achieved using the proposed best code design.

During the past years much more research has become available pertaining to the previous research tabulated in Table 2.1, [17] studied on a generalized differential modulation employing AF relaying protocol wireless relay networks to mitigate the gap difference between the differential modulation and coherent detection. In the proposed scheme, the sequence of information bit are firstly distributed into multiple blocks. Then, the reference symbol of the block was encoded differentially. The rest of the symbols within the block were encoded differentially pertaining to the reference symbol of the similar block.

After the transmission process, the signals from the source as well as the relays were combined as well as differentially detected employing suboptimal combining weights in [53] without channel knowledge. It was demonstrated that the simulation results showed better improvement when the transmission block length increases under slow Rayleigh fading channels, with the assumption that the channels stayed stationary for a block period.

On top of that, in order to resolve the SNR loss in the differential transmission, multiple-symbol differential detection based on ML criterion implementing M-ary DPSK transmitted over an AWGN channel has been introduced in [55]. Unlike the symbol-based detection scheme, this technique can be achieved by increasing the number of symbols in regard to an observation interval to form a block of N consecutive PSK symbols for joint detection of $N-1$ consecutively transmitted data symbols process. It was revealed that the error performance significantly enhanced as the observation window size N increased. Nevertheless, the detection complexity increased exponentially as the number of transmitted symbols grew. Numerous research had attempted to suggested sub-optimal algorithms in [72–74] to solve the MSDD detection complexity pertaining to the subset search approach. Firstly, a reference of block sequence was implemented to determine the maximum candidate list using two consecutive symbols. The candidates was further searched and analysed with the increased window observation size and showed maximization for the MSDD metric. However, it has conclusively been shown that the suboptimal detectors perform only the complexity analysis reduction at the expense of detection decision.

One simpler alternative of MSD detector was suggested in [30] by utilizing the SD concept. The idea was widely modelled in CDMA network, MIMO system and also implemented for the wireless cooperative communication system. In order to mitigate the performance degradation due to the time-selective environment, frame-based transmission

was carried out as opposed to the symbol-based. The idea of the MSDSD can be summarized as follows:

- (a) A decision metric of the multiple-symbol based ML decoding was determined by employing the Cholesky factorization⁴ of the channel statistics.
- (b) Then, the decision metric is adopted with the SD approach. By referring to the idea of SD, the vector search for the ML estimation is limited to a spherical region within certain radius R , centered at the received vector.
- (c) By applying the Schnorr-Euchner search strategy [26], the radius R is dynamically updated when a candidate vector is found and the process was repeated with new radius R .
- (d) The MSDSD process ended when no vector was found within the sphere of radius R . Due to the benefits of the decision extension of the observation window length.

MSDSD has been extensively studied in many other works. Wang and Hanzo in [77] proposed MSDSD in the DAF cooperative system. The research posited that the MSDSD detection provide significant enhancement in contrast to the traditional differential scheme to regain the broadcasted signals. But, most of the existing differential transmission works found in the literature assumed perfect synchronization at the destination. In cooperative diversity whereby all the terminals may be in mobility state, the synchronization between the terminals' oscillators will never be perfectly matched because of the frequency offsets effects. The performance of DPSK and multi-symbol transmission scheme deteriorates in fast varying environment that violate the assumption of static channel which resulting in the

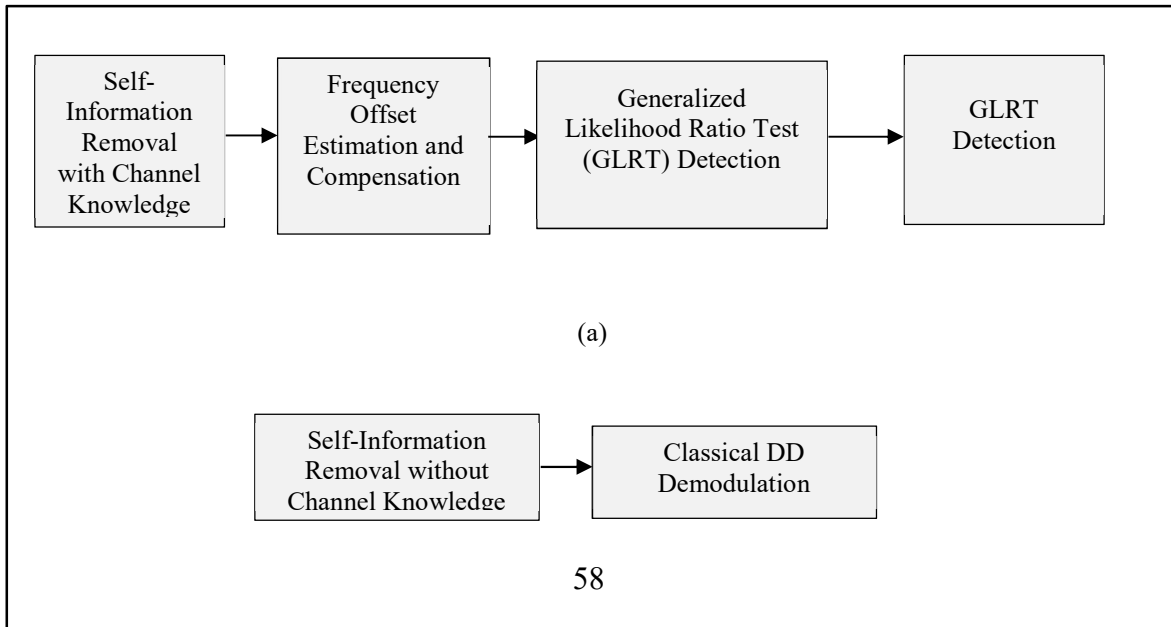
⁴ The Cholesky factorization method factorizes a matrix into a product of a lower triangular matrix with its conjugate to solve numerical problems.

error floor. Thus, DD approach is proposed with the consideration for frequency offsets effect without the knowledge of channel knowledge. However, the DD modulation performs slight degradation than the coherent PSK as differential approach requires higher SNR to obtain the similar average BER of coherent detection. Thus, an attractive approach to alleviate this SNR loss is through the MSD based detection in the presence of frequency offsets.

(b) Multiple Symbol Detection with Double Differential Scheme

MSDD detection becomes an attractive scheme owing to its capability in recovering the desired information symbols without the needs of high SNR in order to obtain the similar average BER performance of the corresponding coherent detection, in contrast to DD modulation [33, 76].

MSDD detection is suggested in [78] and has been further studied in [79] for point-to-point wireless communication and for cooperative communication system in [35]. Wu *et al.* [35] investigated differential detection based on single and double modulation in the AF relaying protocol TWRN flat-fading channel with the effect of frequency offsets. In single differential detection, an approach of four-stage block as depicted in Figure 2.17(a) was executed.



(b)

Figure 2.17: (a) Symbol-By-Symbol based Single Differential Detection (b) Multiple-Symbol Double Differential Detection

Figure 2.17 illustrated the single differential detection, whereby the channel knowledge (i.e. channel gains and amplification factor) estimation was performed upon receiving the broadcasted superimposed information from the relay. Simultaneously the self-information of the source due to the effect of the TWRN, were removed. In order to estimate the frequency offset, only two pilot symbols are transmitted at each of the information block. Then, the compensation of frequency offset was carried out following the frequency offset estimation. Finally, Generalized Likelihood Ratio Test (GLRT)-based algorithm⁵ were utilized to detect the transmitted symbols.

On the other hand, a two-stage mechanism was performed as shown in Figure 2.17(b) based on the DD concept. In the first stage, a single differential detection cum self information elimination was design without the knowledge of the channel. Conventional DD detection was utilized during the second stage due to its attractive characteristics that eliminate not only the highly-complex computationally frequency offset estimation, but also the compensation and tracking circuitry which were required by the single differential detection. The design of MSDD detection can be determined following the concept of single

⁵In a GLRT detection, the likelihood function is estimated or maximized over an unknown parameter (i.e. the channel gains) in order to obtain the decision metrics [112].

differential. Thus, the received signal at either source nodes S_1 or S_2 without additional noise can be expressed as:

$$\hat{y}_1(t) = \bar{y}_1(t)\bar{y}_1^*(t-1) = |v|^2 e^{j(w_2-w_1)} p_2(t-1)z_2(t) = \hat{h}p_2(t-1)z_2(t) \quad (2.17)$$

whereby, $\bar{y}_1$ represented the received signal after the process of compensation with $(*)$ representing the conjugate operation; v is the knowledge of the channel; w_1 and w_2 represents the unknown carrier offsets; p_2 is the power transmitted by the relay and $z_2(t)$ indicates the recovered symbols. Following the iteration process from (2.17), a new expression based on DD detection can be obtained as:

$$\hat{y}_1(t-n) = \hat{h}p_2(t-N+2) \prod_{m=0}^{N-n-3} z_2(t-n-m), n = 0, 1, 2, \dots, N-2 \quad (2.18)$$

with N representing the observation symbol length. The decision rule at the destination can be attained as (2.19) by applying minimum Least Square (LS) over the minimization metric of $\hat{h}p_2^*(t-N+2)$.

$$\max_{z_2(t), z_2(t-1), \dots, z_2(t-N+3)} \left| \sum_{n=0}^{N-2} y_1(t-n) \dots \cdot \prod_{m=0}^{N-n-3} z_2^*(t-n-m) \right|^2 \quad (2.19)$$

Nevertheless, (2.19) yielded an extremely high complexity in the computational process of $(MN - 3M - N)M^{N-1}/(M-1)^2$. Thus, in order to minimize the complexity on the order of $N \log_2 N$ [55], a fast algorithm was implemented.

The BER performance of the proposed schemes were analysed. It was demonstrated that the single differential detection implementing the multiple-symbol based transmission degraded the system performance in despite of the increment of the observation length N . This happened due to the limited number of training symbols transmitted which resulted in an inaccurate estimation of the frequency offset. Therefore, it was suggested that the single symbol transmission should be adopted for the single differential detection in the presence of frequency offset. At the same time, the data rate of the proposed estimator experience

decrement owing to the additional pilot symbols requirement. The estimator became more complex due to the requirement of frequency offset acquisition and tracking circuitry which acquired complex hardware implementation. Unlike the single symbol based detection, multiple symbol detection appeared to have the capability to alleviate the loss of SNR resulting from the DD modulation and improve the system performance significantly when N increased. However, this scheme performed at the expense of a complex computational process. The summary of the reviewed receiver designs in [24, 27, 33, 64] is tabulated in Table 2.4.

Table 2.4: Summary of the reviewed receiver designs

Article	Network	Fading	Relay Protocol	Direct Link	Channel Estimation	Frequency Offset Estimation	Blind/ Pilot
Bhatnagar et. al. [66]	OWRN	Nakagami- m	AF	Yes	Instantaneous SNR is required	No	Partial Pilot
Cano et. al. [24]	Distributed OWRN	Discrete- Time Low- Pass	-	No	No	No	Pilot
Gao et. al. [27]	TWRN	Nakagami- m	AF	No	Yes (Self)	Requires frequency offset estimation and compensation	Pilot (Transmission Overhead)

Table 2.4: continued

Wu et. al. [35]	TWRN	Frequency Flat Rayleigh	AF	No	Yes (SD) No (DD)	Requires frequency offset estimation and compensation	Both
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2.9 Cooperative Differential Distributed Space-Time Coding Transmission

STC can be categorized into two: STTC and STBC as described in Section 2.5.1.2. Due to the attractiveness of the block coded space-time, the scheme has been further implemented and studied in the distributed environment.

DSTBC scheme [1] has sparked increasing interest because of its outstanding capability to provide both significant diversity gains and capacity as well as to combat fading effects. By exploiting the multiple distributed antennas of multiple users, DSTBC has been developed for a multi-user cooperative communications environment to achieve extra spatial diversity. Additionally, in order to achieve the cooperative diversity, multiple users construct and transmit the space-time codewords via statistically independent fading channels. Basically, the distributed space-time coding offers the following three major advantages.

- (i) The DSTBC possess all features of the original STBC and has the potential to achieve significant performance gain in a multipath fading channel.
- (ii) The mobile units in a DSTBC network only need to be installed with a single antenna. With this feature, the mobile phone could offer more compact size for future wireless mobile network. More importantly, a simpler structure would result in an effective power resource to prolong the battery life span.
- (iii) Thirdly, the cooperating users are able to assist the source(s) to forward and relay the signals towards the destination with the help of DSTBC. Through this approach, the wireless network coverage could be expanded without any investment of additional infrastructure.

There has been some works investigating the DSTBC using different cooperative transmission protocols. A Differential Space-Time Modulation (DSTM) scheme was

designed by Hughes in [80] based on the unitary approach under fast time-varying channels. However, a major concern was raised by previous research subjected to the 3 dB performance penalty experienced as compared to the coherent detection scheme. Thus, significant interest has been developed with the intention to minimize the extra performance degradation. Inspired by Tarokh and Jafarkhani's concept of DSTBC, a differential detection has been suggested in [81] whereby a redundancy encoder has been executed to produce extra coding gain. The proposed DSTBC scheme was modelled according to the first-order discrete of Markov process for two transmit antennas. Trellis diagram was incorporated into the proposed DSTBC whereby the decoder employed the Viterbi algorithm where the proposed scheme demonstrated an approximately 1 dB performance enhancement over the traditional DSTBC. The DSTBC scheme had been further explored with a well-known MSDD by Gao et. al. in [82] for MIMO system. The researchers proposed and analysed the DSTBC sequence estimation based on the ML approach and the proposed scheme enhanced the traditional DSTBC performance up to 1.5 dB with the expenses of exponential increase in computational complexity.

For cooperative communication system, a two-stage transmission model for AF relay networks pertaining to DSTBCs is proposed in [83]. In [84], DSTBCs achieving full cooperative diversity were constructed and studied by implementing the division algebra concept. In [85], DSTBCs based on both AF and DF relay protocols were investigated for multiple-relay flat-fading channels. However, the results of these schemes demonstrated that there were trade-off between diversity and rate of data transmission.

Similarly for STC transmission, DSTBCs were proposed individually in [84–87]. In [87], the authors proposed DSTBCs using the unitary matrices. In [88], full diversity DDSTBC was constructed based on circulant matrices applying various differential codes

such as, symplectic group $Sp(2)$ code, square real orthogonal codes and Alamouti code. The proposed schemes were applied in relay networks with a certain number of relays. On the other hand, DSTBC schemes were suggested for any number of relay nodes in [84, 87]. The proposed schemes can achieve full diversity and higher data transmission rate but required a higher decoding delay and at the expense of higher computational complexity in the encoding and decoding process particularly in the scenario of more than four relay nodes. It is important to note that the aforementioned DSTBC schemes lacked in either the ability to gain full cooperative diversity in terms of full data rate simultaneously or the computational complexity is increased, which promote the scheme to be infeasible or even non cost-effective. In fact, Alamouti code was the only low complexity DSTBC scheme which has the ability to perform full data rate transmission and full cooperative diversity with very simple symbol-based ML decoder [90]. However, this code was designed only for the implementation of two single-antenna nodes. Due to that reason, the diversity order was limited to two only and this code may not be able to achieve any array gain.

All the above mentioned schemes assumed the communication channels were flat fading. Furthermore, most of the existing studies made the assumption that perfect channel knowledge was known at all the transmission links for a flat fading and quasi-static channels. The channel knowledge is likely to be acquired using the channel estimation via pilot signals. However, due to the extra transmission of the training sequences or pilot, the overhead increases which, significantly reduces the throughput of the networks. The channel estimation is suitably applicable in quasi-static or slow-varying scenarios where infrequent switching of the pilot signalling occurs. But, it may be a challenging task to obtain perfect channel knowledge when the channel coefficients experiences time-varying environment. In time-varying channel however, the frequency of the pilot switching increases that cause high

computational complexity in the channel estimation. Additionally, the extra pilot symbols increases the complexity in the estimation computation. Thus, non-coherent and differential transmission schemes were proposed to eliminate the need for the extra channel estimation circuitry and frequency acquisition requirement. Although single differential schemes were capable to bypass channel estimation, the schemes fail to perform when frequency offset occurred.

It is well-known that in a wireless mobile communication network, frequency offset is a common and inevitable component due to the imperfect synchronization at the receiver terminals caused by terminals' mobility. In order to mitigate the presence of frequency offsets without knowing the channel knowledge, some studies were analysed in [58, 62, 63]. In [60], iterative algorithms for multiple frequency offset estimation was proposed. The proposed algorithm reduced the estimation complexity without sacrificing the bandwidth performance. However, training sequences were required for the multiple frequency offsets estimation in Space-Division Multiple Access (SDMA) cooperative networks. MMSE detection at the receiver was suggested in [64] for DSTBC AF relaying protocol networks. The results demonstrated that the combination of proposed LS and Differential Evolution (DE) estimator join with the compensator may achieve full diversity in a multi-relay networks. In order to bypass the frequency offsets estimation, DD STBC has been suggested by [63, 89].

Bhatnagar and Hjørungnes in [91] designed distributed DD coding for the implementation of Orthogonal Space-Time Block Coding (OSTBC) over Rayleigh channels in the presence of frequency offsets. The proposed scheme achieved higher performance gain in contrast to the previous distributed DD based on unitary code. Similarly, in another work by [65] a distributed DD TDM scheme with diagonal ST unitary code was proposed to

bypass multiple frequency offset estimation in OSTBC. DD transmission was capable to eliminate both channel knowledge and frequency offset estimation at the cost of reduced frequency offset mitigation range, thus the suggested scheme is preferable for small frequency offset deviation practice. Table 2.5 summarizes the major contributions of differential detection schemes over STC environment.

Table 2.5: Major contributions of differential detection schemes for STC

Year	Author(s)	Contribution
2001	Liu et. al. [20]	Proposed a novel DDSTBC in reference to the diagonal unitary matrix group for the time selective fading channels regardless of the channel and frequency offsets estimation. Performed less effectively in the rapidly fading environments.
2003	Anghel et. al. [92]	Proposed a distributed Alamouti based cooperative system with two relay stations in a slow Rayleigh flat fading channels.

Table 2.5 continued

2006	Jing & Hassibi [83]	Applied the concept of STC for AF wireless relay networks under Rayleigh fading channels employing linear dispersion code and Alamouti code at the relay. Considering the relays do not know the channel information yet perfect synchronization was
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		assumed at the destination. The results showed that diversity can be realized with very high transmitted power.
2006	Kiran & Rajan [84]	Proposed a distributed STC design based on commuting sets and doubling construction that reduced the decoding complexity. In addition two group ML decodable DSTBCs were constructed using division algebra with the aim to reduced the decoding complexity.
2007	Cano et. al. [24]	Proposed a distributed DD modulation across orthogonal channels using TDM as well as simultaneous transmission regardless of channel and frequency offsets estimation. Simulation results show that diversity gain is achieved with coding gain losses that increase with number of relays.
2010	Bhatnagar & Hjørungnes [93]	Proposed a DD coding scheme which is applicable for square OSTBC using M -PSK constellations. The results showed that the proposed schemes were capable to decode the space-time data without channel coefficients and the frequency offset knowledge yet achieved high coding gain.

Table 2.5 continued

2013	Nasir et. al. [64]	Proposed a MMSE detection at the receiver in the presence of multiple frequency offsets and unknown channel knowledge.
2014	Cano et. al. [65]	Proposed distributed DD TDM using diagonal space-time unitary codes in order to eliminate multiple frequency offsets and gain

		space diversity. Simple and robust non-coherent selective as well as adaptive transmission were also studied and analysed demonstrating reduced decoding design complexity whereby full space and multipath diversity can be achieved.
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2.10 Conclusion of Analysis

In this section, conclusion of the critical analysis on works by other researchers is included. Brief overview for different type of diversity combining techniques and differential detection in repetition-based and STBC signalling schemes are revisited, respectively. Desirable design and characteristics of each schemes are analyzed extensively. The simplest differential transmission cooperative network involving two transmit antennas is discussed in depth. Previous research showed that statistical knowledge and average gain for the transmission channels are required. It is also revealed that most researchers consider slow fading channel when the frequency offsets are not large. But for practical environment, it is challenging to obtain the channel knowledge perfectly, especially in high mobility scenarios. Furthermore, the conventional differential transmission performs poorly in the system that is susceptible to frequency offset caused by mobile nodes. Thus, DD encoding and decoding is further studied. The DD is originated from the basic differential transmission eliminating the channel and frequency offset estimation components, whereby it is more interesting in practice due to the implementation simplicity. However the differential transmission experiences performance loss relative to coherent detection and induced irreducible error floor at high Doppler frequency.

In order to mitigate the poor performance of the differential transmission, instead of symbol-based transmission, MSD is further employed and studied. At the destination,

detection and decoding of many existing research, such as GLRT employing LS, MMSE and ML are often computationally infeasible for scenarios comprising a large number of N . According to previous studies, the cooperative communication performance are further enhanced utilizing the diversity combiners such as LC, MRC and SC. Most of the studies employed LC and MRC which requires the knowledge of all transmission channels. This incurs transmission delay and noise due to the detection error. Thus, it is the interest of this research to enhance the performance of the wireless cooperative communication in a time-varying environment without the channel and frequency offset estimations.

2.11 Summary

This chapter described the characteristics of a single link wireless communication, wireless channel model and the wireless cooperative strategies. In addition, various cooperative relaying protocols were described. Mostly importantly, the background of the research in this thesis is needed and helpful in understanding of the systems and the detection schemes considered in the next chapters. In the next chapter, the detailed insight of non-coherent detection schemes will be studied in order to improve the performance in the wireless cooperative network.

CHAPTER 3

MULTIPLE SYMBOL DOUBLE DIFFERENTIAL SPHERE DECODING

3.1 Introduction

In the previous chapter, the related works subjected to the DD based transmission as well as various types detection schemes are studied and analysed in order to develop an

improved performance of an AF MSDD for cooperative wireless communication with different mobility.

In this chapter, DD detection based on multiple symbol SD for AF relaying protocol is considered. In this research, the instantaneous CSI, statistical CSI, noise variances, signal powers, and relay functionalities are unknown a priori under a time-varying environment. It is also considered that frequency offsets is presented in the wireless cooperation networks due to the high mobility of communication terminals. In order to create practical time-varying scenarios, a channel modelling based on auto regression in time series is devised. DD encoding and decoding employing M -PSK constellation for AF cooperative network are developed under repetition-based to omit channel and frequency offset estimation. Then, multiple symbol based transmission is proposed to mitigate the error floor experienced during the symbol-by-symbol based transmission. Finally, SD algorithm is adopted in the MSDD detection to resolve the complexity of the ML detection search space.

Then, the simulation results and the performance of the numerical analysis for existing DD detection based on symbol-by-symbol transmission and MSDDSD scheme is presented. Error performance improvement under time-varying channels particularly in fast fading scenarios is emphasized. Also, a coherent detection for AF relaying protocol under Rayleigh fading is simulated and served as the benchmark for the proposed DDAF networks. Then, a comprehensive error probability is carried over flat fading Rayleigh environment. Specifically, PEP expressions for the detectors are derived, whereby the derivation are valid for and arbitrary sequence length N and arbitrary modulation order M . The superiorities of the proposed detectors over the state-of-the-art differential detectors are justified via comparisons in terms of BER under different channel fading scenarios. This study shows

that both simulation results and theoretical numerical analysis are in agreement. Finally, power allocation expression is obtained to further optimize proposed scheme performance.

The outline of the chapter is as follows. The first part of this chapter in Section 3.2 describes the system and channel model addressed in this thesis. In Section 3.3, the chapter explains the steps which make up the proposed methodology of various detections - conventional DD and MSD. The next part of the chapter derived the new performance analysis of conventional DD and the proposed MSDDSD. The proposed performance analysis and simulations which were run with MATLAB to assess the validity of the proposed methods is further described. The final part of this chapter discusses the results of the simulations in Section 3.4. Finally, Section 3.5 summarizes the chapter.

3.2 System and Channel Model

A simple cooperative communication system portrayed in Figure 3.1 that comprises of source, relay and destination each. The work philosophy behind this cooperative idea is to enable the source to broadcast signals along the individual wireless path towards its desired destination aided by other terminal(s) that act as relay(s). The transmission phase is furnished into the broadcast and relayed phase.

During the broadcast phase, the source transmits its information directly to the destination. Simultaneously, the relay receives and processes the source data. As the TDM scheme is considered for the data transmission, the source remains idle during the relayed phase while the relay retransmits the processed signals to its intended destination. By employing TDM, time is divided into slots wherein the same signals are retransmitted over the different slots until they are successfully received by the receiver. With this, each channel can fully utilize the channel bandwidth and prevent the cross talk problem.

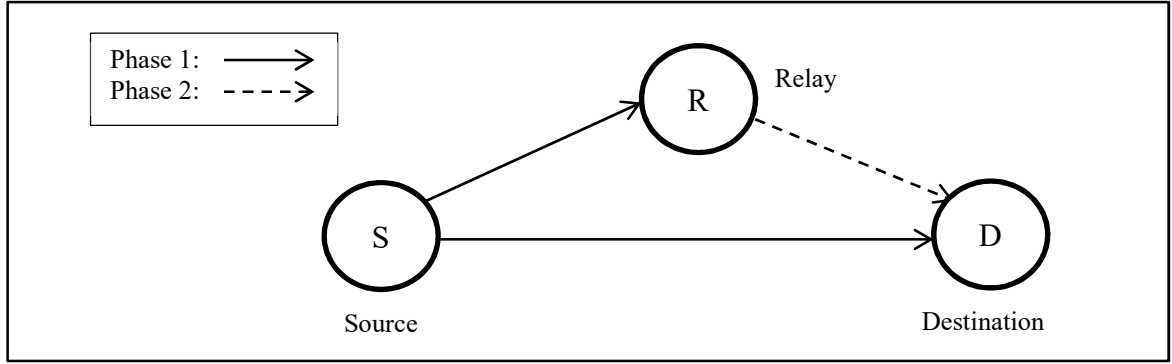


Figure 3.1: Proposed system model

A flat fading Rayleigh channel with $h_{s,d} \sim \mathcal{CN}(0, \sigma_{s,d}^2)$, $h_{s,r} \sim \mathcal{CN}(0, \sigma_{s,r}^2)$ and $h_{r,d} \sim \mathcal{CN}(0, \sigma_{r,d}^2)$, where $\mathcal{CN}(\mu, \sigma^2)$ represents complex Gaussian random variable with μ mean and σ^2 variance is considered. The channel is perturbed in the presence of frequency

offsets. Let the instantaneous SNR of the direct and relayed paths, $\gamma_{s,d} = \frac{P_S |h_{s,d}|^2}{\sigma^2}$,

$\gamma_{s,r} = \frac{P_S |h_{s,r}|^2}{\sigma^2}$ and $\gamma_{r,d} = \frac{P_R |h_{r,d}|^2}{\sigma^2}$ respectively, the average received SNR are, $\bar{\gamma}_{s,d} =$

$\frac{P_S \sigma_{s,d}^2}{N_0}$, $\bar{\gamma}_{s,r} = \frac{P_S \sigma_{s,r}^2}{N_0}$ and $\bar{\gamma}_{r,d} = \frac{P_R \sigma_{r,d}^2}{N_0}$, respectively.

By employing DD modulation in the proposed cooperative communication system, the received data signal at the relay during the broadcast phase transmission can be expressed as:

$$y_{s,r}[n] = P_S h_{s,r} e^{j2\pi f_{s,r} n} z[k] + e_{s,r}[n] \quad (3.1)$$

and at the destination as:

$$y_{s,d}[n] = P_S h_{s,d} e^{j2\pi f_{s,d} n} z[k] + e_{s,d}[n] \quad (3.2)$$

where P_S represents the power transmitted by the source, while $h_{s,r}$ and $h_{s,d}$ are the channel for the source-relay (i.e. relayed) and source-destination (i.e. direct) paths, respectively. $f_{s,r}$ and $f_{s,d}$ are the physical frequency offsets in Hz for both paths. The frequency offsets can

be normalized as $\omega_i = 2\pi f_i n$. $e_{s,r}$ and $e_{s,d}$ are the signal-independent complex AWGN with zero mean and N_o variance on the two paths. Assuming that the noise terms $e_{s,d}[n]$ and $e_{s,r}[n]$ are independent, it can be shown that the noise terms are zero-mean complex Gaussian random variable with distribution $\mathcal{CN} \sim (0, P_s \sigma_{s,d}^2 + N_o)$ and $\mathcal{CN} \sim (0, P_s \sigma_{s,r}^2 + N_o)$ [13].

In the second phase transmission, the source stays in the idle or silent state as a remedy to retain the orthogonality between the transmissions since all the data sequence is transmitted in TDM access manner. At the same time, the relay amplifies the received signal of (3.1) with a fixed gain G_i .

The fixed gain is defined as:

$$G_i = \frac{P_{R_i}}{P_s \sigma_{s,r_i}^2 + N_o} \quad (3.3)$$

where P_{R_i} is the average power transmitted by the i th relay with the assumption that the total power transmitted for all transmission path is $P_T = P_s + P_R$, σ_{s,r_i}^2 denotes the variance of the i th source-relay path and N_o is the variance of the AWGN. i denotes i th relays. When absolutely no CSI available, the gain in (3.3) is an optimal choice [28]. The processed signal is then re-forwards to the destination and is written as:

$$y_{s,r_i,d}[l] = G h_{r_i,d} e^{j2\pi f_{r_i,d} l} y_{s,r_i}[l] + e_{s,r_i,d}[l] \quad (3.4)$$

where $h_{r_i,d}$ denotes the equivalent double Rayleigh channel. It is noted that $h_{r_i,d}$ is Gaussian distributed with mean zero and variances 1. $f_{r_i,d}$ is the frequency offset and $e_{r_i,d}$ represents the AWGN between the relay-destination path. Furthermore, the equivalent noise component $e_{s,r_i,d}$ can be written as:

$$e_{s,r_i,d}[l] = G_i h_{r_i,d}[l] e_{sr_i}[l] + e_{r_i,d}[l] \quad (3.5)$$

Since $h_{r_i,d}$ is Gaussian distributed, $e_{s,r_i,d}$ and $y_{s,r_i,d}$ are complex Gaussian random variables with mean zero and variances of $\sigma_{s,r_i,d}^2 = G_i^2 |h_{r_i,d}[l]|^2 + 1$ and $\sigma_{s,r_i,d}^2(\rho_{s,r_i,d} + 1)$ where

$$\rho_{s,r_i,d} = G_i P_S |h_{r_i,d}[l]|^2 / \sigma_{s,r_i,d}^2 \quad (3.6)$$

denotes the averaged received SNR conditioned on $h_{r_i,d}[l]$.

The following section discusses the channel modelling for simulation and performance analysis purposes.

3.2.1 Channel Simulation Model

Motivated by mobile networks applications, it is important to consider and mathematically model the wireless channels in terms of the channel coefficients with respect to time index n . The significant features of the channel model include: half-duplex operation, time varying Rayleigh multipath fading and AWGN. In order to deploy time-varying channel (i.e. slow fading and fast fading) environment, a mobile-to-mobile simulator based on SoS approach in [94] is studied and applied for the simulation purpose.

So far this simulation model has only been applied to the MIMO systems, for instance, in [95] and references therein. The mobile-to-mobile channels narrowband complex envelope adopted in [94] can be simulated as:

$$h(t) = h_i(t) + jh_q(t) \quad (3.7)$$

$$h_i(t) = \sqrt{\frac{2}{LM}} \sum_{l,m=1}^{L,M} \cos\{2\pi f_1 \cos(\alpha_l) t + 2\pi f_2 \cos(\beta_m) t + \phi_{lm}^i\} \quad (3.8)$$

$$h_q(t) = \sqrt{\frac{2}{LM}} \sum_{l,m=1}^{L,M} \cos\{2\pi f_1 \cos(\alpha_l) t + 2\pi f_2 \cos(\beta_m) t + \phi_{lm}^q\} \quad (3.9)$$

where

$$\alpha_n = \frac{2\pi l - \pi + \theta}{4L} \quad (3.10)$$

$$\beta_m = 2\left(\frac{2\pi m - \pi + \varphi}{4M}\right) \quad (3.11)$$

where the index l and m indicate the number of multipath from the source to L scatterers located on the source end ring and from the M scatterers on the destination end ring to the destination. The random phases $\phi_m \sim U[-\pi, \pi]$ are mutually independent for all l, m pairs. Likewise, θ and φ are mutually independent on $U[-\pi, \pi]$ as well.

3.2.2 Channel Model

In order to create a mobility environment among the users, it is important to model the fading channel that changes over time. The parameter that related to the channel variation is the channel auto-correlation function which is given based on Jakes' model as:

$$\varphi(n) = \sigma_h^2 J_0(2\pi f_D T_S n) \quad (3.12)$$

where J_0 is the Bessel function order of zero for the first kind, f_D is the Doppler frequency and T_S represents the symbol duration. It is noted that the Doppler frequency can be related to the users' mobility and can be obtained as $f_D = v f_c / c$, wherein f_c represents the system carrier frequency and v denotes the user's velocity. $c = 3 \times 10^8$ m/s is speed of light. The channel variation rate can be also called the normalized Doppler frequency $f_D T_S$.

The channel variation – normalized Doppler frequency relation is plotted in Figure 3.2 and Figure 3.3. Both graphs are simulated under three different scenarios as tabulated in Table 3.1. Figure 3.2 illustrates the amplitude for the direct (i.e. the source-destination) Rayleigh fading channel and its corresponding auto-correlation values. On top of that, Figure 3.3 displays the amplitude for the relayed (i.e. the source-relay and relay-destination) Rayleigh fading channel and also its auto-correlation functions. It can be observed from the

figures that, the channel variation rate is related to the normalized Doppler frequency as well as the values of auto-correlation. When $f_D T_S = 0.001$, the channel coefficients are static over time and its corresponding auto-correlation exhibit small curve. Such channel is named slow fading. For $f_D T_S = 0.01$ and 0.05 , the channel varies faster over time and its auto-correlation values also changes rapidly. Such channels are named fast fading. The normalized Doppler frequency can be converted in terms of vehicle speeds for a wireless systems. For instance, in a wireless system where the carrier frequency $f_c = 2\text{GHz}$ and symbol duration $T_S = 0.1\text{ms}$, the corresponding Doppler shifts will be approximately $f_D = \frac{f_D T_S}{T_S} = 10, 100, 500\text{Hz}$. The corresponding speeds can be obtained as $v = \frac{c f_D}{f_c} = 5.4, 54, 270 \text{ km/hr}$.

Table 3.1: The channel condition over a Rayleigh flat fading environment for various fading rate

Case	Normalized Doppler frequency			Channel Condition
	$f_{D_{s,r}} T_S$	$f_{D_{r,d}} T_S$	$f_{D_{s,d}} T_S$	
I	0.001	0.001	0.001	Slow fading
II	0.01	0.01	0.01	Fairly fast fading
III	0.05	0.05	0.05	Fast fading

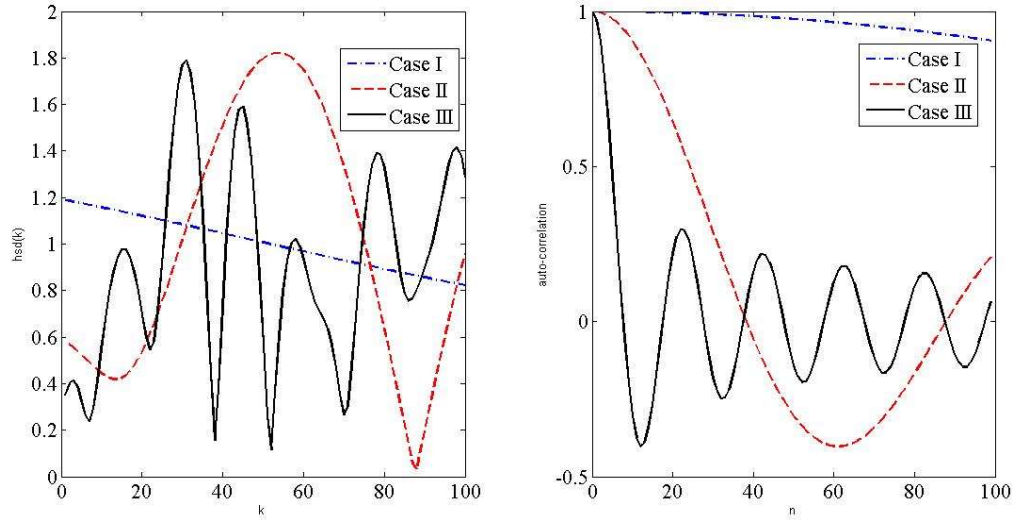


Figure 3.2: Direct link channel coefficients and its corresponding autocorrelation values of different cases

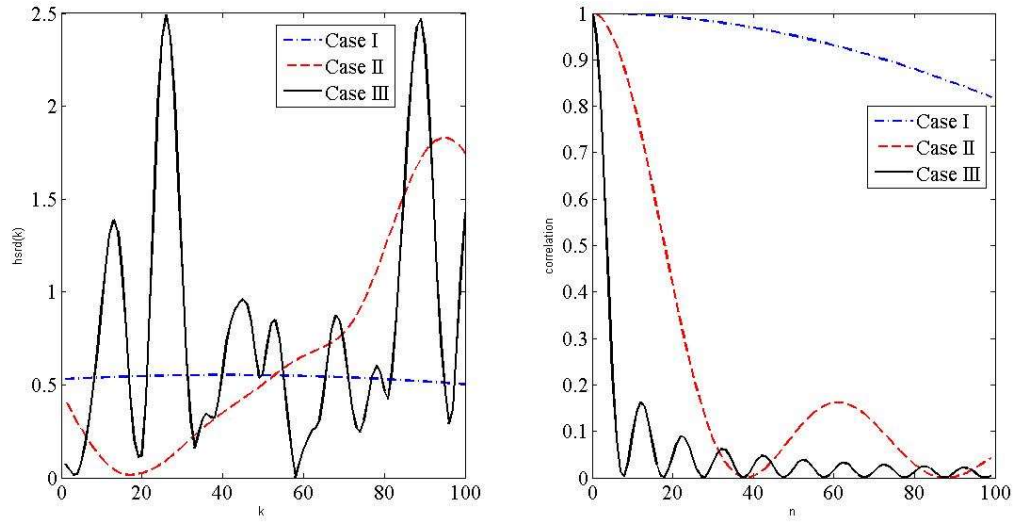


Figure 3.3: Relayed channel and its corresponding autocorrelations of different cases

(i) Scenario with Source and Relay Static, Destination Mobile

By using the AF cooperative relaying protocol, a relayed link is constructed for the source-destination via a relay. In the case where the source and relay are static and the

destination moves, the three consecutive source-relay channel are approximately equal such that:

$$h_{s,r}[k] \approx h_{s,r}[k-1] \approx h_{s,r}[k-2] \quad (3.13)$$

The assumption is valid when the channel is slow fading. Contrastly, the assumption is violated under the scenario of fast fading due to its statistical properties and time-varying properties. Additionally, because of the fast fading environment, it is complex to obtain the perfectly estimated channel estimation. For that reason, in the case when no prior channel knowledge is predictable at the relay and destination, AR model is recommended to describe the time-varying channels due to its stochastic process; the future values are estimated pertaining to a weighted sum of past values. Contingent on autoregression model concept, AR(1) is considered as the first order process, wherein the present value is referred to the immediately preceding value. Since double differentially modulated symbols are transmitted, AR(2) process is implemented with the current value calculated proceeding from the previous two values, particularly in time-varying channel (i.e. when the destination node moves). Accordingly, the relay-destination channel can be pronounced by the AR(2) model as:

$$h_{r,d}[k] = \alpha_{r,d}^2 h_{r,d}[k-2] + \alpha_{r,d} h_{r,d}[k-1] + 2\sqrt{1 - \alpha_{r,d}^2} \varepsilon_{r,d}[k] \quad (3.14)$$

where $\alpha_{r,d} = J_0(2\pi f_{D_{r,d}} n) \leq 1$ is the Jake's autocorrelation model [96]. The autocorrelation value indicates the channel variation of the relay-destination channel. It is noted from (3.14) that the lower value of α , the channel varies greater. When $\alpha = 0$, the channel realization is i.i.d. On the contrary, the channel modelled quasi-static fading when $\alpha = 1$ [28]. $f_{D_{r,d}}$ denotes the normalized Doppler frequency incurred by the relay-destination channel and $\varepsilon_{s,r}[k]$ is the white Gaussian noise with zero mean and variance 1 which is

independent of $h_{s,r}[k-1]$ and $h_{s,r}[k-2]$. Therefore, the relayed link can be obtained by multiplying $h_{s,r}$ and $h_{r,d}$ as:

$$\begin{aligned} h[k] = & \alpha_{r,d}^2 h[k-2] h_{s,r}[k-1] + \alpha_{r,d} h_{s,r}[k-2] h[k-1] \\ & + 2\sqrt{1 - \alpha_{r,d}^2} h_{s,r}[k-2] h_{s,r}[k-1] \varepsilon_{r,d}[k] \end{aligned} \quad (3.15)$$

which is an AR(2) model with parameter $2\sqrt{1 - \alpha_{r,d}^2} h_{s,r}[k-2] h_{s,r}[k-1] \varepsilon_{r,d}[k]$ as the input white noise.

(ii) Scenario with Destination Static, Source and Relay Mobile

In a scenario when destination is static while the source and relay moves, the channel between the source-relay and relay-destination become time-varying. Therefore, the source-destination can be modelled with AR(2) model as follows:

$$h_{s,d}[k] = \alpha_{s,d}^2 h_{s,d}[k-2] + \alpha_{s,d} h_{s,d}[k-1] + 2\sqrt{1 - \alpha_{s,d}^2} \varepsilon_{s,d}[k] \quad (3.16)$$

where $\alpha_{s,d} = J_0(2\pi f_{D_{s,d}} n) \leq 1$ is the autocorrelation of the source-destination channel. $f_{D_{s,d}}$ denotes the normalized Doppler frequency at the source-destination channel and $\varepsilon_{s,d}[k]$ is the white Gaussian noise with zero mean and variance 1 which is independent of $h_{s,d}[k-1]$ and $h_{s,d}[k-2]$.

Similarly, source-relay channel can be described using the AR(2) model as:

$$h_{s,r}[k] = \alpha_{s,r}^2 h_{s,r}[k-2] + \alpha_{s,r} h_{s,r}[k-1] + 2\sqrt{1 - \alpha_{s,r}^2} \varepsilon_{s,r}[k] \quad (3.17)$$

where $\alpha_{s,r} = J_0(2\pi f_{D_{s,r}} n) \leq 1$ is the autocorrelation of the source-relay channel. $f_{D_{s,r}}$ denotes the normalized Doppler frequency at the relay-destination channel and $\varepsilon_{s,r}[k]$ is the white Gaussian noise with zero mean and variance 1 which is independent of $h_{s,r}[k-1]$ and $h_{s,r}[k-2]$.

By multiplying both time-varying channels, (3.17) and (3.14), the relayed links is given as:

$$h[k] = (\alpha_{s,r}\alpha_{r,d})^2 h[k-2] + \alpha_{s,r}\alpha_{r,d} h[k-1] + \Delta[k] \quad (3.18)$$

where

$$\begin{aligned} \Delta[k] = & \alpha_{s,r}^2 h_{s,r}[k-2] \sqrt{1 - \alpha_{r,d}^2 \varepsilon_{r,d}[k]} + 2\alpha_{s,r} h_{s,r}[k-1] \sqrt{1 - \alpha_{r,d}^2 \varepsilon_{r,d}[k]} \\ & + 2\alpha_{r,d}^2 h_{r,d}[k-2] \sqrt{1 - \alpha_{s,r}^2 \varepsilon_{s,r}[k]} + 2\alpha_{r,d} h_{r,d}[k-1] \sqrt{1 - \alpha_{s,r}^2 \varepsilon_{s,r}[k]} \\ & + 4\sqrt{(1 - \alpha_{s,r}^2)(1 - \alpha_{r,d}^2) \varepsilon[k]} \end{aligned}$$

represents the transmitted signals and the white Gaussian noise of the time-varying channel which is independent of $h[k-1]$ and $h[k-2]$.

Additionally, $h[k-2] = h_{s,r}[k-2]h_{r,d}[k-2]$ and $h[k-1] = h_{s,r}[k-1]h_{r,d}[k-1]$ represents the relayed channel for $k-2$ and $k-1$, respectively. However, it is impractical to use the $\Delta[k]$ for the performance analysis. Thus, $\Delta[k]$ is further simplified with a simpler version as:

$$\hat{\Delta}[k] = \left\{ \sqrt{1 - \alpha_{r,d}^2} h_{r,d}[k-2] + \sqrt{1 - \alpha_{r,d}^2} h_{r,d}[k-1] \right\} e_{s,r}[k] \quad (3.19)$$

Substitutes (3.19) into (3.18), the relayed channel model can be written as:

$$\begin{aligned} h[k] = & (\alpha_{s,r}\alpha_{r,d})^2 h[k-2] + \alpha_{s,r}\alpha_{r,d} h[k-1] \\ & + \left\{ \sqrt{1 - \alpha_{r,d}^2} h_{r,d}[k-2] + \sqrt{1 - \alpha_{r,d}^2} h_{r,d}[k-1] \right\} e_{s,r}[k] \end{aligned} \quad (3.20)$$

The theoretical auto-correlation of $h[k]$ can be found by implementing the auto-correlation product of both source-relay and relay-destination channels. By multiplying (3.20) with $h^*[k-2]$ and taking the expectation value of both side of the resulting equation, one gets:

$$E\{h[k]h^*[k-2]\} = \alpha^2 E\{h[k-2]h^*[k-2]\} + E\{\hat{\Delta}[k]h^*[k-2]\} \quad (3.21)$$

As both $\hat{\Delta}[k]$ and $h[k-2]$ are uncorrelated with each other, $E\{\hat{\Delta}[k]h^*[k-2]\} = 0$ and therefore:

$$E\{h[k]h^*[k-2]\} = \alpha^2 = \alpha_{s,r}^2 \alpha_{r,d}^2 \quad (3.22)$$

(iii) Scenario with Source, Relay and Destination Mobile

In this scenario, all terminals are mobile and follows the mobile-to-mobile channel model [94]. The channel model of (3.19) can be adopted for analytical purposes with the condition that the suitable auto correlation of the relayed channels is to be substituted and implemented according to the model.

3.3 Detection Schemes at the Receiver for Cooperative Diversity

This section describes the differential detection schemes at the intended destination in regards to the conventional symbol-by-symbol and proposed multiple symbols transmission for cooperative diversity networks in time-varying channels. In the case at high Doppler frequencies, an irreducible floor is formed when conventional DD scheme is implemented. In order to circumvent the potential-erosion problem, MSDD was proposed for non-coherent schemes. However, the system suffered from the exponential increasing complexity when the symbol detection block length increases. As a further advance, the idea of SD is integrated in the multiple symbol based transmission in order to limit the increased complexity of the system.

3.3.1 Conventional Double Differential Detection

It is assumed that DDPSK modulation is applied in the relayed network as shown in Figure 3.1. To meet the International Mobile Telecommunication (IMT)-2000 standards, a system is required to provide at least 0.2 Mbit/s. Thus, 10^6 information bits are generated

for simulation purposes under mobile communication environment. The generated $\log_2 M$ -bit information are encoded into information symbol $v[n] \in \mathcal{M}_C = \{e^{\frac{j2\pi nm}{M}}, m = 1, 2, \dots, M-1\}$ where C denotes the signal constellation corresponding to a normalized M -PSK modulation. The symbol is then double differentially modulated as (3.22) referring to Figure 3.4. DD modulation is proposed in accordance to eliminate the channel and frequency offsets estimation. The simulation parameters are tabulated in Table 3.2.

Table 3.2: Simulation parameters

Simulation Parameters	Values
Information bits	10^6 bits
SNR	0 – 40 dB
Modulation	BPSK, QPSK
Noise Channels	AWGN, Rayleigh
Source Power, P_s	$0.5P_T$
Relay Power, P_R	$0.5P_T$

Before the data information transmission, the double differentially modulated symbols from the DD Modulation block in Figure 3.4 can be expressed as follows:

$$z[k] = x[k]y[k-1]z[k-1] \quad (3.23)$$

where $z[k-1] = y[k-2]$ with all symbols being unit magnitude i.e. $|x| = |y| = |z| = 1$.

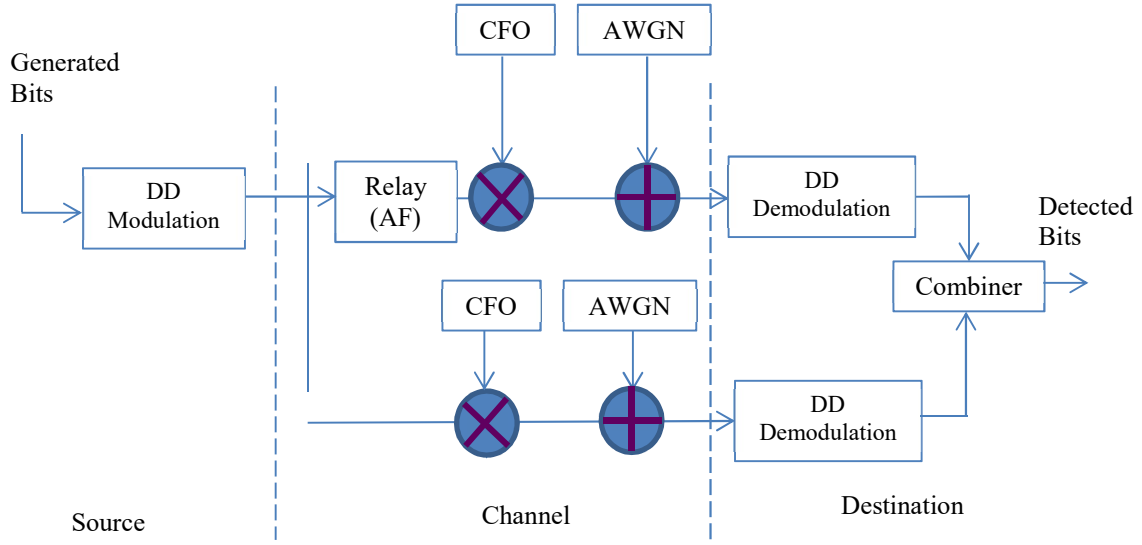


Figure 3.4: Block diagram of DD transmission for single relay cooperative communication system

Since DD modulation is employed in the cooperative communication system with the effect of frequency offsets due to the frequency mismatch at the receiver, the doubly differentiated signal at the destination during the first phase transmission can be expressed as:

$$y_{s,d} = \sqrt{P_S} h_{s,d} e^{2\pi f_{s,d} n} z[n] + e_{s,d}[n], \quad n = 0, 1, \dots, \quad (3.24)$$

and at the relay as:

$$y_{s,r} = \sqrt{P_S} h_{s,r} e^{2\pi f_{s,r} n} z[n] + e_{s,r}[n], \quad n = 0, 1, \dots, \quad (3.25)$$

where P_S indicates the transmitted power by the source, $h_{s,d}$ and $h_{s,r}$ are the channel gains. The channel gains are complex circularly symmetric Gaussian distributed where $h_{s,d} \sim \mathcal{CN}(0, \sigma_{s,d}^2)$ and $h_{s,r} \sim \mathcal{CN}(0, \sigma_{s,r}^2)$. $f_{s,d}$ and $f_{s,r}$ are the normalized frequency offsets between source-destination and source-relay channels, while $e_{s,d}$ and $e_{s,r}$ are the circularly symmetric AWGN noise for both channels with $e_{s,d} \sim \mathcal{CN}(0, N_0)$ and $e_{s,r} \sim \mathcal{CN}(0, N_0)$.

During the second phase transmission, the relay amplifies the signals received from the source, and retransmits the processed signals towards the destination.

$$y_{r,d} = \sqrt{G}h_{r,d}e^{2\pi f_{r,d}l}y_{s,r}[l] + e_{r,d}[l] \quad (3.26)$$

where l is time index to indicate the time difference between the first and second phase transmission, G indicates the gain at the relay, $h_{r,d}$ is the channel gains, $f_{r,d}$ denotes the normalized frequency offsets between the relay-destination channels, and $e_{r,d}$ is the circularly symmetric AWGN noise for the link with $e_{r,d} \sim \mathcal{CN}(0, N_0)$. Since channel knowledge is not available at the relay terminal, a fixed relay gain is proposed based on the channel variance, $\sigma_{s,r}^2$. Hence, referring to Figure 3.4, the relay gain, G in the relay block is given as:

$$G = \frac{P_R}{P_S \sigma_{s,r}^2 + \sigma^2} \quad (3.27)$$

where P_R is the power at the relay and σ^2 is the variance of AWGN. It is assumed that the total transmitted power for the system is $P = P_S + P_R$.

At the destination, direct combiner for the DD demodulation referring to Figure 3.4 combiner block is proposed as:

$$\begin{aligned} d[k] = & (y_{s,d}[n]y_{s,d}^*[n-1])(y_{s,d}[n-1]y_{s,d}[n-2])^* \\ & + (y_{r,d}[l]y_{r,d}^*[l-1])(y_{r,d}[l-1]y_{r,d}[l-2])^* \end{aligned} \quad (3.28)$$

The combined signal is executed and decoded following the ML [22] as:

$$\hat{y}[n] = \text{sign}\{\text{Re}\{q[n]\}\} \quad (3.29)$$

3.3.2 Performance Analysis of Double Differential Detection

In this section, performance of the average BER is numerically analysed for the DD transmission using PSK constellation in the AF Cooperative wireless networks.

The PDFs of $\gamma_{s,d}$, $\gamma_{s,r}^i$ as well as $\gamma_{r,d}^i$ following the Rayleigh fading characteristic, can be expressed as [26]:

$$\begin{aligned}
P_{\gamma_{s,d}}(\gamma_{s,d}) &= \frac{1}{\bar{\gamma}_{s,d}} e^{-\gamma_{s,d}/\bar{\gamma}_{s,d}} \\
P_{\gamma_{s,r}}(\gamma_{s,r}^i) &= \frac{1}{\bar{\gamma}_{s,r}} e^{-\gamma_{s,r}^i/\bar{\gamma}_{s,r}} \\
P_{\gamma_{r,d}}(\gamma_{r,d}^i) &= \frac{1}{\bar{\gamma}_{r,d}} e^{-\gamma_{r,d}^i/\bar{\gamma}_{r,d}}
\end{aligned} \tag{3.30}$$

where $\bar{\gamma}_{s,d} = P_S \sigma_{s,d}^2 \bar{\gamma}$, $\bar{\gamma}_{s,r} = P_S \sigma_{s,r}^2 \bar{\gamma}$ and $\bar{\gamma}_{r,d} = P_R \sigma_{r,d}^2 \bar{\gamma}$ are the average SNR for each channel links.

From (3.24), (3.25) and (3.26) the average SNR of the relayed link for the DD modulation is given by:

$$\bar{\gamma}_{s,r,d} = \frac{G^2 \gamma_{s,r} |h_{s,r}|^2 |h_{r,d}|^2}{1 + G^2 |h_{s,r}|^2 |h_{r,d}|^2} \tag{3.31}$$

Note that the above expression can be seen from [16] Equation (8), [95,96] for a dual-hop transmission scheme with a fixed gain relay. The average SNR is dependent on the channel knowledge, $h_{s,r}$ and $h_{r,d}$ which are unavailable at the destination. Thus, $h_{s,r}$ and $h_{r,d}$ are replaced with its variances, $\sigma_{s,r}^2$ and $\sigma_{r,d}^2$.

The PDF of $\gamma_{s,r,d}$ can be written as:

$$\begin{aligned}
p(\gamma_{s,r,d}) &= \frac{2\sigma^2 P_S \sigma_{s,r}^2 + 2\sigma^4}{P_S P_R \sigma_{s,r}^2 \sigma_{r,d}^2} \exp\left(-\frac{\sigma^2 \gamma}{P_S \sigma_{s,r}^2}\right) \times K_0(2\beta) \\
&+ \frac{2\sigma^4}{P_S \sigma_{s,r}^2 \sigma_{r,d}^2} \times \sqrt{\frac{P_R \sigma_{r,d}^2 (P_S \sigma_{s,r}^2 + \sigma^2)}{P_S \sigma_{s,r}^2 \sigma^2}} \exp\left(-\frac{\sigma^2 \gamma}{P_S \sigma_{s,r}^2}\right) \times K_1(2\beta)
\end{aligned} \tag{3.32}$$

where $K_0(\cdot)$ represents the zeroth-order modified Bessel function of the second kind, and $K_1(\cdot)$ represents the first order modified Bessel function of the second kind.

Equation (3.32) is proven with aid of results given in [97] Section III]. According to [1], the conditional BER of the differential transmission is given by:

$$P_b(\gamma_T) = \frac{1}{2^{2L-1}} e^{-\gamma_T} \sum_{n=0}^{L-1} C_n \gamma_T^n \quad (3.33)$$

where γ_T is the average total SNR at the destination and C_n is defined as:

$$C_n = \frac{1}{n!} \sum_{k=0}^{L-1-n} \binom{2L-1}{k} \quad (3.34)$$

With the independent 2-channel reception, the conditional BER of the binary differential PSK can be expressed as [66], Equation (12.1-13)]:

$$P_b(\gamma) = \frac{1}{8} (4 + \gamma_T) e^{-\gamma_T} \quad (3.35)$$

where the total average SNR at the output of the combining process for the direct and relayed paths is:

$$\gamma_T = \gamma_{s,d} + \gamma_{s,r,d} \quad (3.36)$$

The BER of DDAAF system with BPSK modulation across the joint PDF of $\gamma_{s,d}$ and $\gamma_{s,r,d}$ is averaged over the channels and is expressed by:

$$P_b = \frac{1}{8} \int_0^\infty \int_0^\infty (4 + \gamma_{s,d} + \gamma_{s,r,d}) e^{-(\gamma_{s,d} + \gamma_{s,r,d})} \times p_{s,d}(\gamma_{s,d}) \times p_{s,r,d}(\gamma_{s,r,d}) d\gamma_{s,d} d\gamma_{s,r,d} \quad (3.37)$$

The approximate BER of cooperative diversity with DD modulation is given as 3.3314 [99]:

$$\begin{aligned}
P_b = & \frac{\exp(0.5)}{16} \times \left\{ \frac{4\sigma^6 + 5P_S\sigma_{s,d}^2\sigma^4}{(2\sigma^2 + P_S\sigma_{s,d}^2)^2(2\sigma^2 + P_S\sigma_{s,r}^2)} W_{-1.0,0.5}(\beta) \right. \\
& + \frac{\sqrt{2(\sigma^2 + P_S\sigma_{s,r}^2)}(4\sigma^5 + 5P_S\sigma_{s,d}^2\sigma^3)}{\sqrt{P_R(2\sigma^2 + P_S\sigma_{s,r}^2)}\sigma_{r,d}(2\sigma^2 + P_S\sigma_{s,d}^2)^2} \times W_{-0.5,0}(\beta) \\
& + \frac{16P_S\sigma_{s,r}^2\sigma^4}{(2\sigma^2 + P_S\sigma_{s,r}^2)^2(2\sigma^2 + P_S\sigma_{s,d}^2)} \times W_{-2,0.5}(\beta) \\
& \left. + \frac{8\sqrt{2(\sigma^2 + P_S\sigma_{s,r}^2)}P_S\sigma_{s,r}^2\sigma^4}{\sqrt{P_R\sigma_{r,d}(2\sigma^2 + P_S\sigma_{s,r}^2)^3(2\sigma^3 + P_S\sigma_{s,d}^2\sigma)}} \times W_{-1.5,0}(\beta) \right\}
\end{aligned} \tag{3.38}$$

where

$$\beta = \frac{\sigma^2}{P_R\sigma_{r,d}^2} \tag{3.39}$$

with the assumption that $\sigma^2 = 1$ and symmetric case where $P_S = P_R = 0.5$.

The Whittaker function can be written as [100]:

$$W_{k,\mu}(z) = \exp\left(-\frac{z}{2}\right) z^{\left(\mu+\frac{1}{2}\right)} U\left(\mu - k + \frac{1}{2}, 1 + 2\mu; z\right) \tag{3.40}$$

where $U(a, b, z)$ represents the confluent hypergeometric function and can be expressed as:

$$U(a, b, z) = \frac{\Gamma(1-b)}{\Gamma(a-b+1)} M(a, b, z) + \frac{\Gamma(b-1)}{\Gamma(a)} z^{1-b} M(a-b+1, 2-b, z) \tag{3.41}$$

3.3.3 Multiple Symbol Detection

Although DD transmission is able to circumvent the degradation because of frequency offset as compared to single differential method, higher SNR power is required to achieve the similar BER performance of the coherent detection. Hence, MSDD detection is proposed in [35] to mitigate the SNR loss.

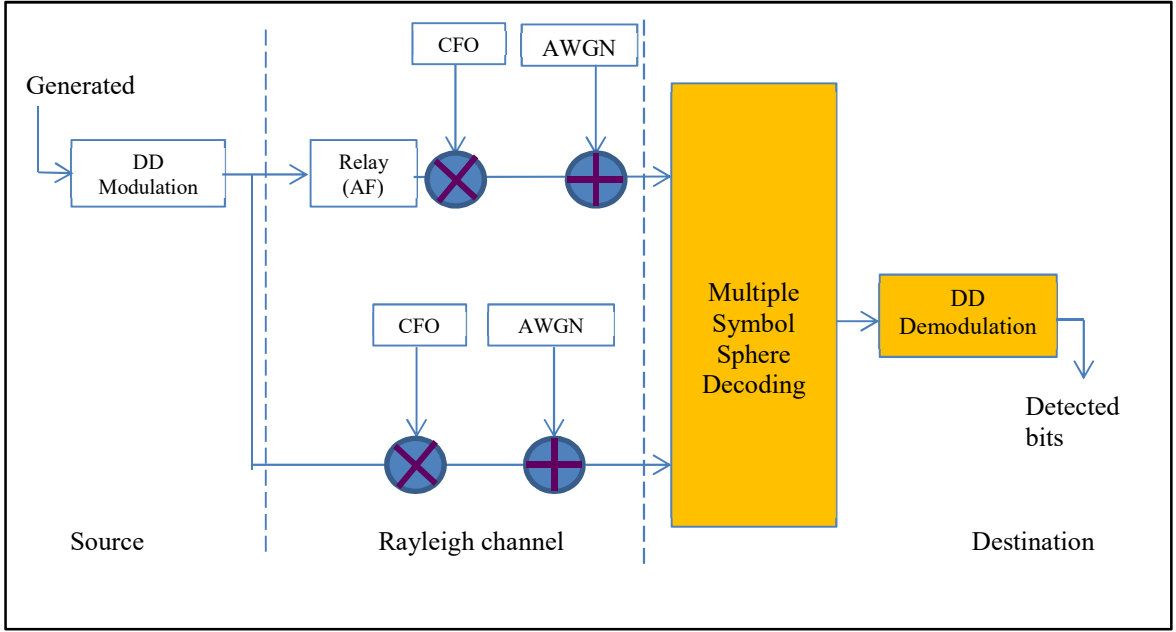


Figure 3.5: Block diagram of proposed MSD detection

When channel knowledge is unavailable at the destination, multiple symbols based detection offers better performance in terms of reduced error-rate over the conventional symbol-by-symbol detection which requires frequent switching at the detection.

The $N > 2$ symbols at the receiver can be written as:

$$\mathbf{y} = (y[1], y[2], \dots, y[N])^t \quad (3.42)$$

Referring to (3.24), (3.25) and (3.26) in Section 3.3.1, the $N > 2$ received symbols at the destination via directly transmitted link from the source can be written as:

$$\mathbf{y}_{sd} = \sqrt{P_S} \cdot e^{jw_{s,d}} \text{diag}\{\mathbf{z}\} \mathbf{h}_{s,d} + \mathbf{e}_{s,d} \quad (3.43)$$

where

$$\mathbf{z} = [z(1), \dots, z(N)]^t$$

$$\mathbf{h}_{s,d} = \text{diag}[h_{s,d}(1), \dots, h_{s,d}(N)]^t$$

$$\mathbf{e}_{s,d} = \text{diag}[e(1), \dots, e(N)]^t$$

$w_{s,d}$ = frequency offsets along the source-destination link

P_S = source power

For a two-hop relay network, the received symbols at the destination can be obtained as:

$$\mathbf{y} = G\sqrt{P_s}e^{jw} \text{diag}\{\mathbf{z}\} \cdot \mathbf{h}_{s,r} \cdot \mathbf{h}_{r,d} + e \quad (3.44)$$

where

$$\mathbf{z} = \text{diag}[z(1), \dots, z(N)]^t$$

$$\mathbf{h}_{r,d} = [h_{r,d}(1), \dots, h_{r,d}(N)]^t$$

$$\mathbf{h}_{s,r} = \text{diag}[h_{s,r}(1), \dots, h_{s,r}(N)]^t$$

$$\mathbf{e} = [e(1), \dots, e(N)]^t = \text{equivalent noise of the relayed links}$$

w = frequency offsets along the relayed links

G = Relay gain

P_s = Source power

At the destination, non-coherent ML detection is proposed due to its optimum performance without requiring the knowledge of channel. Thus, the optimal decision for the relayed and direct link transmission is expressed as:

$$\hat{z} = \arg \max f(\mathbf{r}_{s,d}|\mathbf{z}) \cdot \arg \max f(\mathbf{r}_{r,d}|\mathbf{z}, \mathbf{h}_{r,d}) \quad (3.45)$$

where $f(\cdot | \cdot)$ represents the conditioned PDF of the channel output.

Assuming that the equivalent fading and noise are zero-mean complex Gaussian processes, the PDF of the received $\mathbf{y}_{s,d}$ conditioned on the transmitted data $\mathbf{z}$ is written as [56]:

$$f(\mathbf{y}_{s,d}|\mathbf{z}) = \frac{\exp(-\text{Tr}\{\mathbf{y}_{s,d}^H \boldsymbol{\Psi}_{r,s,d}^{-1} \mathbf{y}_{s,d}\})}{\pi^N \det(\boldsymbol{\Psi}_{r,s,d})} \quad (3.46)$$

and the PDF of $\mathbf{y}_{r,d}$ conditioned on $\mathbf{z}$ and $\mathbf{h}_{r,d}$ is:

$$f(\mathbf{y}_{r,d}|\mathbf{z}, \mathbf{h}_{r,d}) = \frac{\exp(-\text{Tr}\{\mathbf{y}_{r,d}^H \boldsymbol{\Psi}_{y,r,d}^{-1} \mathbf{y}_{r,d}\})}{\pi^N \det(\boldsymbol{\Psi}_{y,r,d})} \quad (3.47)$$

where $\frac{1}{\pi^N \det(\boldsymbol{\Psi})}$ is independent of $\mathbf{z}$.

The conditional variance of $\mathbf{y}_{s,d}$ and $\mathbf{y}_{r,d}$ from (3.46) and (3.47) yield:

$$\boldsymbol{\Psi}_{\mathbf{y}_{s,d}} = E\{\mathbf{y}_{s,d}\mathbf{y}_{s,d}^H|\mathbf{z}\} = P_S \cdot \mathbf{z} \cdot \boldsymbol{\Psi}_{\mathbf{h}_{s,d}} \cdot \mathbf{z}^* + N_{s,d}\mathbf{I}_N \quad (3.48)$$

and

$$\boldsymbol{\Psi}_{\mathbf{y}_{r,d}} = E\{\mathbf{y}_{r,d}\mathbf{y}_{r,d}^H|\mathbf{z}, \mathbf{h}_{r,d}\} = G^2 P_S \cdot \mathbf{z} \cdot \mathbf{h}_{r,d} \cdot \boldsymbol{\Psi}_{\mathbf{h}_{s,r}} \cdot \mathbf{z}^* + (1 + G^2 \cdot \sigma_{r,d}^2)N_{s,d}\mathbf{I}_N \quad (3.49)$$

It is considered that all the fading and noise are complex Gaussian with mean = 0. Therefore, the conditional covariance received vector $\mathbf{y}$, $\boldsymbol{\Psi}_{\mathbf{h}_{s,d}}$ and $\boldsymbol{\Psi}_{\mathbf{h}_{s,r,d}}$ are complex Gaussian vector with mean = 0 where $\boldsymbol{\Psi}_{\mathbf{h}_{s,d}}$ represents the autocorrelation matrix of $\mathbf{h}_{s,d}$ and its expectation can be written as:

$$\boldsymbol{\Psi}_{\mathbf{h}_{s,d}} = E\{\mathbf{h}_{s,d}\mathbf{h}_{s,d}^*\} = \text{toeplitz}\{\varphi_{s,d}(0), \dots, \varphi_{s,d}(N-2)\} \quad (3.50)$$

where $\varphi(i) = E\{\mathbf{h}\mathbf{h}^*\} = J_0(2\pi f_{D_{s,d}})$. Similarly, $\boldsymbol{\Psi}_{\mathbf{h}_{r,d}}$ denotes the autocorrelation matrix of $\mathbf{h}_{r,d}$. Its expectation is expressed as:

$$\boldsymbol{\Psi}_{\mathbf{h}_{r,d}} = E\{\mathbf{h}_{r,d}\mathbf{h}_{r,d}^*\} = \text{toeplitz}\{\varphi_{r,d}(0) \dots, \varphi_{r,d}(N-2)\} \quad (3.51)$$

By substituting (3.46) and (3.47) into (3.45), the ML detection approach is given by:

$$\begin{aligned} \hat{\mathbf{z}}_{\text{ML}} = \arg \max & \left\{ \frac{1}{\pi^N \det(\boldsymbol{\Psi}_{\mathbf{y}_{s,d}})} \exp[\text{Tr}(-\mathbf{y}_{s,d}^H \boldsymbol{\Psi}_{\mathbf{y}_{s,d}}^{-1} \mathbf{y}_{s,d})] \right\} \\ & \left\{ E_{\mathbf{h}_{r,d}} \left\{ \frac{1}{\pi^N \det(\boldsymbol{\Psi}_{\mathbf{y}_{r,d}})} \exp[\text{Tr}(-\mathbf{y}_{r,d}^H \boldsymbol{\Psi}_{\mathbf{y}_{r,d}}^{-1} \mathbf{y}_{r,d})] \right\} \right\} \end{aligned} \quad (3.52)$$

It is noted that from (3.52), the ML detection requires the expectation for $\mathbf{h}_{s,d}$ and $\mathbf{h}_{r,d}$, which does not result in closed-form expression. Therefore, proposed modified decision metric is expressed as follows:

$$\hat{\mathbf{z}}_{\text{ML}} = \arg \max \left\{ \frac{1}{\pi^N \det(\bar{\boldsymbol{\Psi}}_{\mathbf{y}_{s,d}})} \exp[\text{Tr}(-\mathbf{y}_{s,d}^H \bar{\boldsymbol{\Psi}}_{\mathbf{y}_{s,d}}^{-1} \mathbf{y}_{s,d})] \right\} \quad (3.53)$$

$$\frac{1}{\pi^N \det(\bar{\Psi}_{y_{r,d}})} \exp \left[\text{Tr} \left(-\mathbf{y}_{r,d}^H \bar{\Psi}_{y_{r,d}}^{-1} \mathbf{y}_{r,d} \right) \right]$$

where

$$\begin{aligned} \bar{\Psi}_{y_{s,d}} &= E_{\mathbf{h}_{s,d}} \{ \Psi_{y_{s,d}} \} = P_S \cdot \mathbf{z} \cdot \Psi_{\mathbf{h}_{s,d}} \cdot \mathbf{z}^* + N_{s,d} \mathbf{I}_N \\ &= \mathbf{z} (P_S \cdot \Psi_{\mathbf{h}_{s,d}} + N_{s,d} \cdot \mathbf{I}_N) \mathbf{z}^* = \mathbf{z} \mathbf{C}_{s,d} \mathbf{z}^* \end{aligned} \quad (3.54)$$

with

$$\mathbf{C}_{s,d} = P_S \cdot \Psi_{\mathbf{h}_{r,d}} + N_{s,d} \cdot \mathbf{I}_N \quad (3.55)$$

$$\Psi_{\mathbf{h}_{s,d}} = E \{ \mathbf{h}_{s,d} \mathbf{h}_{s,d}^* \} = \text{toeplitz} \{ \varphi_{s,d}(0), \dots, \varphi_{s,d}(N-2) \} \quad (3.56)$$

and

$$\begin{aligned} \bar{\Psi}_{y_{r,d}} &= E_{\mathbf{h}_{r,d}} \{ \Psi_{y_{r,d}} \} = G^2 P_S \cdot \mathbf{z} \cdot \Psi_{\mathbf{h}} \cdot \mathbf{z}^* + (1 + G^2 \sigma_{r,d}^2) N_{s,d} \mathbf{I}_N \\ &= \mathbf{z} (G^2 P_S \cdot \Psi_{\mathbf{h}} + (1 + G^2 \sigma_{r,d}^2) N_{s,d} \mathbf{I}_N) \mathbf{z}^* = \mathbf{z} \mathbf{C} \mathbf{z}^* \end{aligned} \quad (3.57)$$

with

$$\mathbf{C} = G^2 P_S \cdot \Psi_{\mathbf{h}} + (1 + G^2 \sigma_{r,d}^2) N_{s,d} \mathbf{I}_N \quad (3.58)$$

$$\Psi_{\mathbf{h}} = E \{ \mathbf{h}_{r,d} \Psi_{\mathbf{h}_{s,r}} \mathbf{h}_{r,d}^* \} = \text{toeplitz} \{ \varphi_{s,r}(0) \varphi_{r,d}(0), \dots, \varphi_{s,r}(N-2) \varphi_{r,d}(N-2) \} \quad (3.59)$$

Since $\frac{1}{\pi^N \det(\Psi_{y_{s,d}})}$ and $\frac{1}{\pi^N \det(\Psi_{y_{r,d}})}$ is independent of $\mathbf{z}$ using the rule of $\det\{\mathbf{AB}\} = \det\{\mathbf{BA}\}$,

(3.52) is further simplified as:

$$\hat{\mathbf{z}}_{\text{ML}} = \arg \max \{ \exp[\text{Tr}(-\mathbf{y}_{s,d}^H \Psi_{y_{s,d}}^{-1} \mathbf{y}_{s,d})] \} \{ \exp[\text{Tr}(-\mathbf{y}_{r,d}^H \bar{\Psi}_{y_{r,d}}^{-1} \mathbf{y}_{r,d})] \} \quad (3.60)$$

Thus, the MSDDSD decision rule can be obtained by searching the sequence $\hat{\mathbf{z}}$ that maximizes the conditional PDF. The concept is similar to finding the sequence $\hat{\mathbf{z}}$ with the smallest sum of matrix, formulated as:

$$\hat{\mathbf{z}}_{\text{ML}} = \arg \min \{ \sum \mathbf{y}^H \Psi^{-1} \mathbf{y} \}$$

$$= \arg \min \{ \mathbf{y}_{s,d}^* \mathbf{z} \mathbf{C}_{s,d}^{-1} \mathbf{z}^* \mathbf{y}_{s,d} + \mathbf{y}_{r,d}^* \mathbf{z} \mathbf{C}^{-1} \mathbf{z}^* \mathbf{y}_{r,d} \} \quad (3.61)$$

Using Cholesky decomposition approach $\mathbf{C}^{-1} = \mathbf{L}\mathbf{L}^H$,

$$\begin{aligned} \hat{\mathbf{z}}_{\text{ML}} &= \arg \min \left\{ (\mathbf{y}_{s,d} \mathbf{z}^*)^H \mathbf{L}_{s,d} \mathbf{L}_{s,d}^H \mathbf{y}_{s,d} \mathbf{z}^* + (\mathbf{y}_{r,d} \mathbf{z}^*)^H \mathbf{L} \mathbf{L}^H \mathbf{y}_{r,d} \mathbf{z}^* \right\} \\ &= \arg \min \{ |\mathbf{U}_{s,d} \mathbf{z}|^2 + |\mathbf{U}_{r,d} \mathbf{z}|^2 \} \end{aligned} \quad (3.62)$$

where $\mathbf{U}_{r,d} = \{\mathbf{L}_{r,d}^H \mathbf{y}_{r,d}\}^*$ and $\mathbf{U}_{s,d} = \{\mathbf{L}_{s,d}^H \mathbf{y}_{s,d}\}^*$ denotes the upper triangular matrix entry in row and column that further simplify (3.61) as:

$$\hat{\mathbf{z}} = \arg \min \{ |\mathbf{U} \mathbf{z}|^2 \} \quad (3.63)$$

It is noted that the complexity of ML multiple symbol detection grows exponentially with N . In the interest of reducing the exhaustive search of the transmitted signal sequence, SD by Lampe *et. al.* in [30] is proposed to examine only the candidate vectors that are located within the sphere of radius R to obtain $N - 2$ data symbols. The SD adopted algorithm for MSDD scheme is summarized in Algorithm I.

Algorithm 1: MSDDSD

Input: $\sigma_{s,d}^2, \sigma_{s,r}^2, \sigma_{r,d}^2, f_{s,d}, f_{s,r}, f_{r,d}, G, P_S, N_0, N, M, \mathbf{y}$

Output: $\hat{z}[k], \quad k = 1, \dots, M - 1$

1. Compute $\boldsymbol{\varphi}_{h_{s,d}}$ and $\boldsymbol{\varphi}_h$ from (3.56) and (3.59)
2. Compute $\mathbf{C}_{s,d}$ and $\mathbf{C}$ from (3.55) and (3.58)
3. Compute $\mathbf{L}_{r,d}$ and $\mathbf{L}_{s,d}$ from $\mathbf{C}^{-1} = \mathbf{L}\mathbf{L}^H$ and
4. Compute $\mathbf{U}_{r,d} = \{\mathbf{L}_{r,d}^H \mathbf{y}_{r,d}\}^*$ and $\mathbf{U}_{s,d} = \{\mathbf{L}_{s,d}^H \mathbf{y}_{s,d}\}^*$
5. Call function $\hat{\mathbf{z}} = \text{MSDDSD}(\mathbf{U}_{r,d}, \mathbf{U}_{s,d}, N)$ [30]
6. $\hat{z}[k] = \hat{x}^*[k] \hat{x}[k + 1] \hat{x}[k + 2]$

The decision metric is adopted with the SD approach as stated in Algorithm 1 step 5. By referring to SD, the vector search for the ML estimation is limited to a certain radius R spherical region, centered at the received vector. By applying the Schnorr-Euchner search strategy, the radius R is dynamically updated when a candidate vector is found and the process was repeated with new radius R . The process ended when no vector was found within the sphere of radius R .

3.3.4 Performance Analysis of Multiple Symbol Detection

Assume that the data sequence transmitted s is transmitted at the source. The erroneously decoded data instead of the originally transmitted s is represented as $\hat{s}$.

The PEP is defined as the probability of determining the manner when $\mathbf{s} \neq \hat{\mathbf{s}}$ that is expressed as [101]:

$$P_{em}(\mathbf{s} \rightarrow \hat{\mathbf{s}}|\mathbf{h}_m) = Q\left(\frac{d_m(\mathbf{s}, \hat{\mathbf{s}})}{2}\right), \quad m = 1, \dots, N \quad (3.64)$$

where m represents the number of relays (i.e. $m = 1$ corresponds to a single relay and $m = R$ indicates multiple relays) in the system. $d_m(\mathbf{s}, \hat{\mathbf{s}})$ is the Hamming distance between the $\mathbf{s}$ and $\hat{\mathbf{s}}$ at the destination. It is assumed that single relay is used in the system and $Q(x)$ function is given as [102] :

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-\frac{x^2}{2}} dx = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta \quad (3.65)$$

Based on the decision rule, the distance between the s and $\hat{s}$ can be written as:

$$d_1^2(\mathbf{s}, \hat{\mathbf{s}}) = \rho \mathbf{y}^H \text{diag}\{\hat{\mathbf{s}}^* - \mathbf{s}^*\} \mathbf{C}^{-1} \mathbf{y} \text{diag}\{\hat{\mathbf{s}} - \mathbf{s}\} = \rho \mathbf{y}^H \mathbf{Q} \mathbf{y} \quad (3.66)$$

where $\rho = \frac{P_S}{\sigma^2}$ is the instantaneous SNR and $\mathbf{Q} = \text{diag}\{\hat{\mathbf{s}}^* - \mathbf{s}^*\} \mathbf{C}^{-1} \text{diag}\{\hat{\mathbf{s}} - \mathbf{s}\}$

Using the above equations, the average PEP can be calculated by taking the expectation of (3.64).

$$P_{e1}(\mathbf{s} \rightarrow \hat{\mathbf{s}}) = E_{h_1} \left[Q \left(\frac{d_1^2(\mathbf{s}, \hat{\mathbf{s}})}{2} \right) \right] = \frac{1}{\pi} \int_0^\pi E_{h_1} \left[\exp \left(-\rho \frac{\mathbf{y}^H \mathbf{Q} \mathbf{y}}{8 \sin^2 \theta} \right) \right] d\theta \quad (3.67)$$

At high SNR, the average PEP for the single relay is given as:

$$P_{e1}(\mathbf{s} \rightarrow \hat{\mathbf{s}}) = \frac{6}{\det(\text{diag}\{\hat{\mathbf{s}}^* - \mathbf{s}^*\} \text{diag}\{\hat{\mathbf{s}} - \mathbf{s}\})} (\rho^{-2} \ln p) + O\left(\frac{|\ln \det(\text{diag}\{\hat{\mathbf{s}}^* - \mathbf{s}^*\} \text{diag}\{\hat{\mathbf{s}} - \mathbf{s}\})|}{\rho^2}\right) \quad (3.68)$$

According to Divsalar and Simon in [55], upon obtaining the PEP of the multiple symbol sequence detection, the BER over the set of errors is determined by the union bound of the PEP and is derived as:

$$P_b \approx \frac{d}{\log_2(M)(N-1)} P(\mathbf{s} \rightarrow \hat{\mathbf{s}}) \quad (3.69)$$

3.4 Discussion of Simulation Results for Multiple Symbol Based Detection

This section focuses on a cooperative network where channel estimation is not required in the presence of frequency offsets. MATLAB is used as the simulation tool to examine the performance of the proposed ML-DDAF for cooperative diversity networks. BPSK and QPSK signaling are employed, respectively. In the simulation, the channels are affected by the frequency offsets distributed uniformly in $[-\pi, \pi]$ [27, 64] and the fading gains (i.e. the channel variances) for all channels are set to unity (i.e. $\sigma_{sr}^2 = \sigma_{rd}^2 = \sigma_{sd}^2 = 1$). The variance of the Rayleigh fading channels are assumed to be determined by the locations of terminals [103]. Both power for Source P_S and Relay P_R are considered to be equally divided as suggested by Wang and Hanzo in [104]. In all simulation, transmission channels are generated independently according to the simulation approach of [94].

To begin with, a comparison of performance for the coherent scheme in AF relayed system, conventional DD scheme [66] and proposed MSDD scheme, is examined. Figure 3.6 demonstrates the performance comparison of the coherent scheme, the conventional DD scheme ($N_s = 1$) [27], and proposed MSDD detection scheme ($N_s > 1$), in Rayleigh fading environment using BPSK constellation. In this thesis, the observation window size $N_s = 10$ for MSDD scheme is employed.

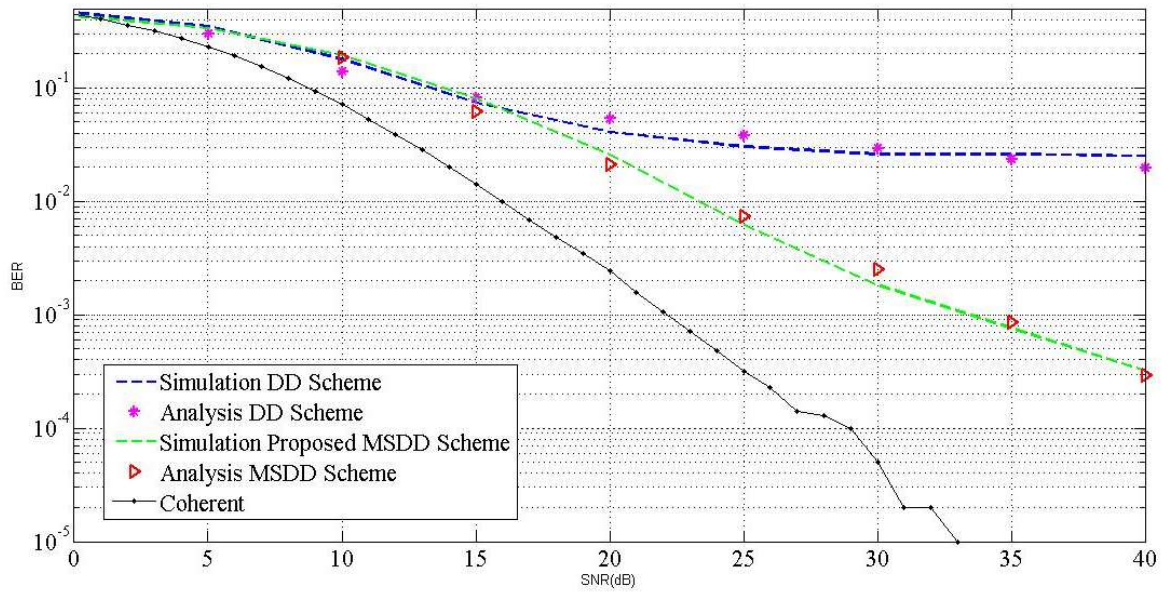


Figure 3.6: Comparison of BER versus SNR performance of MSDD scheme, conventional DD scheme and Coherent scheme when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$; $\sigma_{s,d}^2 = 1$ and $P_S = P_R = 0.5P$ with BPSK constellation.

It is observed in Figure 3.6 that, although the BER curves of coherent with point solid lines and differential detection schemes with dashed lines (different legend) share the same slope, as expected, the coherent detection yields a better performance compare to the DD based scheme. It is witnessed that a performance gap of roughly 5.0 dB incurred when comparing BPSK applying conventional DD detection to BPSK scheme having perfect channel knowledge (coherent) over slow Rayleigh fading environment. The DD detection

scheme is inferior due to lack of accurate channel knowledge. Moreover in the coherent scheme, the detector can exploit additional (i.e. perfect channel knowledge) of a priori parameter(s), while the differential detector cannot. This indicates that the performance of the conventional DD is very sensitive to the accuracy of the required parameter(s) of the channel knowledge.

However in a practical scenario, it is a complicated and difficult task to estimate the channels' properties accurately, particularly in a time-varying channel. Moreover, the coherent scheme does not consider the frequency offsets effect that transpires in a fast fading environment. On that account, the coherent scheme is plotted with the purpose to serve as the benchmark for the cooperative communication system.

Since it is difficult to obtain perfect channel and frequency offsets knowledge in real practice, the DD based transmission allows the detector to omit the requirement of channel knowledge and the frequency offsets estimation, particularly in fast fading environment. The finding is consistent with findings of past studies by [27, 64] that implemented the DD modulation in cooperative communication system to avoid the effect of frequency offsets over the communication channels without estimating the channel gains under slow fading channels.

On the other hand, it can be observed from Figure 3.6 that the performance of the conventional DD scheme when applied in a time-varying channels deteriorates substantially by exhibiting an irreducible error floor at high SNRs [105]. This loss is due to the additional double noise variances that appear in the received signals at the destination and the symbol-based nature of the data signals transmission. As Figure 3.6 suggests, of the two, the additional double noise variances from the received signals appear to be the more significant source of the inefficiency in the protocols. Furthermore, as can be seen from the derivation

of the DD equation, the foundation of this scheme relies upon the assumption that the decision only holds when $h[k] = h[k - 1] = h[k - 2]$. In other words, the decisions are determined based on recent three symbols. Irreducible error floor takes place if the detection scheme fail to comply with the conception of fixed channel coefficients over decision intervals. For example, when the channel varies with respect to time (i.e. fast fading channel), the conventional DD scheme violates the concept of a classical differential system that acquires the channel to be instant over at least three symbol durations. Thus, the scheme fails to perform and results in performance degradation at high SNR regime. This implies that the conventional DD scheme is not favourable in scenarios with the effect of mobility and synchronization error. Hence the aforementioned approach is only valid and presentable for slowly or moderately faded Rayleigh channel (normalized Doppler frequency ($f_D T_s < 0.001$)). On top of that, if the value of f_D grows larger, a DD detector is incapable of compensating for the rapid phase variations of the fading process and therefore, reliable communication is no longer possible.

In order to mitigate the error floor problem, a DD detector based on multiple symbol is proposed to overcomes the losses performed by the conventional DD scheme by forming blocks length of transmitted data signals N_s . Furthermore, the proposed detection scheme is uncorrelated to the aforementioned consideration and as a result is able to perform in relatively fast fading channels. It is presented from the plots with dashed lines that the proposed scheme significantly performs better than the conventional DD symbol based transmission due to a large N_s . It is clearly seen that the MSDDSD scheme with $N_s = 10$ is superior to that with $N_s = 1$ at particularly in the high SNR regime. It is pointed out the performance of MSDD scheme varies with respect to N_s . Hence, the BER expression of MSDD can be employed to obtain the suitable block length at certain fade rate and power.

Next, the corresponding theoretical BER analyses of the CDD and proposed MSDDSD schemes in Section 3.3.2 and 3.3.4, respectively are verified through computer simulation with points (different legends) as depicted in Figure 3.6. It is observed from Figure 3.6 that the analytical BER results based on PDF and PEP match the simulation results. Thus, the conventional DD and MSDDSD theoretical analyses is verified. Note that the PEP subjected to two data sequences with minimum Hamming distance is analyzed and such approach captures the dominant pairwise error. Moreover allocating half of the power transmission to the source terminals helps minimize the BER, especially when SNR is large. Subsequently, it can be observed that the performance of the MSDD cooperative system is dependent on the length of received symbol sequence N_s for data transmission.

On the other hand, the results from Figure 3.7 show the strong impact between N_s and the performance enhancement.

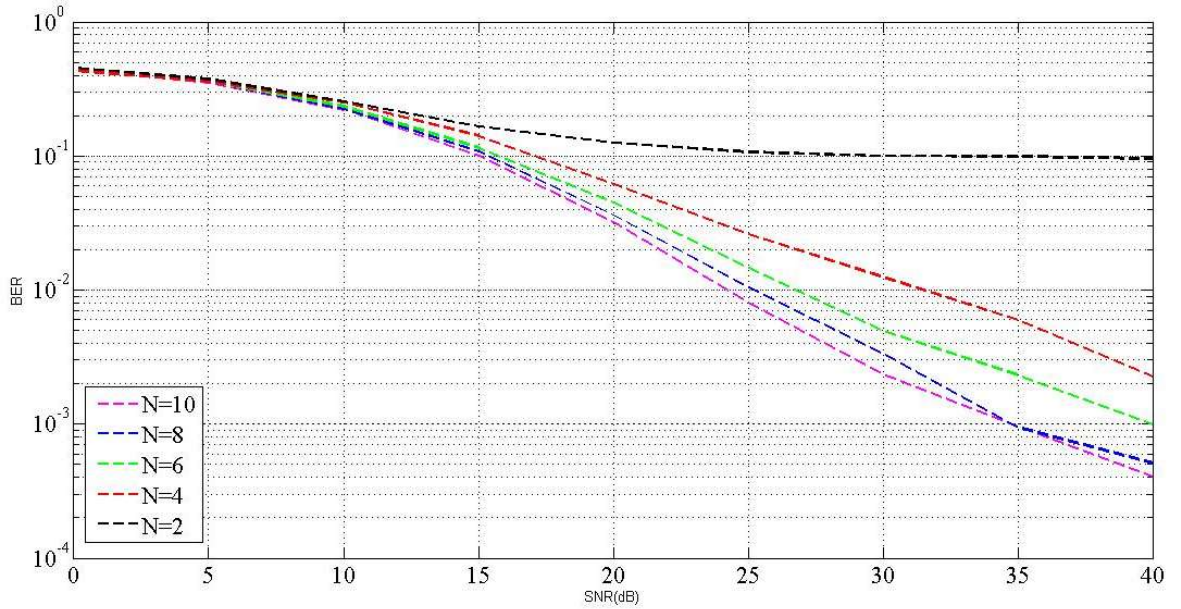


Figure 3.7: Performance of BER for the proposed MSDD scheme with different length received symbol sequence N when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 10$; $\sigma_{s,d}^2 = 1$ and $P_S = P_R = 0.5P$

It can be shown from the trend of the curve that, much of the MSDD performance significantly improves when N_s increases. As seen from the plot, the performance increases significantly by increasing the block length N_s from 2 to 4. However, the BER slowly decreases with increasing N . The curve demonstrated specifically, that at BER of 10^{-3} , the performance improves at about 5 dB as the symbol observation length increases from $N_s = 6$ to $N_s = 8$. It is also observed that upon increasing N_s beyond 8 will produce negligible improvement. Beyond this point, the performance show slight increment and thus, the performance improvement becomes not significant with increasing N_s . Moreover, the likelihood of making correct detection for the received DDPSK symbol sequence at the relay terminals becomes smaller if larger block length is implemented in the source-relay link. Thus, from the result observation of Figure 3.7, the performance of MSDDSD can be determined by the number of observed symbols. It is suggested that the length of the data signal transmission block is kept small to reduce the constrained by this undesirable property in practice. The small value of $N_s = 10$ can be implemented to reduce complexity, albeit to some performance degradation consequence. In addition, the enhancement of the error probability performance indicates that it is more favourable to increase the length of the symbol sequence at the destination than to exploit perfect statistical knowledge of channel and frequency offsets.

3.4.1 Power Allocation

Next, the equal distributed power and power allocation allocation for the conventional DD and proposed scheme is compared. The power is often equally divided and allocated between the source and relay in order to simplify the performance analysis of the DD based cooperative diversity system. It turns out that an equal power strategy produces

good results, yet in general not optimal in the context of cooperative communications. Thus, according to [106] it is required to obtain the optimisation for the allocation of power between the source and relay in highly unbalanced links (i.e. dissimilar channel link quality) to enhance the system performances. It can be observed from (3.37) that the BER is subjected to P_S and P_R . Hence, the BER performance of the proposed MSDD scheme can be further revised by applying a more efficient use of distributed power resources among the source and relay terminals. P_S and P_R can be obtained by minimizing the BER formulation in (3.37) using the power constraint, $P_T = P_S + P_R$ for the system. Figure 3.8 depicts the power allocation factor for different channel variances scenarios with the error probability performance versus the PAF $p = P_S/P_T$.

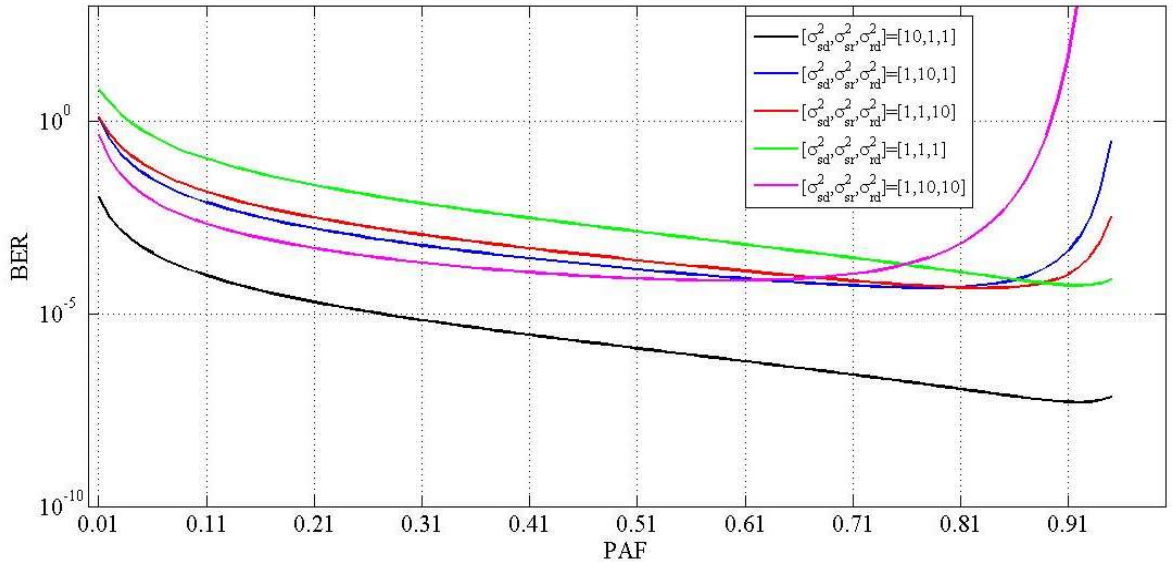


Figure 3.8: Power Allocation Factor (PAF) at $\frac{P}{N_0} = 20dB$ for different channel variances scenarios

The obtained optimum power allocation for $SNR = 20dB$ corresponding to different channel variances (channel link quality) is further tabulated in Table 3.3.

Table 3.3 : Power allocation for various channel variances

Scenarios	Channel Variances			Power Allocation	
	SD	SR	RD	P_S	P_R
Symmetric	1	1	1	0.92	0.08
Strong SD	10	1	1	0.92	0.08
Strong SR	1	10	1	0.78	0.22
Strong RD	1	1	10	0.83	0.17
Strong SR and RD	1	10	10	0.59	0.41

An interesting result observed is that the optimized power allocation is uncorrelated with the point-to-point link. The power allocation factor is subjected only to the relayed channel. Specifically, the simulation results from Figure 3.8 and Table 3.3 indicating that the optimum transmitted power P_S at the source over the total power P_T is in between 0.5 and 1. On the contrary, the optimum ratio of the power P_R used at the relay over the total power P_T is larger than 0 and less than 0.5 (i.e $0.5 < \frac{P_S}{P_T} < 1$ and $0 < \frac{P_R}{P_T} < 1$), respectively. It is pointed out that more power should always be allocated at the source and lesser power at the relay. If the source-relay channel quality is much less than that of relay-destination channel, then P_S approaches P_T while P_R approaches 0. Since the system is a case of diversity reception, the system adds more power to the source to drive both the first hop and direct link. In contrast, if the channel quality between source-relay and relay-destination is much larger than the relay-destination, the system will perform better with both P_S and P_R approaches $\frac{P_T}{2}$. It means that symmetrical power is distributed to the source and relay.

Also, it is observed from Figure 3.9 that the optimized power allocation (i.e. $P_S:P_R = 0.92:0.18$) achieve better results as compare to the case when the power is equally divided between the source and relay nodes when the channels are symmetry. For instance, at SNR = 20dB, the simulated BER of the equally distributed power equal to 2×10^{-2} , while the BER from the power allocation analytical approach equal to 2.8×10^{-2} . It can be seen that the improved resource (power) allocation potentially leads to better coverage areas and extended terminal lifetime. Therefore, it can be concluded that the optimum power allocation depends on the quality of the communication channel, especially when relay is involved.

Since all channel links $h_{s,d}$, $h_{s,r}$ and $h_{r,d}$ are considered available in the system, the cooperation approach is likely to achieve second order of diversity. It is certain that the system is producing a performance diversity of order one because of the point-to-point (i.e. source-destination) link. Nevertheless, to obtain cooperative diversity order of two, the source-relay and relay-destination channels ought to be appropriately balanced. For instance, if the source-relay channel link exhibit poor quality, then it is extremely hard for the relay to relayed the transmitted symbol. Thus, the relay channel becomes less significant and it makes sense to allocate more power at the source. On the contrary, if the source-relay quality link is very good, the relay can always forward the transmitted symbol effectively (i.e. the amplified symbol at the relay is almost similar as the source). The relay can be considered as a replica of the source and an equally distributed power can be allocated on them. However, it is highlighted that this interpretation is good only for sufficiently high SNR scenario and under the assumption that all of the channel links $h_{s,d}$, $h_{s,r}$ and $h_{r,d}$ are available.

Figure 3.9 potrays the AF cooperative system performance employing BPSK signals for optimum power allocation versus equal power scheme under different channel quality

condition as shown in Table 3.4. In case of symmetrical channel, Case I, the optimum power ratio are $\frac{P_S}{P_T} = 0.92$ and $\frac{P_R}{P_T} = 0.08$. It is depicted in the figure that the optimum power allocation results is slightly better than the equally distributed power. Similar trend of curves are shown for Case II and Case III whereby the optimum power allocation scheme outperforms the equal power scheme with performance enhancement of about 1 dB. Also, it is noted that if the ratio of the link quality ratio $\sigma_{s,r}^2/\sigma_{r,d}^2$ or $\sigma_{r,d}^2/\sigma_{s,r}^2$ becomes larger, more performance improvement of the optimum power allocation over the equal power can be observed.

Table 3.4: Various channel qualities based on channel variances

Case	Channel Variances $[\sigma_{s,d}^2; \sigma_{s,r}^2; \sigma_{r,d}^2]$	Channel status
I	1;1;1	Symmetric
II	1;10;1	Strong SR
III	1;1;10	Strong RD

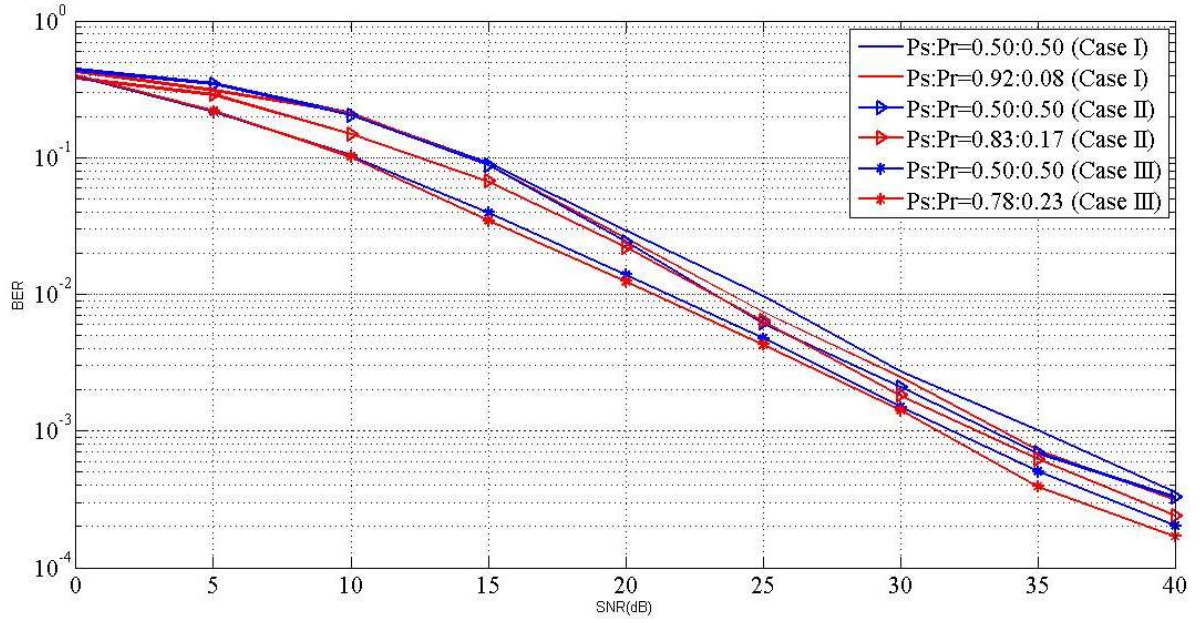


Figure 3.9: BER performance comparison of the proposed MSDD scheme with symmetric power allocation and optimized power allocation under various channel quality link

It is worth noting that the power allocation optimization which is devised from minimizing the BER approximation (3.37), concides with the result presented by [106], in which the power allocation optimization was obtained from a minimizing outage probability point of view for multi-hop communication systems.

3.4.2 Multiple Symbol Double Differential Scheme Under Various Mobility Effect

The proposed MSDD scheme is further concentrated and simulated in various channel qualities based on the channel variances as symmetric channels with $\sigma_{sr}^2 = 1; \sigma_{rd}^2 = 1; \sigma_{sd}^2 = 1$, nonsymmetric channels with $\sigma_{sr}^2 = 1; \sigma_{rd}^2 = 1; \sigma_{sd}^2 = 10$ which indicates strong source-destination (SD) link and non-symmetric channels with $\sigma_{sr}^2 = 1; \sigma_{rd}^2 = 10; \sigma_{sd}^2 = 1$ which represents strong source-relay (SR) link.

Additionally, referred to the normalized Doppler frequencies of the channels, three different channel fading cases as tabulated in Table 3.5 are considered – (I) all the channels exhibit slow fading; (II) the source-destination (SD) channel is fast fading while the source-relay (SR) and relay-destination (RD) channels are slow fading; (III) all the channels are fast fading.

Table 3.5: Simulation case based on normalized Doppler frequency

Case	Normalized Doppler frequency $(f_{D_{s,d}}T_s, f_{D_{s,r}}T_s, f_{D_{r,d}}T_s)$	Channel condition
I	0.001, 0.001, 0.001	SD, SR, RD slow fading
II	0.01, 0.01, 0.001	SD fast fading; SR, RD slow fading
III	0.01, 0.01, 0.01	SD, SR, RD fast fading

The normalized Doppler frequency values can be interpreted to the speed of the mobile users or various vehicle speed of communication terminals in a typical wireless system. Refer to the simulated values of optimal power allocation in Table 3.3 and various cases of channel conditions (i.e. slow fading and fast fading) in Table 3.5, the BER performance of the conventional DD and proposed MSDD are simulated. The simulations are plotted in Figure 3.10, Figure 3.11 and Figure 3.12, respectively for the symmetry and non-symmetry cases (i.e. different channel variances) for DD BPSK. From Figure 3.10, it can be observed that for case I, the BER is decreasing with SNR and the error floor does not take place. However for case II and case III of conventional symbol-by-symbol based DD scheme, which involve fast fading channels, error floor occurs at high SNR. The irreducible

error floor is formed due to the increment of the frequency offsets caused by the mobility of the communication terminals. In order to solve the error floor problem, MSDD scheme employing SD is applied. It is observed that the proposed MSDDSD increases the BER performance under time-varying channels. Figure 3.10 also compares the simulation results based on 10^6 bits information data with the numerical analysis results in Section 3.3.4. It is demonstrated that the simulation results match closely with the numerical analysis results.

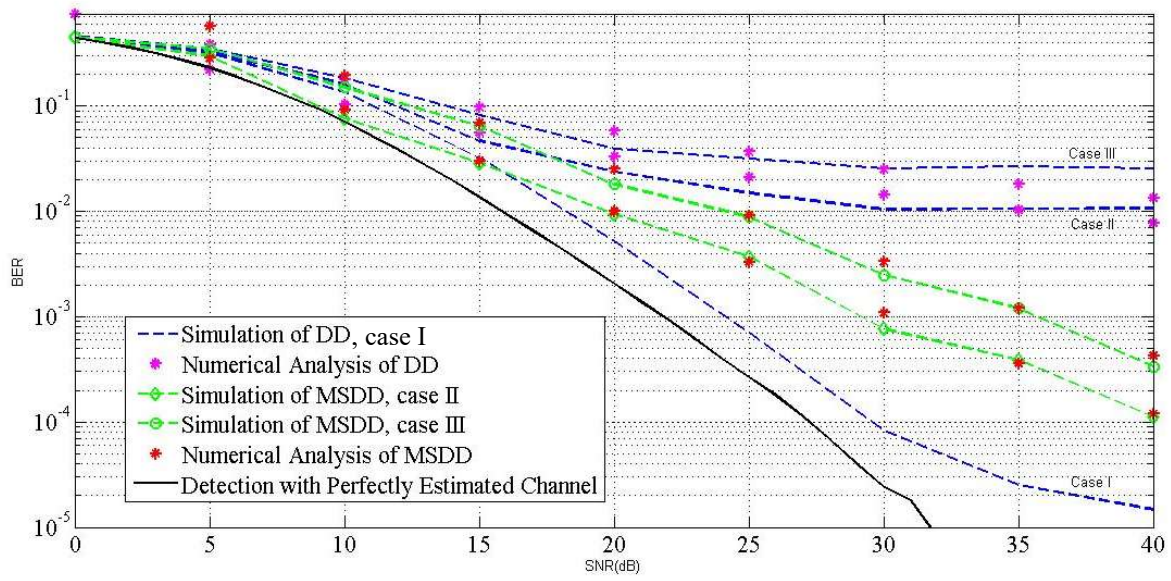


Figure 3.10: Comparison of the BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$; $\sigma_{s,d}^2 = 1$ for DDBPSK

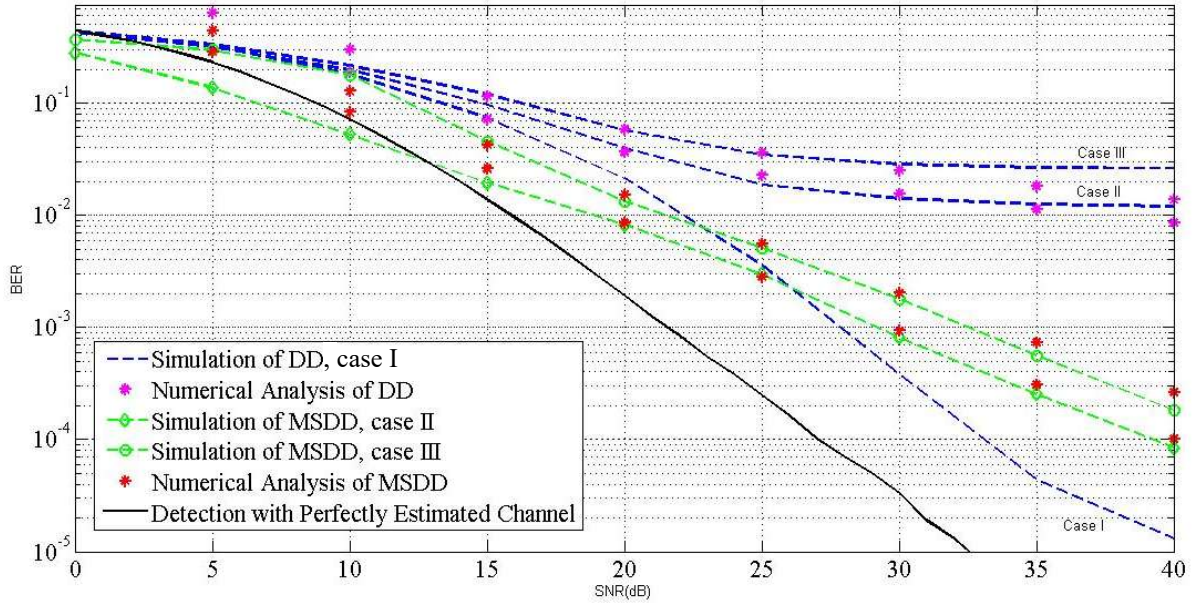


Figure 3.11: Comparison of the BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$; $\sigma_{s,d}^2 = 10$ for DDBPSK

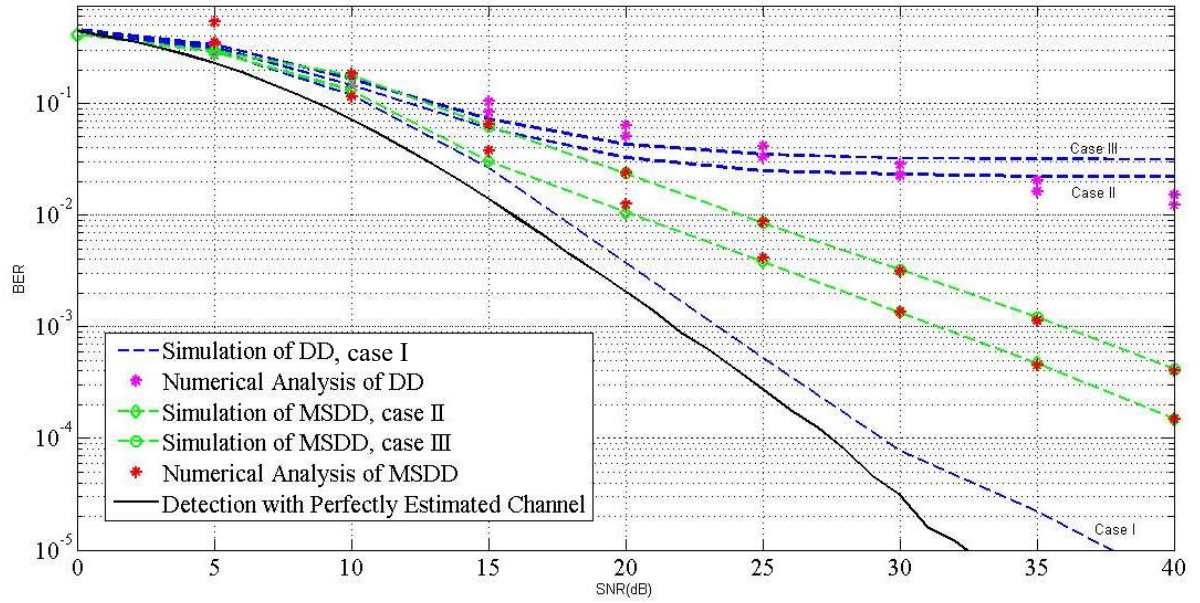


Figure 3.12: Comparison of the BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 10$; $\sigma_{s,d}^2 = 1$ for DDBPSK

Additionally, it is also observed that the performance trend in the non-symmetry case of $\sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1; \sigma_{s,d}^2 = 10$ and $\sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 10; \sigma_{s,d}^2 = 1$ as shown in Figure 3.11 and Figure 3.12 are similar to the symmetry case in Figure 3.10.

On top of that, the error performance versus the system SNR for perfectly estimated channel detection, conventional DD and proposed MSDD scheme is further analysed utilizing QPSK modulation. The simulation results for symmetrical case are depicted in Figure 3.13. The simulation setup is similar to Figure 3.10. From Figure 3.13, it can be observed that the curves exhibit similar trend when different modulation level ($M = 4$) is utilized. The results are further analysed for non-symmetrical case as demonstrated in Figure 3.14 and Figure 3.15.

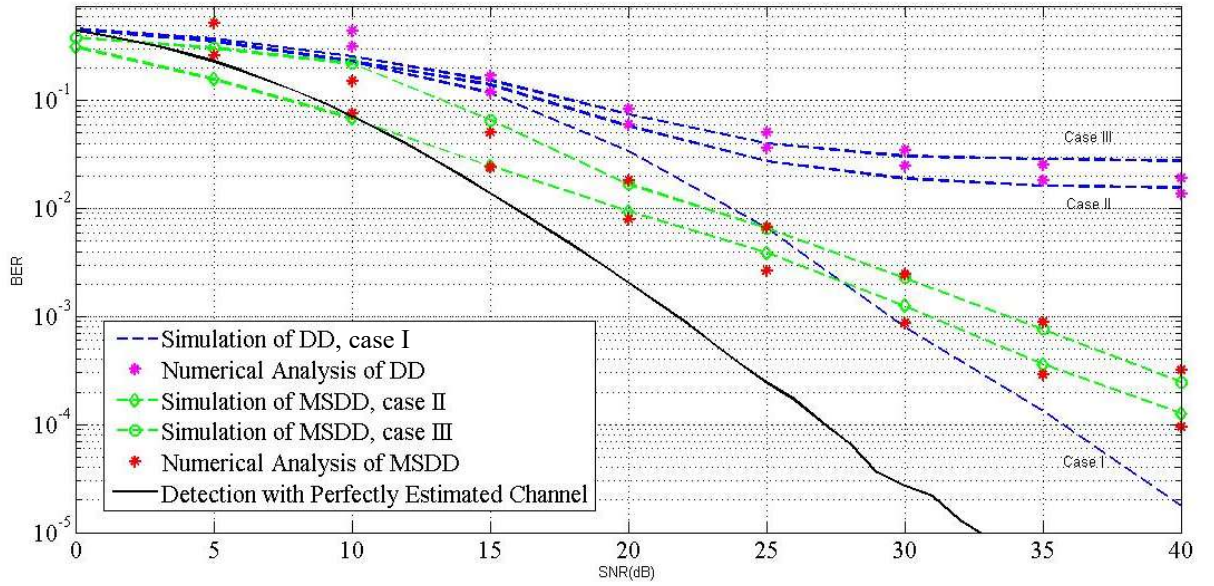


Figure 3.13: Comparison of the BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1; \sigma_{s,d}^2 = 1$ for DDQPSK

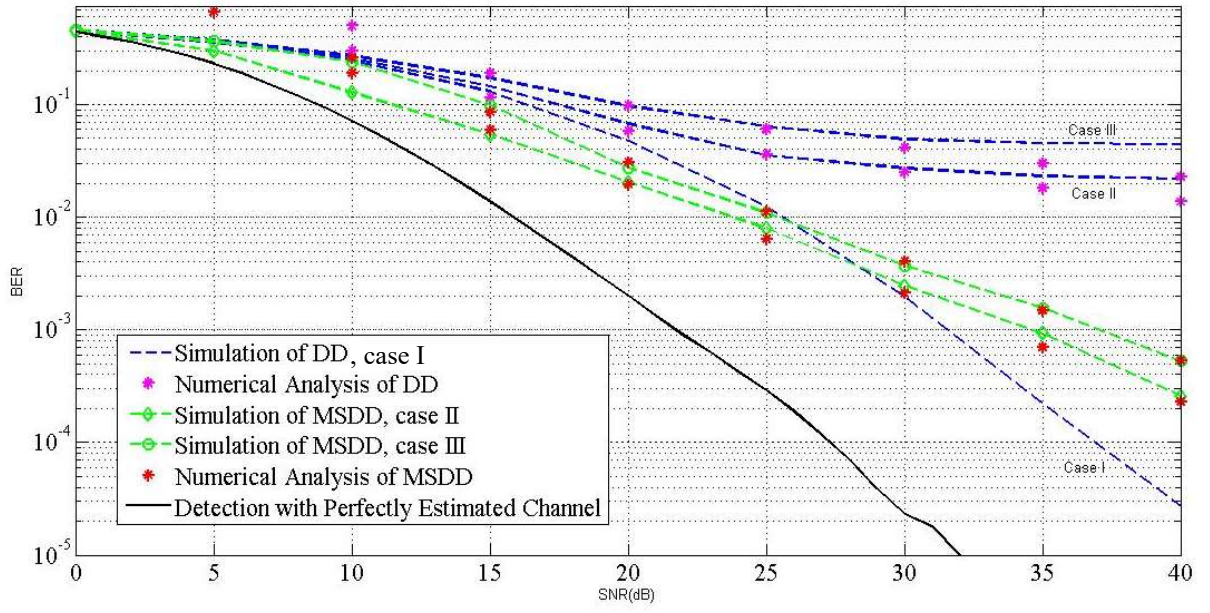


Figure 3.14: Comparison of the BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$; $\sigma_{s,d}^2 = 10$ for DDQPSK

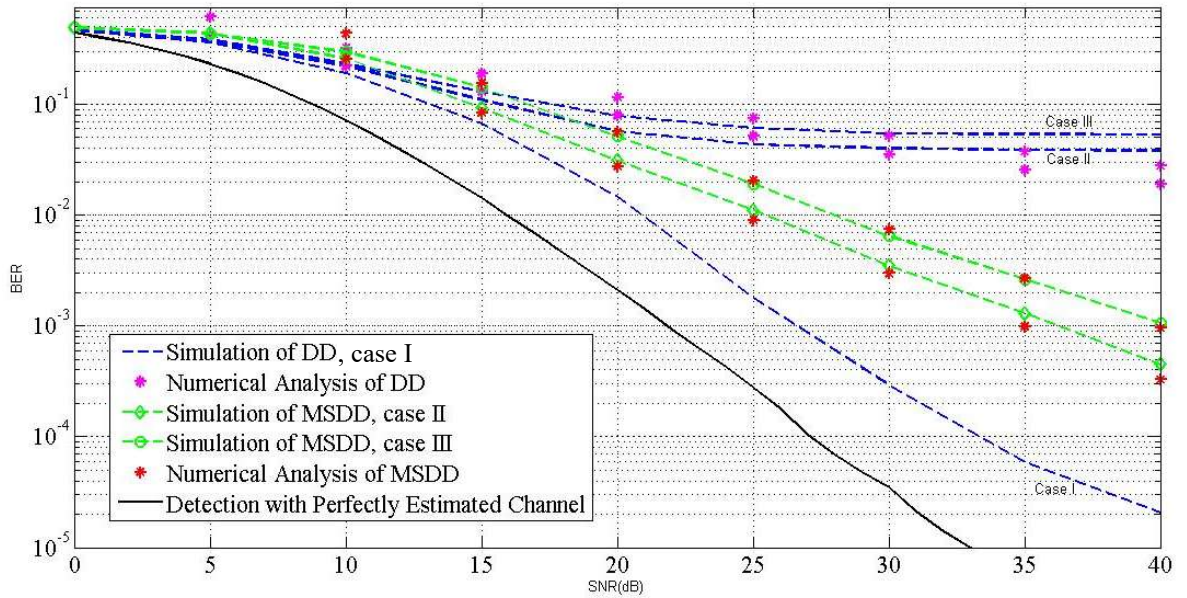


Figure 3.15: Comparison of the BER performance for the perfectly estimated channel detection, conventional DD and proposed MSDD scheme with optimized power allocation when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 10$; $\sigma_{s,d}^2 = 1$ for DDQPSK

For the BER performance in Case II, the BER curve gradually diverge from the results acquired in Case I. The curve for the conventional DD and proposed MSDD scheme eventually reaches the error floor as the Doppler frequency increases. Even though the existence of the error floor is unavoidable in the schemes, the simulation results significantly show that the proposed MSDD scheme exhibit lower error floor over the conventional DD scheme. In Case III, the performance degrades severely as compared to the performance in Case II due to the condition wherein all channels are fast fading. The results clearly portray that the proposed scheme is capable to narrow down the performance gap and increases the performance of the system over fast fading channel owing to the time diversity provided by the time-varying channel.

In case of QPSK modulation, it can be observed that the larger the signal constellation size, the less performance gain is achieved compare to the case of BPSK modulation. This is because as the Doppler frequency increases, the channel becomes time-selective. Despite the advantage of time-selective channel that provides time diversity, noise enhancement and frequency offsets effect also grow for higher order modulations. It is noted that the performance degradation from the noise and frequency offsets is more dominant from the time-selective channel over higher order modulations.

As the results in this section indicate, for AF; the error probability performance minimizes as SNR increases. The proposed multiple symbol based DD transmission employing SD is compared to the conventional DD symbol based and verified with the theoretical numerical analysis. The impacts of the transmission block lengths and power allocation are examined in terms of BER. Simulation results show that the proposed MSDDSD is effective in removing the frequency offsets in a time-varying channel environment.

3.5 Summary

In this chapter, differential cooperative communication system employing AF relaying protocol has been studied and analysed over the Rayleigh time-varying channels. This chapter described the system and channel model subjected in this thesis. A time-series model was developed for the communication channels wherein the time-varying channels were related to the values of auto-correlation based on the AR(2) approach. It was showed that the fading rates from the auto-correlation values affected the network error performance. Specifically under fast fading channels, the large fading rates of the channels lead to system poor performance. It was observed that an error floor occurred in the conventional two symbol DD transmission, particularly in the fast fading channel. The analysis was supported with simulation in various time-varying (i.e. slow fading, fairly fast fading and fast fading) scenarios. This analysis is beneficial for the system to predict the error floor phenomenon and control its value by varying the parameters such as the symbol interval time. Due to the limitation of the conventional DD especially in fast fading channels, MSDDSD is proposed. It is shown that the proposed scheme is able to perform significantly better over that achieved with the conventional DD transmission although the improvement is achieved at the cost of higher complexity. In the next chapter, various combining schemes will be studied since the proposed scheme in this chapter applied direct combining scheme at the receiver.

CHAPTER 4

DIVERSITY COMBINING SCHEMES

4.1 Introduction

In the previous Chapter 3, two-symbol and MSDDSD scheme is discussed to improve the performance of an AF MSDD transmission with direct combiner. This chapter focuses on the diversity combining schemes for MSDDSD detection scheme proposed in the previous chapter. Four diversity combiners, including direct combiner, MRC, semi MRC and SC are demonstrated and explained in details. In the previous section, MSDDSD is proposed wherein, direct combiner is implemented at the desired destination. Based on the proposed system model and multiple symbol based DD detection in Section 3.2 and Section 3.3 respectively, one step further is taken at the destination to improve the network's performance. The previous MSDDSD employing direct combiner is extended with other diversity combining techniques such as, conventional MRC, semi MRC and SC under time-varying channels. SC is selected among the other diversity combining schemes due to its approach that does not depends on the channel knowledge. A comprehensive error probability and outage probability performance analysis are carried through the flat fading Rayleigh environment for semi MRC and SC. Specifically, PEP expressions for the semi MRC detectors are derived based on the MGF. Contrarily, PDF analysis expression is derived to measure the system outage probability. For SC detectors, the error performance analysis is obtained using the PDF. Finally, power allocation expression based on error performance minimization approach is presented for the proposed SC performance optimization.

The chapter organization is presented as follows. Section 4.2 briefly introduced the proposed combining schemes in this thesis. Section 4.3 highlights the methods to obtain the proposed MRC and followed by the numerical analysis derivation in Section 4.4. The simulation and numerical analysis results are discussed in Section 4.5. The second proposed diversity combining scheme proposed, diversity SC method is highlighted in Section 4.6 while its numerical analysis is derived in detail in Section 4.7. Section 4.8 shows that the analysis of simulations were further validated with the new performance analysis for the proposed schemes. Finally, Section 4.9 summarizes the important conclusions deduced during the analysis.

4.2 Proposed Combining Schemes

Consider the wireless AF relaying system and channel model as detailed in Section 3.2 and its sub-sections, respectively. A group of fixed gain based on the second-order statistics of all wireless transmission channels are studied without estimating the channel and frequency offsets knowledge at the desired destination. This combiner is referred to as MRC. As the MRC performance analysis is challenging, instantaneous combining weights are taken into account to validate the performance of the system.

However, obtaining the second-order statistics of all channels at the destination becomes an issue due to the fast fading channels, a simpler combiner named SC is proposed. In the implementation of SC, decision variable is computed for each link and only the largest magnitude value is selected at the detector. This in fact simplify the detection process at the destination because it does not require the second-order statistics of all the channels.

4.3 Maximal Ratio Combiner

Based on AR(2), using the individual and relayed channels from Section 3.2.2 and substitutes equation (3.15) and (3.19) into (3.2) and (3.4), accordingly, gives:

$$\begin{aligned}
 y_{s,d}[k] &= P_S[\alpha_{s,d}^2 y_{s,d}[k-1] y_{s,d}[k-2] \\
 &+ \alpha_{s,d} x[k] y_{s,d}[k-1]] e^{j2\pi f_{s,d}[k]} + n_{s,d}[k] \\
 n_{s,d}[k] &= w_{s,d}[k] - \alpha_{s,d}^2 y[k-1] w_{s,d}[k-1] - \alpha_{s,d} x[k] w_{s,d}[k-2] \\
 &+ 2\sqrt{1 - \alpha_{s,d}^2} \sqrt{P_S} z[k] \varepsilon_{s,d}[k]
 \end{aligned} \tag{4.1}$$

and

$$\begin{aligned}
 y_{s,r_i,d}[k] &= G[\alpha_{s,r_i,d}^2 y[k-1] y_{s,r_i,d}[k-2] + \\
 &\alpha_{s,r_i,d} x[k] y_{s,r_i,d}[k-1]] e^{j2\pi f_{s,r_i,d}} + n_{s,r_i,d}[k]
 \end{aligned} \tag{4.2}$$

where

$$\begin{aligned}
 n_{s,r_i,d}[k] &= w_{s,r_i,d}[k] - \alpha_{s,r_i,d}^2 y[k-1] w_{s,r_i,d}[k-1] \\
 &- \alpha_i x[k] w_{s,r_i,d}[k-2] + 2G\sqrt{1 - \alpha_i^2} \sqrt{P_S} z[k] h_{r_i,d}[k] \varepsilon_{s,r_i}[k]
 \end{aligned} \tag{4.3}$$

Since the equivalent noise $n_{s,d}[k]$ and $n_i[k]$ are complex Gaussian random variables, thus their average noise variances are also complex Gaussian. The average noise variances employing DD transmission are written as:

$$\sigma_{n_{s,d}}^2 = 1 + \alpha_{s,d}^4 + \alpha_{s,d}^2 + 2P_S(1 - \alpha_{s,d}^2) \tag{4.4}$$

$$\sigma_{n_{s,r_i,d}}^2 = \sigma_{s,r_i,d}^2 \left(1 + \alpha_{s,r_i,d}^4 + \alpha_{s,r_i,d}^2 + 2\rho_{s,r_i,d}(1 - \alpha_{s,r_i,d}^2) \right) \tag{4.5}$$

The received signals are combined at the destination based on MRC [40]:

$$\begin{aligned}
 \varsigma &= \beta_0 (y_{s,d}[n] y_{s,d}^*[n-1]) (y_{s,d}[n-1] y_{s,d}^*[n-2])^* \\
 &+ \beta_1 (y_{s,r_i,d}[n] y_{s,r_i,d}^*[n-1]) (y_{s,r_i,d}[n-1] y_{s,r_i,d}^*[n-2])^*
 \end{aligned} \tag{4.6}$$

where

$$\beta_0 = \frac{\alpha_{s,d}}{\sigma_{n_{s,d}}^2} \quad (4.7)$$

$$\beta_1 = \frac{\alpha_{r,d}}{\sigma_{n_{r,d}}^2} \quad (4.8)$$

$\alpha_{s,d}$ and $\alpha_{r,d}$ are the autocorrelation for source-destination and relay-destination, respectively. For slow fading channels, as the autocorrelation value is 1. It can be seen from (4.3) that the noise variance is dependent on the channel coefficient $h_{r,d}[k]$. Since channel coefficient is considered to be unknown in the system, the mean of the noise variance is used as the combining weight for the conventional DD transmission scheme. Thus, the weight for the slow fading channels can be written as:

$$\beta_{0(slow)} = \frac{1}{3} \quad (4.10)$$

$$\beta_{1(slow)} = \frac{1}{3(1 + G_i^2)}, \quad i = 1, \dots, R \quad (4.11)$$

For fast fading channels, the mean value of the noise variances can be calculated as:

$$E\{\sigma_{n_{s,d}}^2\} = 1 + \alpha_{s,d}^4 + \alpha_{s,d}^2 + 2(1 - \alpha_{s,d}^2)P_S \quad (4.12)$$

and

$$E\{\sigma_{n_{s,r_i,d}}^2\} = (1 + G_i^2)(1 + \alpha_{s,r_i,d}^4 + \alpha_{s,r_i,d}^2) + 2(1 - \alpha_{s,r_i,d}^2)G_i^2 P_S \quad (4.13)$$

Thus, the proposed and modified weight factors for the time-varying channels are expressed as:

$$\beta_{0(fast)} = \frac{\alpha_{s,d}}{1 + \alpha_{s,d}^4 + \alpha_{s,d}^2 + 2(1 - \alpha_{s,d}^2)P_S} \quad (4.14)$$

and

$$\beta_{1(fast)} = \frac{\alpha_{s,r_i,d}}{(1 + G_i^2)(1 + \alpha_{s,r_i,d}^4 + \alpha_{s,r_i,d}^2) + 2(1 - \alpha_{s,r_i,d}^2)G_i^2 P_S} \quad (4.15)$$

For the fast fading environment, it can be observed that the proposed weight factor is dependent on the auto-correlation of the channel, amplification gain of the relay(s) and the average source power. The proposed weight that depends on the fade rate of the channels, provides dynamic weight to the received signals. The faster the channels varies in a communication links, the received signals are attenuated and thus, only a small amount of signal magnitude is taken into account for the detection process. That is, the channels with strong signals are further amplified.

4.4 Performance Analysis of Maximal Ratio Combiner

In this subsection, the focus is shifted to the numerical evaluation of the proposed scheme subjected to error probability and outage probability performance. In order to analyse and resolve the theoretical error probability performance, MGF method is employed while PDF based approach is employed for the outage probability analysis. MGF method is chosen because it offers a unified-based solution to measure the PEP or error probability of a specific system communicating over diverse fading channels such as Rayleigh fading. This subsection commences from the introduction of mathematical preliminaries, and then focuses on the error probability analysis of the average end-to-end error probability and outage probability performance based on a PEP and PDF approach, respectively.

4.4.1 Error Probability Performance Analysis

This section evaluates the performance of the MSDD AF network employing MRC under time-varying fading channels. Suppose that x_1 is sent, and x_2 is decoded erroneously by the decoder. According to Zhang in [107] an erroneous event occurs when:

$$||\zeta - x_1||^2 > ||\zeta - x_2||^2 \quad (4.16)$$

which can be simplified to:

$$||n|| > ||a|| \quad (4.17)$$

whereby n denotes the noise terms and a represents the signal received with $\varsigma(x_1 - x_2)$ and $x_1 - x_2 = d_{min}$. By substituting ς from (4.7) into (4.17) and employing the optimal combining weight β_0 and β_1 , the erroneous event can be further written as:

$$\begin{aligned} \xi \triangleq & |d_{min}|^2 (\alpha_{s,d} \beta_0 |y_{s,d}[k-1]|^2 |y_{s,d}[k-2]|^2 + \\ & \alpha_{s,r_i,d} \sum_{i=1}^R \beta_{1i} |y_{s,r_i,d}^*[k-1]|^2 |y_{s,r_i,d}^*[k-2]|^2) + \\ & d_{min}^* (\beta_0 y_{s,d}^*[k-1] y_{s,d}^*[k-2] n_{s,d}[k] \\ & + \sum_{i=1}^R \beta_{1i} y_{s,r_i,d}^*[k-1] y_{s,r_i,d}^*[k-2] n_{s,r_i,d}[k]) < 0 \end{aligned} \quad (4.18)$$

Conditioned on $y_{s,d}$, $y_{s,r_i,d}$ and $h_{r_i,d}$, the last 2 terms of the decision parameter ξ constitute of Gaussian noise. It is also noted that the 2 noise terms are independent and mutually uncorrelated with each other because they are complex conjugate.

The mean of the noise component n can be expressed as:

$$\begin{aligned} \mu_n = & E\{n|y_{s,d}\{k-1\}y_{s,d}[k-2], y_{s,r_i,d}[k-1]y_{s,r_i,d}[k-2], \\ & h_{r_i,d}[k-1]h_{r_i,d}[k-2]\} \\ = & d_{min}^* (\beta_0 y_{s,d}^*[k-1] y_{s,d}^*[k-2] E\{n_{s,d}[k]|y_{s,d}[k-1]y_{s,d}[k-2]\} \\ & + \sum_{i=1}^R \beta_{1i} y_{s,r_i,d}^*[k-1] y_{s,r_i,d}^*[k-2] \\ & \times E\{n_{s,r_i,d}[k]|y_{s,r_i,d}[k-1]y_{s,r_i,d}[k-2], h_{r_i,d}[k-1]h_{r_i,d}[k-2]\}) \end{aligned} \quad (4.19)$$

It can be seen that since $n_{s,d}[k]$ and $n_{s,r_i,d}[k]$ are conditioned on $y_{s,d}[k-1]y_{s,d}[k-2]$ and $y_{s,r_i,d}[k-1]y_{s,r_i,d}[k-2]$ respectively, both terms can be represented as $w_{s,d}[k-1]w_{s,d}[k-2]$ and $w_{s,r_i,d}[k-1]w_{s,r_i,d}[k-2]$. Thus, (4.19) becomes:

$$\begin{aligned}
\mu_n &= d_{min}^* (\beta_0 y_{s,d}^*[k-1] y_{s,d}^* \\
&[k-2] E\{w_{s,d}[k-1] w_{s,d}[k-2] | y_{s,d}[k-1] y_{s,d}[k-2]\} \\
&+ \sum_{i=1}^R \beta_{1_i} y_{s,r_i,d}^*[k-1] y_{s,r_i,d}^*[k-2] \times
\end{aligned} \tag{4.20}$$

$$E\{w_{s,r_i,d}[k-1] w_{s,r_i,d}[k-2] | y_{s,r_i,d}[k-1] y_{s,r_i,d}[k-2], h_{r_i,d}[k-1] h_{r_i,d}[k-2]\}$$

The conditional means of Gaussian noise $w_{s,d}[k-1] w_{s,d}[k-2]$ and $w_{s,r_i,d}[k-1] w_{s,r_i,d}[k-2]$ can be computed as:

$$\begin{aligned}
&E\{w_{s,d}[k-1] w_{s,d}[k-2] | y_{s,d}[k-1] y_{s,d}[k-2]\} \\
&= \frac{1}{2P_S + 1} d_{min} y_{s,d}[k-1] y_{s,d}[k-2]
\end{aligned} \tag{4.21}$$

and

$$\begin{aligned}
&E\{w_{s,r_i,d}[k-1] w_{s,r_i,d}[k-2] | y_{s,r_i,d}[k-1] y_{s,r_i,d}[k-2], h_{r_i,d}[k-1] h_{r_i,d}[k-2]\} \\
&= \frac{1}{2\rho_{s,r_i,d}} d_{min} y_{s,r_i,d}[k-1] y_{s,r_i,d}[k-2]
\end{aligned} \tag{4.22}$$

Substitutes (4.21) and (4.22) into (4.20) gives:

$$\begin{aligned}
\mu_n &= |d_{min}|^2 \left(\frac{\alpha_{s,d} \beta_0}{2P_S + 1} |y_{s,d}[k-1] y_{s,d}[k-2]|^2 \right. \\
&\quad \left. + \sum_{i=1}^R \frac{\alpha_{s,r_i,d} \beta_{1_i}}{2\rho_{s,r_i,d}} |y_{s,r_i,d}[k-1] y_{s,r_i,d}[k-2]|^2 \right)
\end{aligned} \tag{4.23}$$

Then, the variance of the complex Gaussian noise is computed as:

$$\begin{aligned}
\sigma_n^2 &= Var\{n | y_{s,d}[k-1] y_{s,d}[k-2], y_{s,r_i,d}[k-1] y_{s,r_i,d}[k-2], \\
&\quad h_{r_i,d}[k-1] h_{r_i,d}[k-2]\} \\
&= d_{min}^2 (\beta_0^2 |y_{s,d}[k-1] y_{s,d}[k-2]|^2 Var\{n_{s,d}[k] | y_{s,d}[k-1] y_{s,d}[k-2]\} +
\end{aligned} \tag{4.24}$$

$$\sum_{i=1}^R (\beta_{1_i})^2 |y_{s,r_{i,d}}[k-1] y_{s,r_{i,d}}[k-2]|^2$$

$$\times \text{Var}\{n_{s,r_{i,d}}[k] | y_{s,r_{i,d}}[k-1] y_{s,r_{i,d}}[k-2], h_{r_{i,d}}[k-1] h_{r_{i,d}}[k-2]\})$$

where

$$\text{Var}\{n_{s,d}[k] | y_{s,d}[k-1] y_{s,d}[k-2]\} = 1 + \alpha_{s,d}^4 + \alpha_{s,d}^2 + 2P_S(1 - \alpha_{s,d}^2) = \frac{\alpha_{s,d}}{b_0} \quad (4.25)$$

and

$$\text{Var}\{n_{s,r_{i,d}}[k] | y_{s,r_{i,d}}[k-1] y_{s,r_{i,d}}[k-2], h_{r_{i,d}}[k-1] h_{r_{i,d}}[k-2]\})$$

$$= 1 + \alpha_{s,r_{i,d}}^4 + \alpha_{s,r_{i,d}}^2 + \rho_{s,r_{i,d}}(1 - \alpha_{s,r_{i,d}}^2) = \frac{\alpha_{s,r_{i,d}}}{b_{1_i}} \quad (4.26)$$

By substituting (4.25) and (4.26) into (4.24), one gets:

$$\sigma_n^2 = |d_{min}|^2 (\alpha_{s,d} \beta_0 |y_{s,d}[k-2] y_{s,d}[k-2]|^2$$

$$+ \sum_{i=1}^R \alpha_{s,r_{i,d}} \beta_{1_i} |y_{s,r_{i,d}}[k+1] y_{s,r_{i,d}}[k+2]|^2) \quad (4.27)$$

Therefore the conditional PEP can be further written as:

$$P_E(x \rightarrow \hat{x} | y_{s,d}, y_{s,r_{i,d}}, h_{r_{i,d}}) = Q\left(\frac{a - \mu_n}{\sigma_n}\right) = Q\left(\sqrt{\Gamma_{s,d} + \sum_{i=1}^R \Gamma_{s,r_{i,d}}}\right) \quad (4.28)$$

where $Q(x)$ represents the Gaussian Q function and

$$\Gamma_{s,d} = \frac{\gamma_{s,d} |d(x \rightarrow \hat{x})|^2}{2P_S + 1} |y_{s,d}[k-1]|^2 |y_{s,d}[k-2]|^2 \quad (4.29)$$

$$\Gamma_{s,r_{i,d}} = \frac{\gamma_{s,r_{i,d}} |d(x \rightarrow \hat{x})|^2}{2\rho_{s,r_{i,d}} + 1} |y_{s,r_{i,d}}[k-1]|^2 |y_{s,r_{i,d}}[k-2]|^2 \quad (4.30)$$

with $\rho_{s,r_{i,d}} = G_i P_S |h_{r_{i,d}}[l]|^2 / \sigma_{s,r_{i,d}}^2$ denotes the averaged received SNR conditioned on

$h_{r_{i,d}}[l]$ as derived in section 3.2. $\gamma_{s,d}$ and $\gamma_{s,r_{i,d}}$ are defined as:

$$\gamma_{s,d} = \frac{\alpha_{s,d}^2 P_S}{2P_S(1 - \alpha_{s,d}^2) + 4 + \frac{2}{P_S}} \quad (4.31)$$

$$\gamma_{s,r_i,d} = \frac{\alpha_{s,r_i,d}^2 P_R}{2\rho_{s,r_i,d}(1 - \alpha_{s,r_i,d}^2) + 4 + \frac{2}{P_R}} \quad (4.32)$$

and $|d(x \rightarrow \hat{x})|^2$ is the Euclidean distance between x and $\hat{x}$.

Then, take the average over the channel distribution of $|y_{s,d}[k-1]y_{s,d}[k-2]|^2$ and $|y_{s,r_i,d}[k-1]y_{s,r_i,d}[k-2]|^2$ utilizing the MGF approach [108] as follows:

$$P_E(x|h_{r_i,d}) = \int_0^{\pi/2} M_{\Gamma_{s,d}}\left(\frac{1}{\sin^2\theta}\right) \prod_{i=0}^R M_{\Gamma_{s,r_i,d}}\left(\frac{1}{\sin^2\theta}\right) d\theta \quad (4.33)$$

where $M_{\Gamma_{s,d}}$ and $M_{\Gamma_{s,r_i,d}}$ are the MGFs of $\Gamma_{s,d}$ and $\Gamma_{s,r_i,d}$, respectively. The MGFs of $\Gamma_{s,d}$ and $\Gamma_{s,r_i,d}$ can be written as [109]:

$$\begin{aligned} M_{\Gamma_{s,d}}(s) &= \frac{1}{1 - s|d_{min}|^2 \gamma_{s,d}} \\ M_{\Gamma_{s,r_i,d}}(s) &= \frac{1}{1 - s|d_{min}|^2 \gamma_{s,r_i,d}} \end{aligned} \quad (4.34)$$

By substituting (4.34) into (4.33), one gets:

$$P_E(x|h_{r_i,d}) = \frac{1}{\pi} \int_0^{\pi/2} \left(\frac{1}{1 + \frac{\gamma_{s,d}}{\sin^2\theta} |d_{min}|^2} \right) \prod_{i=0}^R \left(\frac{1}{1 + \frac{\gamma_{s,r_i,d}}{\sin^2\theta} |d_{min}|^2} \right) d\theta \quad (4.35)$$

In order to obtain unconditioned PEP, averaged the expression (4.35) over the distribution of $|h_{r_i,d}|^2$, $\int h_i = e^{-h_i}$. Thus, the unconditioned PEP is given as follows:

$$P_E(x) = \frac{1}{\pi} \int_0^{\pi/2} \frac{\prod_{i=0}^R H_i(\theta)}{1 + \frac{\gamma_{s,d}}{\sin^2\theta} |d_{min}|^2} d\theta \quad (4.36)$$

wherein by manipulating 3.3526⁸ in [99] gives:

$$\prod_{i=0}^R H_i(\theta) = \int_0^\infty \frac{e^{-h_i}}{1 + \frac{\gamma_{s,r_i,d}}{\sin^2 \theta} |d_{min}|^2} dh_i = e^{\epsilon_i(\theta)} E_1(\epsilon_i(\theta)) \quad (4.37)$$

with $\epsilon_i(\theta)$ defined as:

$$\epsilon_i(\theta) = \frac{4}{\frac{1}{\sin^2 \theta} \alpha_{s,r_i,d}^2 G_i^2 |d_{min}|^2 + 2(1 - \alpha_{s,r_i,d}^2) G_i^2 P_S + 4G_i^2} \quad (4.38)$$

and $E_1(\beta) = \int_{-\beta}^\infty \frac{e^{-t}}{t} dt$ denotes the exponential integral function. (4.36) can be computed to get the PEP. It is noted from $\gamma_{s,d}$ and $H_i(\theta)$ that the error performance is subjected to the channel fading rates, $\alpha_{s,d}$ and $\alpha_{s,r_i,d}$. When all the channels are in slow fading (i.e. $\alpha_{s,d} = 1$ and $\alpha_{s,r_i,d} = 1$) it can be seen that $\gamma_{s,d}$ is directly proportional to P_S and $H_i(\theta)$ is indirectly proportional to P_S . Thus, the spatial diversity of the network is attained. However, when the channels are in fast fading condition, the denominator of $\gamma_{s,d}$ and $H_i(\theta)$ become significant. This effects the $\gamma_{s,d}$ and $\gamma_{s,r_i,d}$ particularly in high SNR region. The decrement performance of $\gamma_{s,d}$ and $\gamma_{s,r_i,d}$ consequently decrease the overall performance and the diversity of the system.

It should be mentioned that the numerical analysis PEP is expressed based on the optimum combining weights. Thus the simulation results will give a lower bound for the system employing the proposed combining weights.

4.4.2 Outage Probability

This section investigates the outage probability of signal received at the intended destination. So as to analyse the network outage probability performance, the instantaneous SNR distribution at the MRC combining coefficient output of (4.7) and (4.8) is derived. The optimum combining coefficient is utilized due to the complexity of obtaining the

performance using the practical combining weights of (4.14) and (4.15). Thus, the performance based on the optimum combining coefficient is served as a benchmark for the time-varying performance using the proposed weight. It is observed that the performance analysis approach is similarly applied in [12].

In the outage probability performance, it is defined that outage occurs if the instantaneous SNR γ_T received at the destination falls below a certain specific threshold γ_{th} . Thus, the probability of outage can be written as:

$$P_{out} = \Pr(\gamma_T \leq \gamma_{th}) \quad (4.39)$$

where the instantaneous SNR $\gamma_T = \gamma_{s,d} + \gamma_{s,r,d}$ with $\gamma_{s,d}$ and $\gamma_{s,r,d}$ are two independent random variables.

By referring to [12], the instantaneous SNR received of the direct and relayed links, $\gamma_{s,d}$ and $\gamma_{s,r,d}$ can be represented as:

$$\gamma_{s,d} = \frac{\alpha_{s,d}^2 P_S}{2P_S(1 - \alpha_{s,d}^2) + 4 + 2/P_S} \quad (4.40)$$

and

$$\gamma_{s,r,d} = \frac{\alpha_{s,r,d}^2 P_R}{2P_R(1 - \alpha_{s,r,d}^2) + 4 + 2/P_R} \quad (4.41)$$

Since the outage probability for DDAF is the CDF of $\gamma_T = F_{\gamma_T}(\gamma_{th})$, then the outage probability of DDAF relaying is:

$$P_{out} = F_{\gamma_T}(\gamma_{th}) \quad (4.42)$$

where the CDF of γ_T is written as:

$$\begin{aligned} F_{\gamma_T}(\gamma_{th}) &= \Pr(0 \leq \gamma_T \leq \gamma_{th}) \\ &= \int_0^{\gamma_{th}} \Pr\{0 \leq \gamma_{s,r,d} \leq x - \gamma_{s,d}\} P_{\gamma_{s,d}}(\gamma_{s,d}) d\gamma_{s,d} \end{aligned} \quad (4.43)$$

When $\gamma_{s,r} = \frac{\alpha_{s,r}^2 P_S}{2P_S(1-\alpha_{s,r}^2)+4+2/P_S}$ and $\gamma_{r,d} = \frac{\alpha_{r,d}^2 P_R}{2P_R(1-\alpha_{r,d}^2)+4+2/P_R}$, the CDF of $\gamma_{s,r,d}$

can be determined as 5.41¹¹ and 12.111 in [99]:

$$\begin{aligned}
F_{\gamma_{s,r,d}}(\gamma_{th}) &= \Pr\{0 \leq \frac{\gamma_{s,r}\gamma_{r,d}}{1 + \bar{\gamma}_{s,d} + \gamma_{r,d}} \leq \gamma_{th}\} \\
&= \int_0^\infty \Pr\left\{0 \leq \gamma_{s,r} \leq \frac{(1 + \bar{\gamma}_{s,d})\gamma_{th}}{\gamma_{r,d}} + \gamma_{th}\right\} P_{\gamma_{r,d}}(\gamma_{r,d}) d\gamma_{r,d} \\
&= 1 - \frac{e^{-\frac{x\gamma_{s,d}}{\bar{\gamma}_{s,r}}}}{\bar{\gamma}_{r,d}} \int_0^\infty e^{-\frac{\gamma_{r,d}}{\bar{\gamma}_{r,d}} - \frac{\left(1 + \left(\frac{1}{\bar{\gamma}_{s,r}}\right)\right)\gamma_{th}}{\gamma_{r,d}}} d\gamma_{r,d}
\end{aligned} \tag{4.44}$$

Substitute (4.44) into (4.43) and since the probability of Rayleigh distribution is $p(t) =$

$\frac{1}{\sigma^2} e^{-t/\sigma^2}$, the CDF of γ_T becomes:

$$\begin{aligned}
&F_{\gamma_T}(\gamma_{th}) \\
&= \int_0^{\gamma_{th}} \left(1 - \frac{e^{-\frac{(\gamma_{th}-\gamma_{s,d})\gamma_{s,d}}{\bar{\gamma}_{s,r}}}}{\bar{\gamma}_{r,d}} \int_0^\infty e^{-\frac{t}{\bar{\gamma}_{r,d}} - \frac{\left(1 + \left(\frac{1}{\bar{\gamma}_{s,r}}\right)\right)(\gamma_{th}-\gamma_{s,d})}{t}} dt\right) \left(\frac{1}{\bar{\gamma}_{s,d}} e^{-\frac{\gamma_{s,d}}{\bar{\gamma}_{s,d}}}\right) d\gamma_{s,d} \\
&= \int_0^{\gamma_{th}} \left(\frac{1}{\bar{\gamma}_{s,d}} e^{-\frac{\gamma_{s,d}}{\bar{\gamma}_{s,d}}}\right) d\gamma_{s,d} \\
&\quad - \int_0^{\gamma_{th}} \frac{e^{-\frac{(\gamma_{th}-\gamma_{s,d})\gamma_{s,d}}{\bar{\gamma}_{s,r}}}}{\bar{\gamma}_{r,d}} \left[\int_0^\infty e^{-\frac{t}{\bar{\gamma}_{r,d}} - \frac{\left(1 + \left(\frac{1}{\bar{\gamma}_{s,r}}\right)\right)(\gamma_{th}-\gamma_{s,d})}{t}} dt\right] \\
&\quad \times \left(\frac{1}{\bar{\gamma}_{s,d}} e^{-\frac{\gamma_{s,d}}{\bar{\gamma}_{s,d}}}\right) d\gamma_{s,d}
\end{aligned} \tag{4.45}$$

In (4.45), the expression comprises of two parts and thus can be written as:

$$F_{\gamma_T}(\gamma_{th}) = F_1 + F_2 \tag{4.46}$$

where

$$F_1 = \int_0^{\gamma_{th}} \left(\frac{1}{\bar{\gamma}_{s,d}} e^{-\frac{\gamma_{s,d}}{\bar{\gamma}_{s,d}}}\right) d\gamma_{s,d} = 1 - e^{-\frac{\gamma_{th}}{\bar{\gamma}_{s,d}}} \tag{4.47}$$

and by integrating in terms of variable γ_{th} , then follows by t , the second part of the integral in (4.45) can be expressed as:

$$\begin{aligned}
 F_2 &= \int_0^\infty \int_0^{\gamma_{th}} \frac{1}{\bar{\gamma}_{rd}\bar{\gamma}_{s,d}} e^{-\frac{\gamma_{th}-\gamma_{s,d}}{\bar{\gamma}_{s,r}} - \frac{t}{\bar{\gamma}_{r,d}} - \frac{\left(1+\frac{1}{\bar{\gamma}_{s,r}}\right)(\gamma_{th}-\gamma_{s,d})}{t} - \frac{\gamma_{s,d}}{\bar{\gamma}_{s,d}}} d\gamma_{s,d} dt \\
 &= \int_0^\infty \frac{1}{\bar{\gamma}_{rd}\bar{\gamma}_{s,d}} e^{-\frac{t}{\bar{\gamma}_{r,d}}} \times \left(\frac{1}{\bar{\gamma}_{s,d}} - \frac{1}{\bar{\gamma}_{s,r}} - \left(\frac{1+\frac{1}{\bar{\gamma}_{s,r}}}{t} \right) \times \left(e^{-\left(\frac{1+\frac{1}{\bar{\gamma}_{s,r}}}{t} + \frac{1}{\bar{\gamma}_{s,r}} \right) \gamma_{th}} - e^{-\frac{\gamma_{th}}{\bar{\gamma}_{s,d}}} \right) \right) dt
 \end{aligned} \tag{4.48}$$

Thus combining both parts, the CDF of the SNR γ_T is obtained and the outage probability of the DDAF relaying is given as:

$$\begin{aligned}
 P_{out}(\gamma_{th}) &= 1 - e^{-\frac{\gamma_{th}}{\bar{\gamma}_{s,d}}} - \int_0^\infty \frac{1}{\bar{\gamma}_{rd}\bar{\gamma}_{s,d}} e^{-\frac{t}{\bar{\gamma}_{r,d}}} \times \left(\frac{1}{\bar{\gamma}_{s,d}} - \frac{1}{\bar{\gamma}_{s,r}} - \left(\frac{1+\frac{1}{\bar{\gamma}_{s,r}}}{t} \right) \right. \\
 &\quad \times \left. \left(e^{-\left(\frac{1+\frac{1}{\bar{\gamma}_{s,r}}}{t} + \frac{1}{\bar{\gamma}_{s,r}} \right) \gamma_{th}} - e^{-\frac{\gamma_{th}}{\bar{\gamma}_{s,d}}} \right) \right) dt
 \end{aligned} \tag{4.49}$$

It is noted that further simplification of equation (4.49) is difficult, thus symmetric case of $\bar{\gamma}_{s,d} = \bar{\gamma}_{s,r}$ is considered. Hence, (4.49) is simplified and its closed-form expression can be written as:

$$\begin{aligned}
 P_{out}(\gamma_{th}) &= 1 - e^{-\frac{\gamma_{th}}{\bar{\gamma}_{s,d}}} - e^{-\frac{\gamma_{th}}{\bar{\gamma}_{s,r}}} \frac{\bar{\gamma}_{r,d}}{\bar{\gamma}_{s,r} \left(1 + \frac{1}{\bar{\gamma}_{s,r}} \right)} \\
 &\quad + 2e^{-\frac{\gamma_{th}}{\bar{\gamma}_{s,r}}} \left(\frac{\gamma_{th}}{\bar{\gamma}_{s,r}} \right) K_2 \left(\sqrt{\left(1 + \frac{1}{\bar{\beta}_{s,r}} \right) \frac{\gamma_{th}}{\bar{\gamma}_{r,d}}} \right)
 \end{aligned} \tag{4.50}$$

4.5 Discussion of Simulation Results for Maximal Ratio Combining Scheme

As in a pure differential system, the destination cannot possess the exact channel knowledge. Therefore, an MRC employing set of fixed weights pertaining to the second-order statistics of all channels has been studied and analysed in [12, 17] to achieve diversity. In this thesis, in order to eliminate the channel and frequency offset estimation, the channel gains of all channels are replaced by their variances and a semi MRC receiver is suggested at the destination.

In this section, a DDAF cooperative network with multiple relays is simulated and presented in various cases of different fading channels and communication terminals' mobility to demonstrate the robust performance of the proposed combiners. All the channels are perturbed with frequency offsets and are generated independently following the mobile-to-mobile wireless communication simulation method of [94]. Also, two performance metrics, BER and outage probability are focused and simulated under slow and fast fading channels. Slowly-faded Rayleigh channel is commenced utilizing normalized Doppler frequency of 0.001, which is increased later up to 0.05 to model a high mobility scenario.

Conventional MRC and proposed semi MRC schemes are incorporated into the cooperative wireless multiple-relay networks, combining the signals received in both stages (i.e. the directly transmitted and relay link) to jointly detect the transmitted data. The information signal is encoded with double differential modulation for BPSK and QPSK constellations. The signals are transmitted in blocks. Since prior knowledge of the channel is unavailable, thus an amplification factor at the relay is set to $G = (\frac{P_{Ri}}{P_S+1})^{1/2}$. For the purpose of simulation, the amplification gain is specified in such approach to ensure that the average transmission power at the relay is fixed to P_R in a long term period. Moreover, to facilitate numerical performance evaluations, symmetry power allocation at the source $P_S =$

$\frac{P_T}{2}$ and relay $P_{Ri} = \frac{P_T}{2R}$ is assumed, respectively in the network. P_T represents the total of the power utilized in the wireless network where $P_T = P_S + P_R$.

4.5.1 Error Probability Performance

Figure 4.1 compares the BER performance of the conventional MRC and proposed semi MRC scheme with DDBPSK modulation. It is considered that the channel variances $\sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1; \sigma_{s,d}^2 = 1$. Slow fading channels are considered, that is, normalized Doppler frequency $f_{D_{s,d}}T_s = f_{D_{s,r}}T_s = f_{D_{r,d}}T_s = 0.05$.

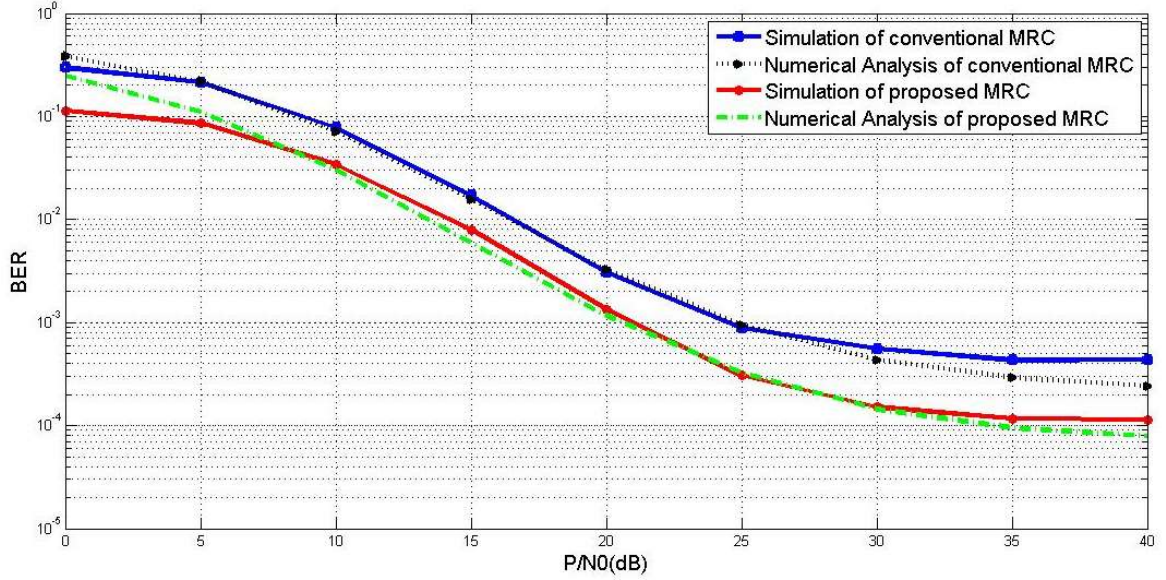


Figure 4.1: Comparison of the BER performance for conventional MRC and proposed MRC utilizing DDBPSK with 2 relays for $f_{D_{s,d}}T_s = f_{D_{s,r}}T_s = f_{D_{r,d}}T_s = 0.05$

It is noteworthy that the total power P_T is also similar to SNR since the variance of the channel gains and AWGN is normalized to 1. At the destination, the received signals are combined with the proposed weights so that Euclidean-distance detection can be executed to recover back the originally transmitted signals. For comparison wise, the similar simulation process is carried out but employing the optimum combining weights given in

(4.8) and (4.9) for the MSDDSD system. Both schemes are simulated and plotted for BER versus SNR. It is shown that even when slow fading channels are involved, the conventional MRC scheme suffers severe performance degradation while the proposed semi MRC is able to achieve excellent performance. For instance, the performance gap reduces to 2-3 dB throughout the SNR value. This is because, the new combiner weights of semi MRC from (4.14) and (4.15) are dependent on the auto-correlation of the channel, relay gain and average source power. On the contrary, the conventional MRC combiner consider only the relay gain. Also, the simulated BER curves with solid lines (different legends) for both combining schemes are plotted along with the theoretical numerical analysis expression of BER based on PEP approach given in (4.36) with dashed lines. As observed, both analytical expression matches tightly with their simulated results, even in the low SNR region.

Figure 4.2 demonstrates the BER performance of MSDDSD with semi MRC serves as combiner at the intended destination under fast fading channels. It is demonstrated from the figure that the proposed scheme is capable of minimizing the error performance when the number of relays increased from 2 to 4 throughout the range of $\frac{P}{N_0}$. As the number of relays increases to 4, it indicates that diversity order is achieved and is in agreement with the derived Equation 4.38. The achieved diversity order indicates that more diversity gain is

provided than the direct transmission scheme. Thus, the performance of the system is further improved where the error is minimized.

4.5.2 Outage Probability

The outage performance versus transmitted power for the DDAF network is performed based on the theoretical numerical analysis equation formed in Section 4.3.2.2. The numerical analysis is validated with the multiple symbol transmission where semi MRC serves as the combiner at the intended destination in the case where time-varying Rayleigh channel is considered. Random mutually independent and uniformly distributed frequency offset $2\pi f_d T$ is perturbed over the range of $[-\pi, \pi]$. Figure 4.3 demonstrates the simulated outage probabilities in line with their corresponding numerical analysis approximations. As can be observed from the curve, the approximate numerical results plot matches well with the simulation results throughout the entire SNR range. Moreover, the simulation results show tightness at high SNR region. The numerical expression significantly shows that the outage probability can be efficiently forecasted. It can be observed also, that the achieved diversity order is dependent on the number of relays.

Figure 4.3 depicts the effect of relay configuration on the outage probability performance of the MSDDSD-AF system employing semi MRC.

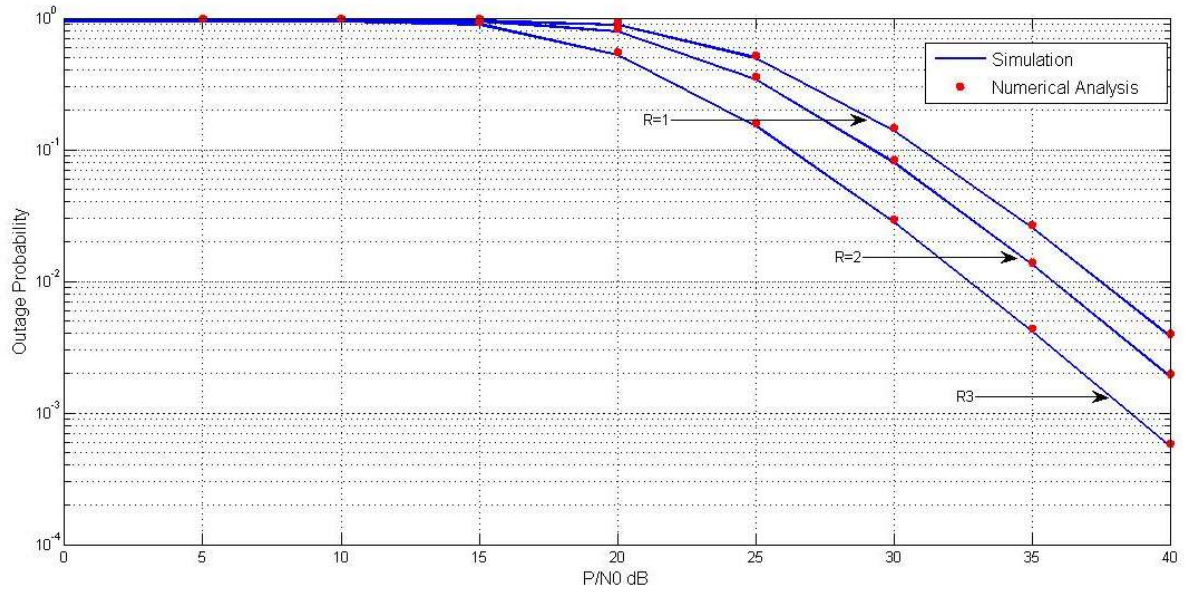


Figure 4.3: Outage probabilities of the MSDDSD-AF network for BPSK with semi MRC employing different numbers of relay terminals

At transmitted power of 40 dB, the relay configuration of 1, 2 and 3 give 0.15, 0.08 and 0.03, respectively. This result reviews that the system outage reduces by an average of 54.59% with increasing number of relays.

The diversity order increases proportional against the number of relays used. It is noted also, to fairly compare the performances, the total transmit power at the source and relay are assumed to be similar; in other words, $P_T = P_S + P_R$.

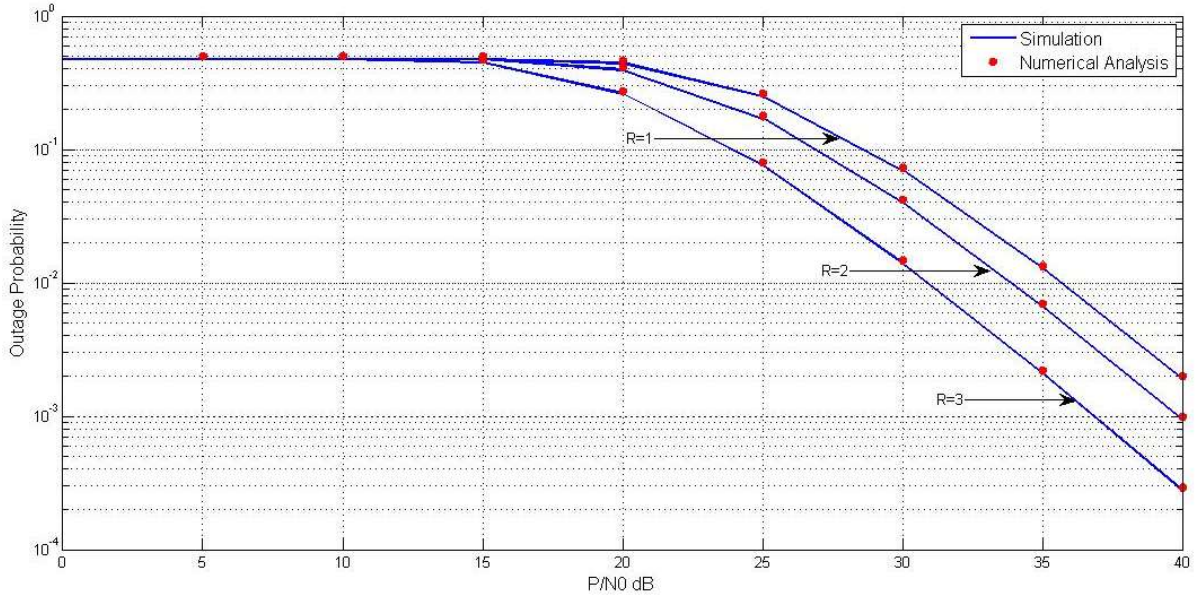


Figure 4.4: Outage probabilities of the MSDDSD-AF network for QPSK with semi MRC employing different numbers of relay terminals

The outage probabilities of the MSDDSD AF relay network is further analysed for different terminals' mobility environments employing the semi MRC. The obtained theoretical outage probability of the semi MRC scheme is verified by simulation results. Considering a symmetrical channels with $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$; $\sigma_{s,d}^2 = 1$, and equally distributed power at both source and relay terminals, the simulation results are plotted in Figure 4.5 and Figure 4.6 for DDBPSK and DDQPSK, respectively. Three cases are taken into consideration and tabulated in Table 3.5 (Section 3.4.2).

At transmitted power of 40 dB, the outage probability of the semi MRC for case I, case II and case III are 0.8, 0.015 and 0.005, respectively. This shows that the semi MRC employed under case III has 66.67% reduced outage probability compared to semi MRC in case 2 and For all values of transmitted power, the semi MRC for fast fading channels give lower outages compared to semi MRC that is utilized in case I and case II fading channels.

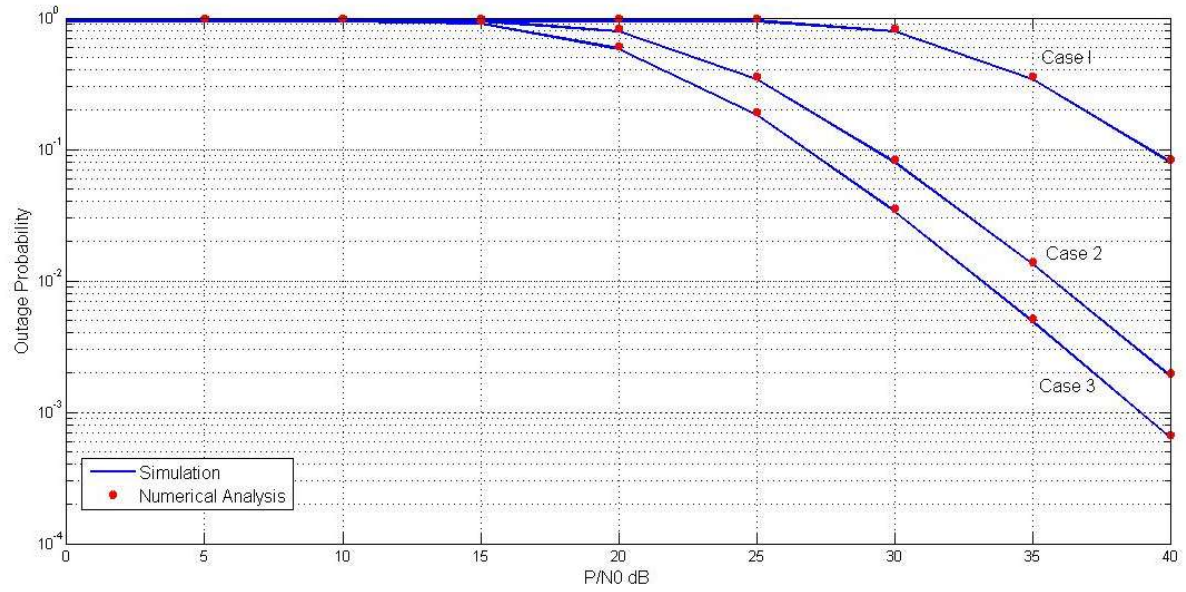


Figure 4.5: Outage probability performance of proposed MRC under different faded channels for MSDDSD-AF system for BPSK constellation

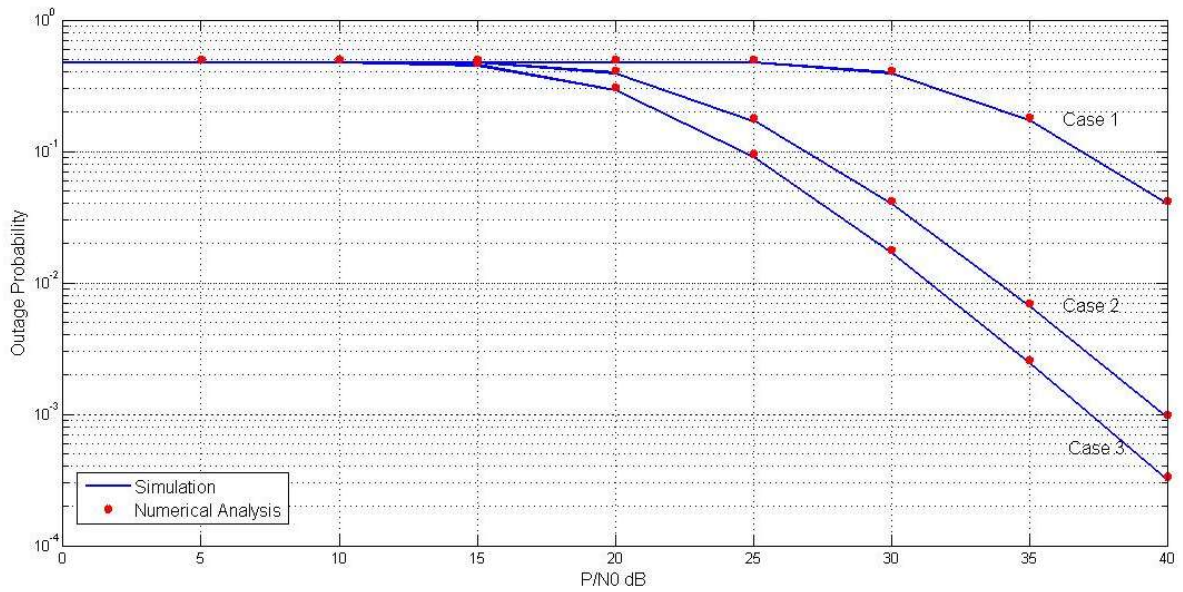


Figure 4.6: Outage probability performance of proposed semi MRC under different faded channels for MSDDSD-AF system for QPSK constellation

4.6 Selection Combiner

Considering three terminals of one source, one relay and one destination, the system model in this sub section is similar to Figure 3.1. By exploiting the wireless channel, a doubly-differential encoded signal is transmitted from the source via two transmission links i.e., towards the fixed-in AF relay and directly to the destination. In order to attain the diversity of cooperation network, the transmitted signals from both relayed and direct links are combined using some combining schemes. From the previous section 4.3.1, semi MRC has been proposed under slow fading channels as:

$$\zeta = \frac{1}{N_0} \zeta_{s,d} + \frac{1}{N_0(1 + G\sigma_{r,d}^2)} \zeta_{s,r,d} \quad (4.51)$$

It can be seen from (4.51) that statistical knowledge of all wireless channels are required. Thus, in order to avoid the channel and frequency offset estimation, SC scheme is proposed. For selection combining approach, only the best relayed signal possessing the highest signal magnitude among the relayed signals is chosen. Then, the receiver at the destination will consider the signal from the source as well as the one from the best relayed signal and select the one with a larger SNR for decoding purpose. The block diagram of the DDAF network employing SC scheme at the destination is portrayed in Figure 4.7.

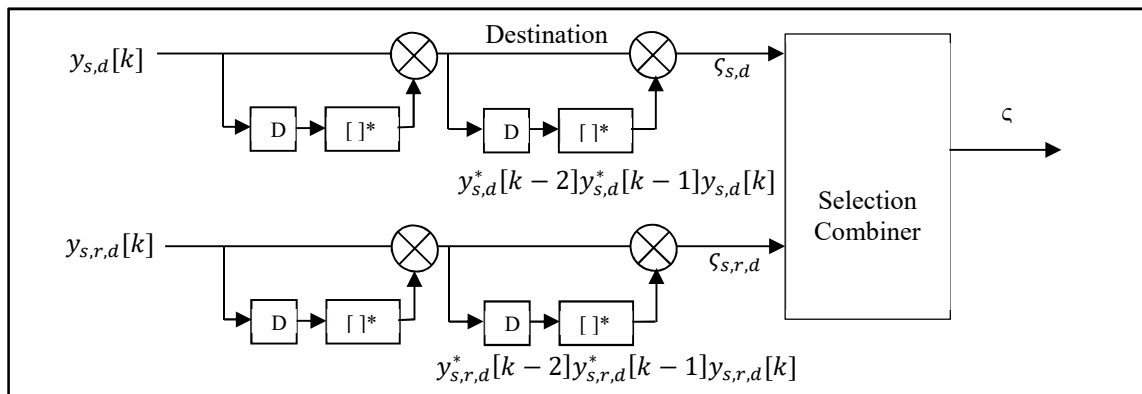


Figure 4.7: The block diagram of SC scheme at the receiver

For DDAF network employing BPSK constellations, the decision variables at the destination for direct and relayed link are computed from the three latest received symbols are written as:

$$\varsigma_{s,d} = R\{(z_{s,d}[n]z_{s,d}^*[n-1])(z_{s,d}[n-1]z_{s,d}^*[n-2])^*\} \quad (4.52)$$

$$\varsigma_{s,r,d} = R\{(z_{s,r,d}[l]z_{s,r,d}^*[l-1])(z_{s,r,d}[l-1]z_{s,r,d}^*[l-2])^*\} \quad (4.53)$$

The decision variables are then, compared in terms of the larger magnitude of the links.

Hence the output of the SC can be obtained as:

$$\varsigma = \begin{cases} \varsigma_{s,d} & \text{if } |\varsigma_{s,d}| > |\varsigma_{s,r,d}| \\ \varsigma_{s,r,d} & \text{if } |\varsigma_{s,d}| < |\varsigma_{s,r,d}| \end{cases} \quad (4.54)$$

Then, the combiner output is used to decode the transmitted signal as:

$$\hat{x}[k] = \begin{cases} -1 & \text{if } \varsigma \leq 0 \\ +1 & \text{if } \varsigma > 0 \end{cases} \quad (4.55)$$

4.7 Performance Analysis of Selection Combiner

It is assumed that $s[0] = s[1] = p[1] = 1$ as an initialization. Since only direct link and single relay link is considered, the probability of error can be given by [110] as:

$$P_b(E) = \Pr(|\varsigma_{s,d}| > \varsigma_{s,r,d}, \varsigma_{s,d} < 0) + \Pr(|\varsigma_{s,r,d}| > \varsigma_{s,d}, \varsigma_{s,r,d} < 0) \quad (4.56)$$

Since $\varsigma_{s,d}$ and $\varsigma_{s,r,d}$ have different distributions, the probability of error (4.56) is computed separately. By averaging over the PDF of $\varsigma_{s,d}$ and $\varsigma_{s,r,d}$, the average error probability can be expressed as:

$$\begin{aligned} P_b(E_1) &= \Pr(|\varsigma_{s,d}| > \varsigma_{s,r,d}, \varsigma_{s,d} < 0) = \int_{-\infty}^0 \int_0^{-z} f_{\varsigma_{s,d}}(z) f_{\varsigma_{s,r,d}}(r) dr dz \\ &= \int_{-\infty}^0 f_{\varsigma_{s,d}}(z) [F_{|\varsigma_{s,r,d}|}(-z) - F_{|\varsigma_{s,r,d}|}(0)] dz \end{aligned} \quad (4.57)$$

and

$$\begin{aligned}
P_b(E_2) &= \Pr(|\zeta_{s,d}| > \zeta_{s,r,d}, \zeta_{s,d} < 0) = \int_{-\infty}^0 \int_0^{-z} f_{\zeta_{s,r,d}}(z) f_{\zeta_{s,d}}(r) dr dy \\
&= \int_{-\infty}^0 f_{\zeta_{s,r,d}}(z) [F_{|\zeta_{s,d}|}(-z) - F_{|\zeta_{s,d}|}(0)] dz
\end{aligned} \tag{4.58}$$

From (4.57) and (4.58), it is known that, the PDF of ζ is related to its CDF $F_\zeta(\zeta)$ by:

$$f_\zeta(z) = \frac{dF_\zeta(z)}{d\zeta} \tag{4.59}$$

The CDF of $\zeta_{s,d}$ and $\zeta_{s,r,d}$ can be obtained as:

$$\begin{aligned}
F_{|\zeta_{s,d}|}(z) &= \Pr(|\zeta_{s,d}| \leq z) = \Pr(-z \leq \zeta_{s,d} \leq z) \\
&= F_{\zeta_{s,d}}(z) - F_{\zeta_{s,d}}(-z)
\end{aligned} \tag{4.60}$$

and

$$\begin{aligned}
F_{|\zeta_{s,r,d}|}(z) &= \Pr(|\zeta_{s,r,d}| \leq z) = \Pr(-z \leq \zeta_{s,r,d} \leq z) \\
&= F_{\zeta_{s,r,d}}(z) - F_{\zeta_{s,r,d}}(-z)
\end{aligned} \tag{4.61}$$

In order to proceed with the computation of the probability error, the PDFs and CDFs of the decision variables, $\zeta_{s,d}$ and $\zeta_{s,r,d}$ are required. As formulated in Section 3.2.2, the individual time-varying channel based on AR(2) is given by:

$$h_{s,r}[k] = \alpha_{s,r}^2 h_{s,r}[k-2] + \alpha_{s,r} \varepsilon_{s,r}[k-1] + \varepsilon_{s,r}[k] \tag{4.62}$$

And for a relayed channel, it is given as:

$$h[k] = \alpha^2 h[k-2] + \alpha h_{s,r}[k-2] \varepsilon_{s,r}[k] \tag{4.63}$$

where $\alpha^2 = (\alpha_{s,r} \alpha_{r,d})^2 \leq 1$ is the autocorrelation of the relayed channel.

By substituting the source-destination channel model of (3.15) and the transmitted symbols $x[k]$ into $y_{s,d}$ (3.2), one gets:

$$y_{s,d}[k] = \alpha^2 x[k] y_{s,d}[k-2] e^{j2\pi f_{s,d}[k]} + \tilde{\varepsilon}_{s,d}[k] \tag{4.64}$$

where

$$\begin{aligned}\tilde{\varepsilon}_{s,d}[k] &= \alpha P_S h_{s,d}[k-2] \varepsilon_{s,d}[k-1] e^{j2\pi f_{s,d}[k]} x[k] + \\ &P_S h_{s,d}[k-2] \varepsilon_{s,d}[k] e^{j2\pi f_{s,d}[k]} x[k] + \varepsilon_{s,d}[k]\end{aligned}\quad (4.65)$$

Similarly, by substituting the relayed channel model of (3.19) and the transmitted symbols into $y_{s,r,d}$, one gets:

$$y_{s,r,d}[k] = \alpha^2 x[k] y_{r,d}[k-2] e^{j2\pi f_{r,d}[k]} + \tilde{\varepsilon}[k] \quad (4.66)$$

where

$$\begin{aligned}\tilde{\varepsilon}_{r,d}[k] &= \alpha P_S h_{r,d}[k-2] \varepsilon_{r,d}[k-1] e^{j2\pi f_{r,d}[k]} P_S x[k] + \\ &P_S h_{r,d}[k-2] \varepsilon_{r,d}[k] e^{j2\pi f_{r,d}[k]} x[k] + \varepsilon_{r,d}[k]\end{aligned}\quad (4.67)$$

Then, substitutes $y_{s,d}$ and $y_{s,r,d}$ into the decision variables from (4.52) and (4.53) gives:

$$\varsigma_{s,d} = R\{\alpha^2 x[k] |y_{s,d}[k-1]|^3 y_{s,d}[k-2] + |y_{s,d}[k-1]|^2 y_{s,d}^*[k-2] \tilde{\varepsilon}[k]\} \quad (4.68)$$

and

$$\varsigma_{s,r,d} = R\{\alpha^2 x[k] |y_{s,r,d}[k-1]|^3 y_{s,r,d}[k-2] + |y_{s,r,d}[k-1]|^2 y_{s,r,d}^*[k-2] \tilde{\varepsilon}[k]\} \quad (4.69)$$

It can be observed that, conditioned on $y_{s,d}[k-1]$ and $y_{s,d}[k-2]$, the conditional mean and variance of the decision variables are computed as:

$$\begin{aligned}\mu_{\varsigma_{s,d}} &= E\{\varsigma_{s,d} | y_{s,d}[k-1] y_{s,d}[k-2], x[k] = +1\} \\ &= \alpha_{s,d}^2 |y_{s,d}[k-2]|^3 y_{s,d}[k-2] + E\{R\{|y_{s,d}[k-1]|^2 y_{s,d}^*[k-2] \tilde{\varepsilon}[k]\}\} \\ &= \alpha_{s,d}^2 |y_{s,d}[k-1]|^3 y_{s,d}[k-2] - \alpha_{s,d}^2 E\{R\{\varepsilon[k-1] y_{s,d}[k-1] |^2 y_{s,d}[k-2]\}\} \\ &= \alpha_{s,d}^2 |y_{s,d}[k-1]|^3 y_{s,d}[k-2] - \frac{\alpha_{s,d}^2}{\rho_o + 1} [|y_{s,d}[k-1]|^3 y_{s,d}[k-2]] \\ &= \frac{\alpha_{s,d}^2 \rho_o}{\rho_o + 1} [|y_{s,d}[k-1]|^3 y_{s,d}[k-2]]\end{aligned}\quad (4.70)$$

and

$$\sigma_{\varsigma_{s,d}}^2 = Var\{\varsigma_{s,d} | y_{s,d}[k-1] y_{s,d}[k-2], x[k] = +1\} \quad (4.71)$$

$$\begin{aligned}
&= \text{Var}\{\alpha_{s,d}^2 |y_{s,d}[k-2]|^3 y_{s,d}[k-2] + |y_{s,d}[k-1]|^2 y_{s,d}^*[k-2] \tilde{\varepsilon}[k]\} \\
&= \text{Var}\{|y_{s,d}[k-1]|^2 y_{s,d}^*[k-2] \tilde{\varepsilon}[k]\} \\
&= \frac{1}{2} N_0 |y_{s,d}[k-2]|^2 |y_{s,d}[k-1]|^2 + \frac{1}{2} N_0 \frac{\alpha_{s,d}^2 \rho_0}{\rho_0 + 1} |y_{s,d}[k-2]|^2 |y_{s,d}[k-1]|^2 + \\
&\quad \frac{1}{2} N_0 (1 - \alpha_0^2) \rho_0 |y_{s,d}[k-2]|^2 |y_{s,d}[k-1]|^2 \\
&= \frac{1}{2} N_0 \left\{ 1 + \frac{\alpha_{s,d}^2 \rho_0}{\rho_0 + 1} + (1 - \alpha_{s,d}^2) \rho_0 \right\} |y_{s,d}[k-2]|^2 |y_{s,d}[k-1]|^2
\end{aligned}$$

Additionally, conditioned on $y_{r,d}[k-1]$ and $y_{r,d}[k-2]$, the conditional mean as well as variance of the decision variables are computed as:

$$\begin{aligned}
\mu_{\zeta_{s,r,d}} &= E\{\zeta_{s,r,d} | y_{r,d}[k-1] y_{r,d}[k-2], x[k] = +1\} \\
&= \alpha_{r,d}^2 |y_{r,d}[k-2]|^3 y_{r,d}[k-2] + E\{R\{|y_{r,d}[k-1]|^2 y_{r,d}^*[k-2] \tilde{\varepsilon}[k]\}\} \\
&= \alpha_{r,d}^2 |y_{r,d}[k-1]|^3 y_{r,d}[k-2] - \alpha_{r,d}^2 E\{R\{\varepsilon[k-1] y_{r,d}[k-1] |y_{r,d}[k-2]\}\} \\
&= \alpha_{r,d}^2 |y_{r,d}[k-1]|^3 y_{r,d}[k-2] - \frac{\alpha_{r,d}^2}{\rho_1 + 1} [|y_{r,d}[k-1]|^3 y_{r,d}[k-2]] \\
&= \frac{\alpha_{r,d}^2 \rho_1}{\rho_1 + 1} [|y_{r,d}[k-1]|^3 y_{r,d}[k-2]]
\end{aligned} \tag{4.72}$$

and

$$\begin{aligned}
\sigma_{\zeta_{s,r,d}}^2 &= \text{Var}\{\zeta_{s,r,d} | y_{r,d}[k-1] y_{r,d}[k-2], x[k] = +1\} \\
&= \text{Var}\{\alpha_{r,d}^2 |y_{r,d}[k-2]|^3 y_{r,d}[k-2] + |y_{r,d}[k-1]|^2 y_{r,d}^*[k-2] \tilde{\varepsilon}[k]\} \\
&= \text{Var}\{|y_{r,d}[k-1]|^2 y_{r,d}^*[k-2] \tilde{\varepsilon}[k]\} \\
&= \frac{1}{2} N_0 |y_{r,d}[k-2]|^2 |y_{r,d}[k-1]|^2 + \frac{1}{2} N_0 \frac{\alpha_{r,d}^2 \rho_1}{\rho_0 + 1} |y_{r,d}[k-2]|^2 |y_{r,d}[k-1]|^2 + \\
&\quad \frac{1}{2} N_0 (1 - \alpha_2^2) \rho_1 |y_{r,d}[k-2]|^2 |y_{r,d}[k-1]|^2
\end{aligned} \tag{4.73}$$

$$= \frac{1}{2} N_0 \left\{ 1 + \frac{\alpha_{r,d}^2 \rho_1}{\rho_0 + 1} + (1 - \alpha_2^2) \rho_0 \right\} |y_{s,d}[k-2]|^2 |y_{s,d}[k-1]|^2$$

The conditional PDF of $\zeta_{s,d}$ can be computed based on [56]:

$$f_{\zeta_{s,d}}(z|y_{s,d}) = \frac{1}{(2\pi\sigma_{\zeta_{s,d}}^2)^{1/2}} e^{-\frac{(z-\mu_{\zeta_{s,d}})^2}{2\sigma_{\zeta_{s,d}}^2}} \quad (4.74)$$

Since the distribution of the channel $h_{s,d}$ is $y_{s,d} \sim \mathcal{CN}(0, N_0(1 + \rho_0))$, therefore $|y_{s,d}|^2 = 0.5N_0^2(1 + \rho_0)^2$. Its PDF is obtained as [1]:

$$f_{y_{s,d}}(X_0) = \frac{\exp\left(-\frac{X_0}{N_0(1 + \rho_0)}\right)}{N_0(1 + \rho_0)} \quad (4.75)$$

By taking the average of (4.75) over the distribution of $y_{s,d}$, the PDF of $\zeta_{s,d}$ is given as [110]:

$$f_{\zeta_{s,d}}(z) = \begin{cases} B_0 \exp(C_0 z), & z \leq 0 \\ B_0 \exp(D_0 z), & z \geq 0 \end{cases} \quad (4.76)$$

where

$$B_0 = \frac{1}{N_0(1 + \rho_0)} \quad (4.77)$$

$$C_0 = \frac{2}{N_0(1 + (1 - \alpha_0)\rho_0)} \quad (4.78)$$

$$D_0 = \frac{-2}{N_0(1 + (1 + \alpha_0)\rho_0)} \quad (4.79)$$

Then, the CDF of $\zeta_{s,d}$ is obtained as:

$$F_{\zeta_{s,d}}(z) = \begin{cases} \frac{B_0}{C_0} \exp(C_0 z), & z \leq 0 \\ 1 + \frac{B_0}{D_0} \exp(D_0 z), & z \geq 0 \end{cases} \quad (4.80)$$

The CDF of $|\zeta_{s,d}|$ is expressed by substituting (4.80) into (4.59) as:

$$\begin{aligned} F_{|\zeta_{s,d}|}(z) &= F_{\zeta_{s,d}}(z) - F_{\zeta_{s,d}}(-z) \\ &= 1 + \frac{B_0}{D_0} \exp(D_0 z) - \frac{B_0}{C_0} \exp(-C_0 z) \end{aligned} \quad (4.81)$$

Similarly, for the relayed link, the conditional PDF of $\zeta_{s,r,d}$ is written as:

$$f_{\zeta_{s,r,d}}(z|y_{s,r,d}, h_{r,d}) = \frac{1}{(2\pi\sigma_{\zeta_{s,r,d}}^2)^{1/2}} e^{-\frac{(z-\mu_{\zeta_{s,r,d}})^2}{2\sigma_{\zeta_{s,r,d}}^2}} \quad (4.82)$$

Since the distribution of the channel $h_{s,r,d}$ is $y_{s,r,d} \sim \mathcal{CN}(0, \sigma_w^2(1 + \rho_1))$, therefore $|y_{s,r,d}|^2 = 0.5\sigma_w^2(1 + \rho_1)^2$. Its PDF is obtained as [1]:

$$f_{y_{s,r,d}}(X_1) = \frac{\exp\left(-\frac{X_1}{\sigma_w^2(1 + \rho_1)}\right)}{\sigma_w^2(1 + \rho_1)} \quad (4.83)$$

By taking the average of (4.83) over the distribution of $y_{s,r,d}$, the PDF of $\zeta_{s,r,d}$ is given as [110]:

$$f_{\zeta_{s,r,d}}(z|h_{r,d}) = \begin{cases} B_1 \exp(C_1 z), & z \leq 0 \\ B_1 \exp(D_1 z), & z \geq 0 \end{cases} \quad (4.84)$$

where

$$B_1 = \frac{1}{\sigma_w^2(1 + \rho_1)} \quad (4.85)$$

$$C_1 = \frac{2}{\sigma_w^2(1 + (1 - \alpha_1)\rho_1)} \quad (4.86)$$

$$D_1 = \frac{-2}{\sigma_w^2(1 + (1 + \alpha_1)\rho_1)} \quad (4.87)$$

Then, the CDF of $\varsigma_{s,r,d}$ is obtained as:

$$F_{\varsigma_{s,r,d}}(z) = \begin{cases} \frac{B_1}{C_1} \exp(C_1 z), & z \leq 0 \\ 1 + \frac{B_1}{D_1} \exp(D_1 z), & z \geq 0 \end{cases} \quad (4.88)$$

The CDF of $|\varsigma_{s,r,d}|$ conditioned on $h_{r,d}$ is expressed by substituting (4.88) into (4.61) as:

$$\begin{aligned} F_{|\varsigma_{s,r,d}|}(z|h_{r,d}) &= F_{\varsigma_{s,r,d}}(z) - F_{\varsigma_{s,r,d}}(-z) \\ &= 1 + \frac{B_1}{D_1} \exp(D_1 z) - \frac{B_1}{C_1} \exp(-C_1 z) \end{aligned} \quad (4.89)$$

Therefore,

$$\begin{aligned} P_b(E_1|h_{r,d}) &= \int_{-\infty}^0 f_{\varsigma_{s,d}}(z) [F_{|\varsigma_{s,r,d}|}(-z|h_{r,d}) - F_{|\varsigma_{s,r,d}|}(0|h_{r,d})] dz \\ &= \int_{-\infty}^0 B_0 e^{C_0 z} \left(\frac{B_1}{D_1} e^{-D_1 z} - \frac{B_1}{C_1} e^{C_1 z} - \frac{B_1}{D_1} + \frac{B_1}{C_1} \right) dz \\ &= \frac{B_0 B_1}{C_0} \left(\frac{1}{C_0 - D_1} + \frac{1}{C_0 + C_1} \right) \end{aligned} \quad (4.90)$$

Similarly,

$$\begin{aligned} P_b(E_2|h_{r,d}) &= \int_{-\infty}^0 f_{\varsigma_{s,r,d}}(z|h_{r,d}) [F_{|\varsigma_{s,d}|}(-z) - F_{|\varsigma_{s,d}|}(0)] dz \\ &= \int_{-\infty}^0 B_2 e^{C_2 z} \left(\frac{B_0}{D_0} e^{-D_0 z} - \frac{B_0}{C_0} e^{C_0 z} - \frac{B_0}{D_0} + \frac{B_0}{C_0} \right) dz = \frac{B_0 B_1}{C_1} \left(\frac{1}{C_1 - D_0} + \frac{1}{C_0 + C_1} \right) \end{aligned} \quad (4.91)$$

Hence, by combining the equation (4.90) and (4.91), the conditional probability of error can be expressed as:

$$P_b(E|h_{r,d}) = P_b(E_1|h_{r,d}) + P_b(E_2|h_{r,d}) = \frac{B_0 B_1}{C_0 C_1} + \frac{B_0 B_1}{C_0(C_0 - D_1)} + \frac{B_0 B_1}{C_1(C_1 - D_0)} \quad (4.92)$$

Taking the average over the distribution of $|h_{r,d}|^2 = h$, the conditional probability of error is calculated as:

$$P_b(E) = \int_0^\infty \left(\frac{B_0 B_1}{C_0 C_1} + \frac{B_0 B_1}{C_0(C_0 - D_1)} + \frac{B_0 B_1}{C_1(C_1 - D_0)} \right) f_h(h) dh$$

$$P_b(E) = P_{b1}(E) + P_{b2}(E) + P_{b3}(E) \quad (4.93)$$

Based on 3.352.4 in [99], $P_{b1}(E)$, $P_{b2}(E)$ and $P_{b3}(E)$ can be solved as:

$$P_{b1}(E) = \int_0^\infty \frac{B_0 B_1}{C_0 C_1} f_h(h) dh = \frac{B_0}{C_0} \int_0^\infty \frac{b_1}{c_1} \frac{h+c_1}{h+b_1} \frac{e^{-\frac{h}{\sigma_{r,d}^2}}}{\sigma_{r,d}^2} dh$$

$$= \frac{B_0 b_1}{C_0 c_1} \left(1 + \frac{b_1 - c_1}{\sigma_{r,d}^2} \exp\left(\frac{c_1}{\sigma_{r,d}^2}\right) E_1\left(\frac{c_1}{\sigma_{r,d}^2}\right) \right) \quad (4.94)$$

where

$$b_1 = \frac{1}{G^2(1 + \rho_1)} \quad (4.95)$$

$$c_1 = \frac{1}{G^2(1 + (1 - \alpha)\rho_1)} \quad (4.96)$$

with $E_i(\mu) = \int_1^\infty \frac{e^{-\mu x}}{x} dx$ indicates the exponential integral function.

$P_{b2}(E)$ is computed as follows:

$$P_{b2}(E) = \int_0^\infty \frac{B_0 B_1}{C_0(C_0 - D_1)} f_h(h) dh = \frac{B_0}{C_0} \int_0^\infty \frac{b_2}{h+b_1} \frac{1}{\sigma_{r,d}^2} e^{-\frac{h}{\sigma_{r,d}^2}} dh \quad (4.97)$$

where

$$\check{b}_1 = \frac{1 + (1 - \alpha_{s,d})\rho_0\rho_1 + 1 + (1 - \alpha_{s,d})\rho_0}{G^2(1 + (1 + \alpha)\rho_1 + (1 + \alpha)\rho_1^2)} \quad (4.98)$$

$$b_2 = \frac{1 + (1 - \alpha)\rho_1 + 1 + (1 - \alpha_{s,d})\rho_0}{G^2(1 + (1 + \alpha)\rho_1 + (1 + (1 - \alpha_{s,d})\rho_0))} \quad (4.99)$$

and $P_{b3}(E)$ can be determined as:

$$P_{b3}(E) = \int_0^\infty \frac{B_0 B_1}{C_1(C_1 - D_0)} f_h(h) dh = \frac{B_0}{C_1} \int_0^\infty \tilde{b}_2 \frac{h + \tilde{b}_0}{h + \tilde{b}_1} dh \quad (4.100)$$

where

$$\tilde{b}_0 = \frac{2}{G^2(1 + (1 - \alpha)\rho_1)} \quad (4.101)$$

$$\tilde{b}_1 = \frac{1 + (1 - \alpha_{s,d})\rho_0\rho_1 + 1 + (1 - \alpha_{s,d})\rho_0}{G^2(1 + (1 + \alpha)\rho_1 + (1 + \alpha)\rho_1^2)} \quad (4.102)$$

$$\tilde{b}_2 = \frac{1 + (1 - \alpha)\rho_1 + 1 + (1 + \alpha_{s,d})\rho_0}{G^2(1 + (1 - \alpha)\rho_1 + (1 + (1 + \alpha_{s,d})\rho_0))} \quad (4.103)$$

4.8 Discussion of Simulation Results for Selection Combiner

In this section, Matlab simulation is conducted in order to examine the overall system performance of the proposed semi MRC and SC. The proposed SC scheme is the simplest among all other combining schemes, where only the maximum amplitude of the link is required to determine the near optimum performance of signal combining. DD modulation and demodulation is implemented to the system at the source and destination, respectively. The proposed methods are tested with respect to various channel fading environments. It is assumed that the variance of the complex channel between any pairs of the terminals is normalized to 1 and 10 for fading gains. For instance, $\sigma_{s,d}^2 = \sigma_{s,r}^2 = \sigma_{r,d}^2 = 1$ refers to the scenarios when the relay is located in the middle between the source and destination terminals. $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 10$ refers to the scenarios where the relay are located nearer to the destination terminals. On the other hand, $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 10; \sigma_{r,d}^2 = 1$ denotes the scenarios where the relays are located nearer to the source terminal. For all simulations, the relay is fixed to 1 and all channels are slow fading with $f_{D,s,r}T_s = f_{D,r,d}T_s = f_{D,s,d}T_s = 0.001$ is investigated.

In order to give a fair comparison between various combining schemes, the transmit power of the source and the relay(s) of each combiner schemes is equally divided.

4.8.1 Performance of Error Probability

Figures 4.8 and Figure 4.9 show the performances of various diversity combining schemes (i.e. direct combiner, MRC, semi MRC and SC) under symmetric case of Rayleigh fading channel. The baseline is the double differentially multiple symbol cooperative communication with conventional MRC scheme. All schemes demonstrate the almost identical curve of performance (i.e. the BER decreases as SNR increases). It is clearly shown that the proposed semi MRC and SC scheme offer better performance as compared to the conventional MRC and direct combiner scheme for a symmetric case with $\sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1; \sigma_{s,d}^2 = 1$ and slow fading channels throughout the whole SNR range. Moreover, it can be observed that the performance of the conventional MRC and direct combiner are unable to achieve optimum performance, due to the channel noise variance and the presence of the frequency offsets in the system. Both MRC and direct combining scheme of signal combining is lagging behind 8dB to 10dB and 2dB to 3dB to the optimum combining performance, respectively.

Another interesting phenomenon is that the proposed semi MRC demonstrates a very close performance with the proposed SC scheme. It is reiterated that the semi MRC scheme requires second order channel statistics, which implies that the combining scheme is not suitable for fast fading environment. On the other hand with the proposed SC scheme, the receivers are able to decode the transmitted information signal by selecting the largest magnitudes among the communication links (i.e. the relayed and direct link) without estimating the channel gains or the frequency offsets.

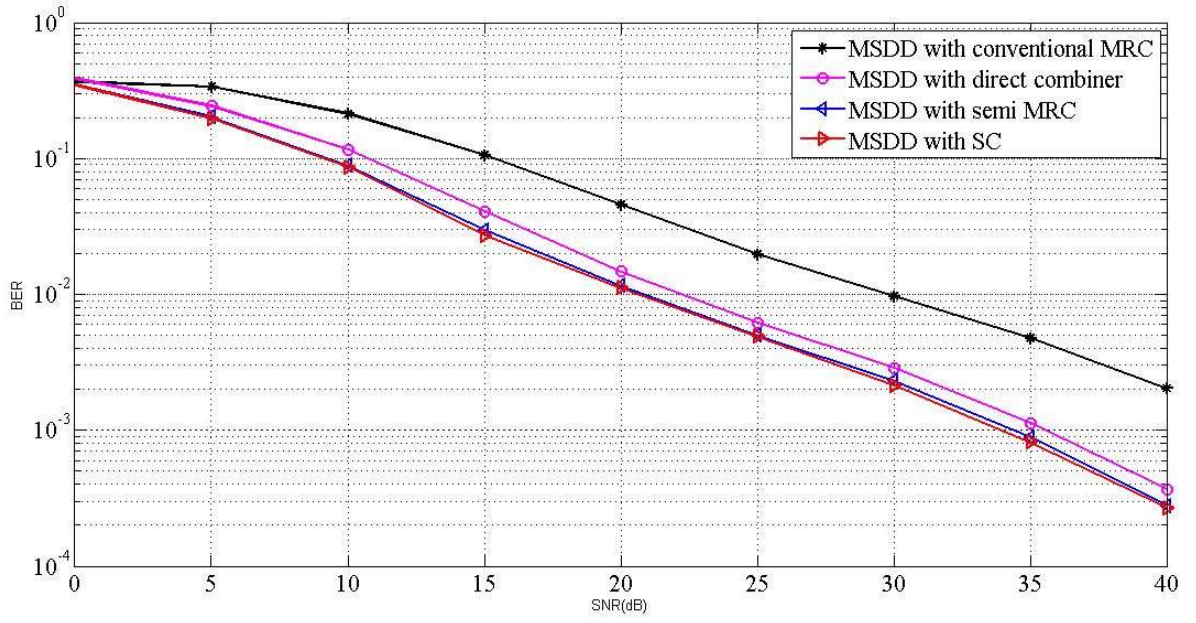


Figure 4.8: Comparison of the BER performance for the conventional MRC, semi MRC, direct combiner and proposed SC scheme based on MSDD when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$; $\sigma_{s,d}^2 = 1$ for DDBPSK

The simulations are extended to include DDQPSK in Rayleigh channels. Figure 4.9 presents results for DDQPSK in a Rayleigh channel. The shape and the relative positioning of the error probability plots are very similar to those of DDBPSK in Rayleigh fading channel as shown in Figure 4.8. The plots for the diversity combining schemes in DDQPSK are shifted at least 4 dB to the right along the SNR axis, compared to the DDBPSK case. For example to obtain the same error probability at 10^{-3} , the SC scheme requires 4 dB more SNR than DDBPSK). This occurs for all the probability plots of other diversity combining schemes.

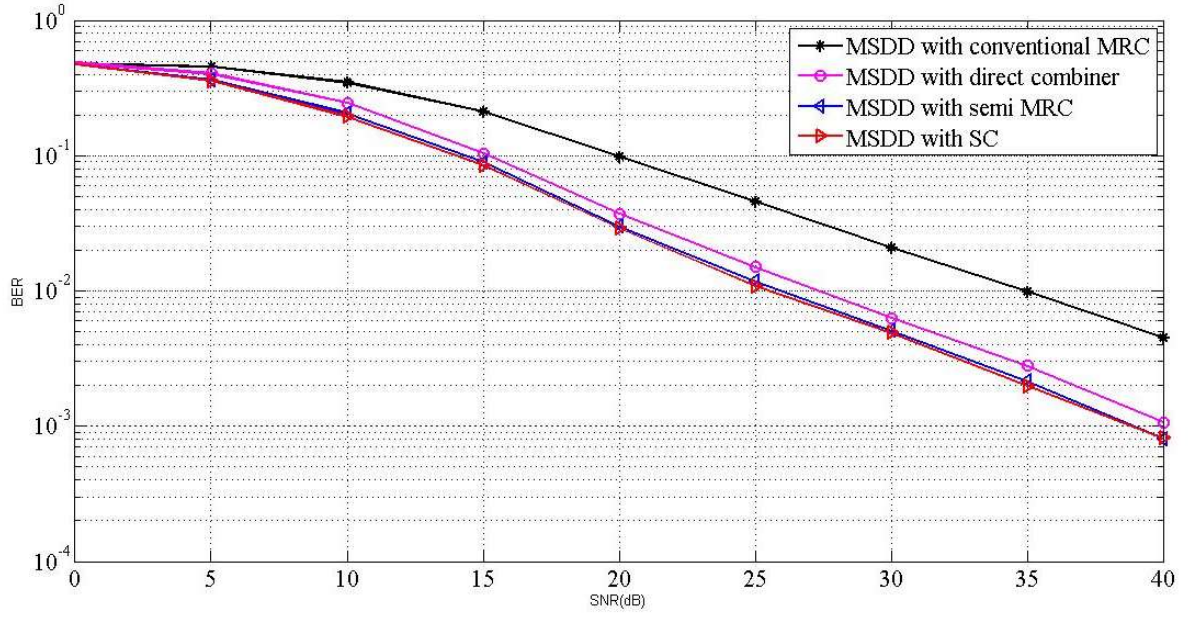


Figure 4.9: Comparison of the BER performance for the conventional MRC, semi MRC, direct combiner and proposed SC scheme based on MSDD when $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$; $\sigma_{s,d}^2 = 1$ for DDQPSK

On the other hand, the error probability plot of conventional MRC for the AF scheme is shifted even more to the right, increasing the performance difference between the SC and conventional MRC in the cooperative systems. Similar to the results of the semi MRC and SC scheme for DDBPSK, both schemes demonstrate a very close performance with each other. It can be observed that the SC scheme performance are slightly better than that of the semi MRC method. This is due to the reason that the semi MRC used the fixed combining weight based on the second-order channel statistics, instead of the instantaneous channel knowledge that are not optimum. It is noted that the conventional MRC requires the noise variance of each communication links in order to obtain the optimum combining weights. However, the noise variance of the relayed link is not available especially in fast fading channels which restrict the receiver from obtaining the perfect instantaneous channel knowledge of the relay-destination link. On the other hand, the tight results between the

semi MRC as well as SC allows the used of SC scheme analysis as the performance approximation for the semi MRC scheme under time-varying channels. It is noted that the derivation of the performance analysis for semi MRC under time-varying environment is complicated and challenging.

The SC scheme again outperforms all other schemes and has a performance difference over conventional MRC of around 5 dB at 10^{-2} when the communication channels varies with time. Both simulation results from Figure 4.8 and Figure 4.9 show that the proposed SC scheme is able to show very close performance as compare to the semi MRC scheme. But the analysis is limited to symmetric fading powers. Thus, the analysis is one step further extended for the network in the context of various terminal's location or different quality channels. These imply that the system will experience different fading powers and fading rates.

In order to calculate the allocation of the optimized power between the source and the relay, the expression of BER in (4.93) is simulated for a range of power allocation factor $q = P_S/P_T$ where $P_T = P_S + P_R$ is the total power transmitted in the system. The BER curves are plotted versus PAF in Figure 4.10 for $\frac{P}{N_0} = 20 \text{ dB}$ with DDBPSK. Similar results are obtained when DDQPSK is utilized. Note that for AF relaying network, the optimum power allocation is not modulation-dependent. This is because in the AF cooperative system, the relay amplifies the received signals and retransmit the amplified signals regardless the types of the signals. The power allocation for symmetrical and non-symmetrical cases are tabulated in Table 4.1 referring to the simulation results from Figure 4.10.

The results from Table 4.1 and Figure 4.10 demonstrate that to achieve a low BER around 10^{-3} and 10^{-4} , for strong source-destination and strong source-relay channels, more power is devoted to relay compared to the source. In contrast, when the relay-destination

channel becomes stronger than the source- relay channel, equal power allocation between the source and relay is more beneficial to enhance the system performance.

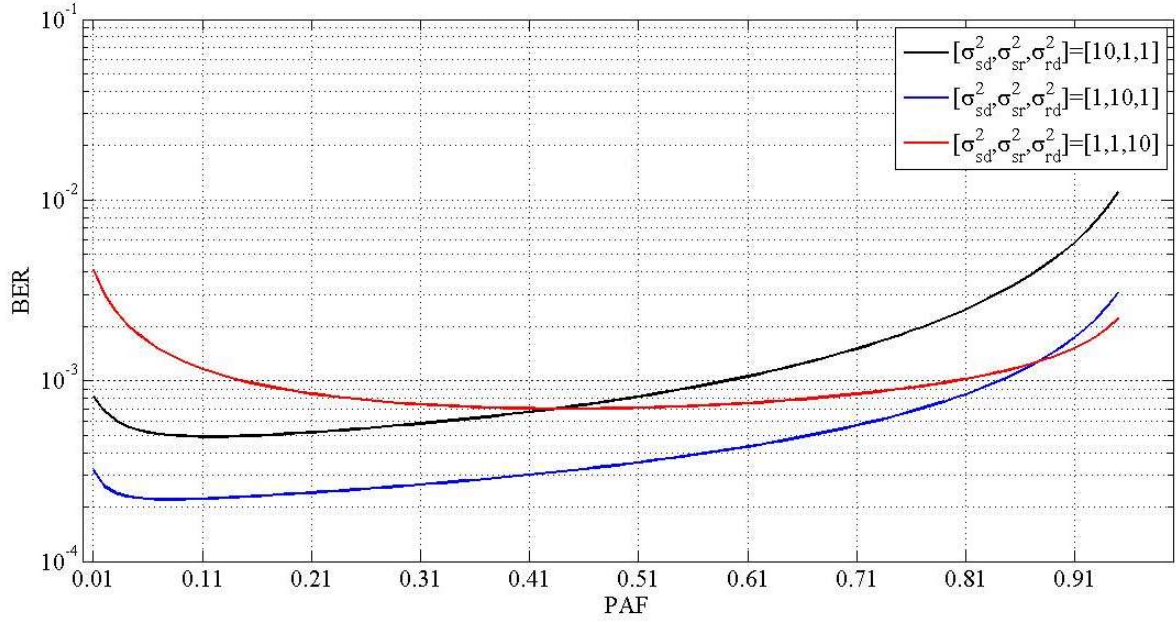


Figure 4.10: Function of BER for PAF at $\frac{P}{N_0} = 20 \text{ dB}$

Table 4.1: Power allocation for different channel variances

Scenarios	Channel Variances			Power Allocation	
	$\sigma_{s,d}^2$	$\sigma_{s,r}^2$	$\sigma_{r,d}^2$	P_S	P_R
Strong SD	10	1	1	0.08	0.92
Strong SR	1	10	1	0.11	0.89
Strong RD	1	1	10	0.44	0.56

The simulation of the SC scheme with blue solid lines (different legends) and the semi MRC scheme with red blue solid lines under time-varying channels are further

analysed. The simulations are depicted in Figure 4.11 and Figure 4.12 under different fading rates with symmetrical channels.

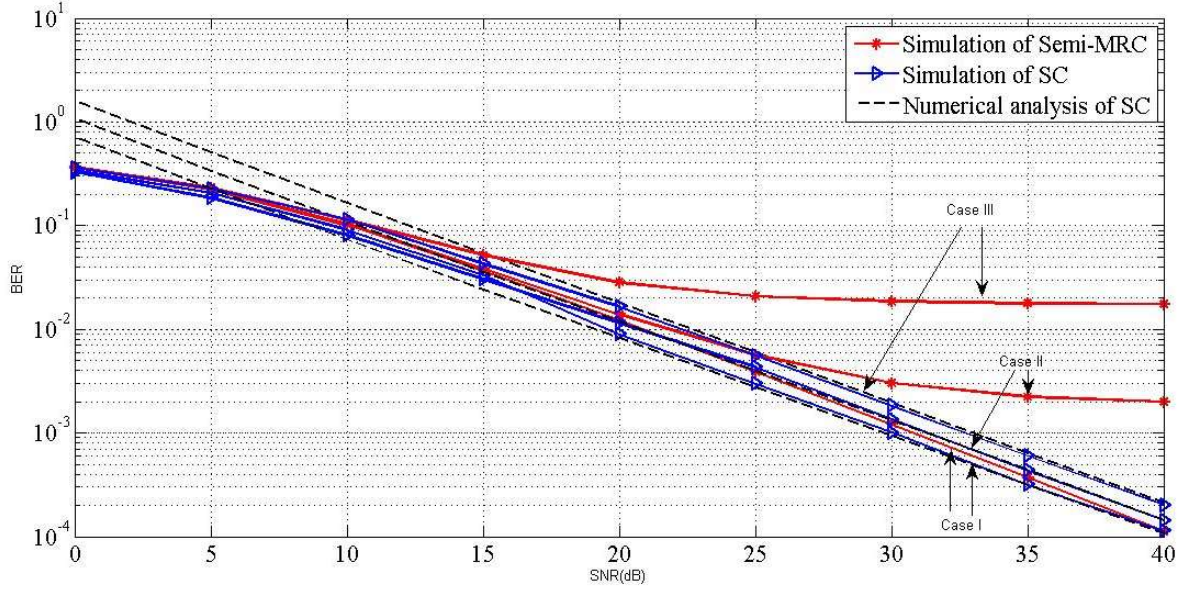


Figure 4.11: Theoretical numerical analysis and simulation BER of the DD-AF system with semi MRC and SC using BPSK under different fading rates and symmetrical channels ($\sigma_{s,d}^2 = 1$; $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 1$)

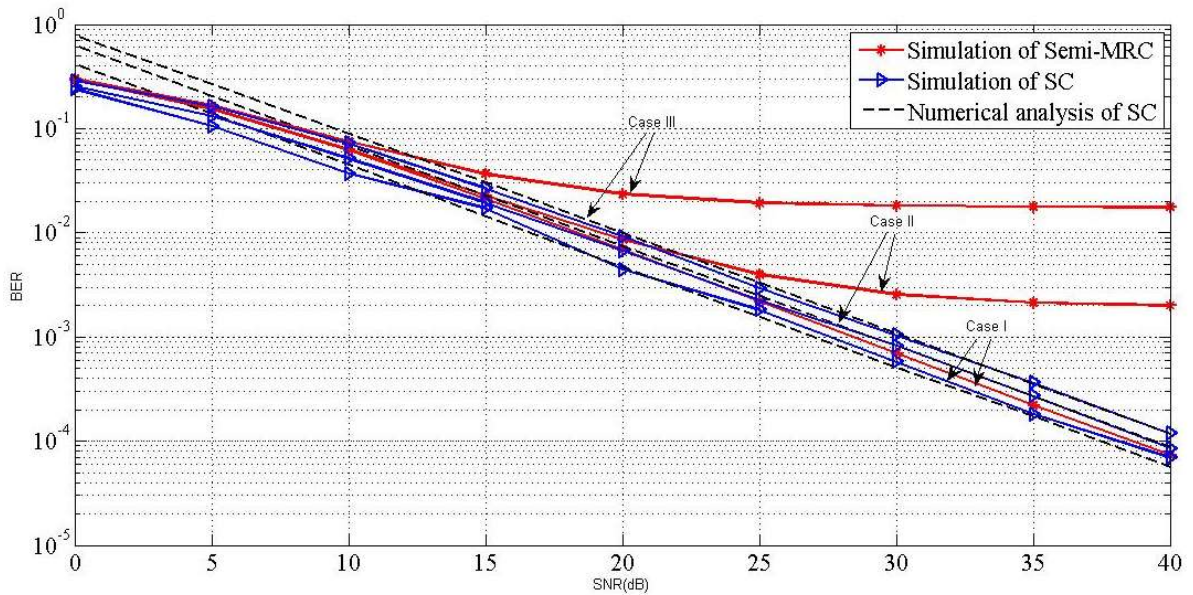


Figure 4.12: Theoretical numerical analysis and simulation BER of the DD-AF system with semi MRC and SC using BPSK under different fading rates and symmetrical channels ($\sigma_{s,d}^2 = 1$; $\sigma_{s,r}^2 = 10$; $\sigma_{r,d}^2 = 1$)

It is observed in Case I (slow fading channels) that the BER of the semi MRC and SC scheme decrease consistently with SNR and achieve diversity of two. In Case I for both schemes, error floor does not exist because all the channels involved are slow fading. Furthermore, it can be observed from Figure 4.11 and Figure 4.12, the proposed scheme benefits from substantial amount of improvement in slowly-faded channel when compared with the semi MRC scheme.

However for Case II, even more performance gain is acquired by for semi MRC scheme due to the time diversity provided by the time-varying channels. The phenomenon leads to irreducible error floor. In fast fading channels, the channels vary and the phenomenon effects both semi MRC and SC scheme for $\text{SNR} > 20\text{dB}$. In Case I, the BER for the semi MRC gradually diverge from the Case I results. To be more specific, as the SNR increases, the plot finally come to an error floor between 10^{-3} and 10^{-2} . However, the results of the SC scheme shows significant enhancement as compare to the semi MRC scheme for high SNR regime. Furthermore, the SC scheme simulation results are tight to the theoretical numerical analysis results for all fading channel cases.

Moreover, compared to the results of Case II, the BER performance of semi MRC scheme exhibits further degradation (i.e. error floor) that happened earlier at $\text{SNR} > 15\text{ dB}$ for Case III. In this case, a severe deterioration is observed for semi MRC scheme due to the scenario where all channels are fast fading, especially in high SNR regime. As for the SC scheme, the performance suffers slight degradation when compared to the results of Case II due to the effect of channel fadings. However, it can be highlighted that the SC scheme is able to reduce the effect of error floor experienced by the semi MRC scheme when

incorporated in fast fading environment. It is observed for all cases that the theoretical numerical analysis results are tight with the simulation results.

Due to the frequency offsets existence in the network, the semi MRC scheme in Case II and III demonstrate poor performance which leads to error floor. Thus, multiple-codeword of $N_C = 10$ with SC scheme is utilized to Case II and III. The BER results are exhibited in Figure 4.11 and 4.12. From the plots, it can be seen that the semi MRC scheme employed in Case I achieved good performance and due to that reason, its plot can be used as the performance benchmark to see the effectiveness of the SC scheme. Overall, it is seen from the curves that the SC scheme is more appealing in terms of the error performance. The SC scheme has the ability to produce the system performance for Case II and III near to Case I and furthermore effectively alleviate the detrimental effect of frequency offsets. Moreover, the presented scheme is computationally simple and particularly useful when channel knowledge and frequency offset is unknown in the system.

4.9 Summary

An MRC and differential SC scheme received signals of a DD AF relaying network employing DDBPSK were studied in time-varying channels at the destination. The new combining weights of MRC was provided based on the second order statistic of the channel. It was observed that the MRC experience severe degradation in the probability of error with the simulation analysis in different time-varying channels. Although the proposed combining weight outperform over the direct combining weights in Chapter 2, it requires the knowledge of fading channels. Thus, in order to avoid the estimation of channel and frequency offset at the destination, differential SC is proposed and analysed under different mobility scenarios. Error performance and system outage probability were derived and it showed that the simulation results are validated with the numerical performance analysis.

Also, the results showed that the SC performed closely to the MRC scheme which is more complicated, whereby second-order statistic of the channels are required. As discussed in Chapter 2, relays can be applied in either repetition-based or DSTC strategy. The strategy based on repetition has been applied in this chapter and the previous chapter – Chapter 4. Although repetition strategy is easier to be implemented but it suffers from low spectral efficiency. Thus, the next chapter will discuss on the detection scheme based on DSTC strategy.

CHAPTER 5

MULTIPLE CODEWORD DOUBLE DIFFERENTIAL SPHERE DECODING FOR DISTRIBUTED SPACE-TIME BLOCK CODING

5.1 Introduction

Cooperative wireless networks inherited the distributed nature which contributes towards diversity and performance gain by exploiting a group of distributed antennas in a multi-users environment. In the previous chapter 3 and chapter 4, the proposed detection and combining schemes are proposed in repetition strategy. In the repetition approach transmission, low data rate is unavoidable owing to the TDD transmission. Thus in order to solve the low data rate problem, space-time coding is deployed in a distributed manner along

with the cooperation from the relays. For the distributed STC, distinctive relays received numerous replicas of the similarly transmitted data. Then, the received signals are processed and being forwarded by the relays simultaneously towards the desired destination.

Although DSTBC is an attractive scheme, two crucial problems need to be solved. The first one involved the error propagation at the source-relay link. The error propagation causes severe degradation of the system performance if the destination is unable to decode the transmitted signals correctly. The second issue, in which need to be addressed and solved in this thesis is the synchronization problem at the destination terminals, especially in a fast fading channel causes by the presence of frequency offset. It is extremely challenging and difficult to perform an accurate decision with different copies of information signals received at the destination. Furthermore, most of the DSTC studies have been carried out and it has a rather high relay decoding complexity and the direct link between the communicating terminals is not implemented in the network. Thus, in order to alleviate the performance degradation and to restore the information transmitted from the source terminal, a decision rule is proposed based on multiple-codeword and further simplified using SD algorithm.

The chapter outline is presented as follows. In the first part of this chapter, wireless system model under DSTBC environment applied to this thesis is explained in Section 5.2. Section 5.2 also discusses and provides the methods for the two-codeword and proposed multiple codeword DD detection. Section 5.3 provides the analysis of the two-codeword and multiple codeword detection with relayed only link as well as relayed and direct link communication. The final part of the chapter which outlines the discussion of the obtained results are discussed in Section 5.4.

5.2 System Model

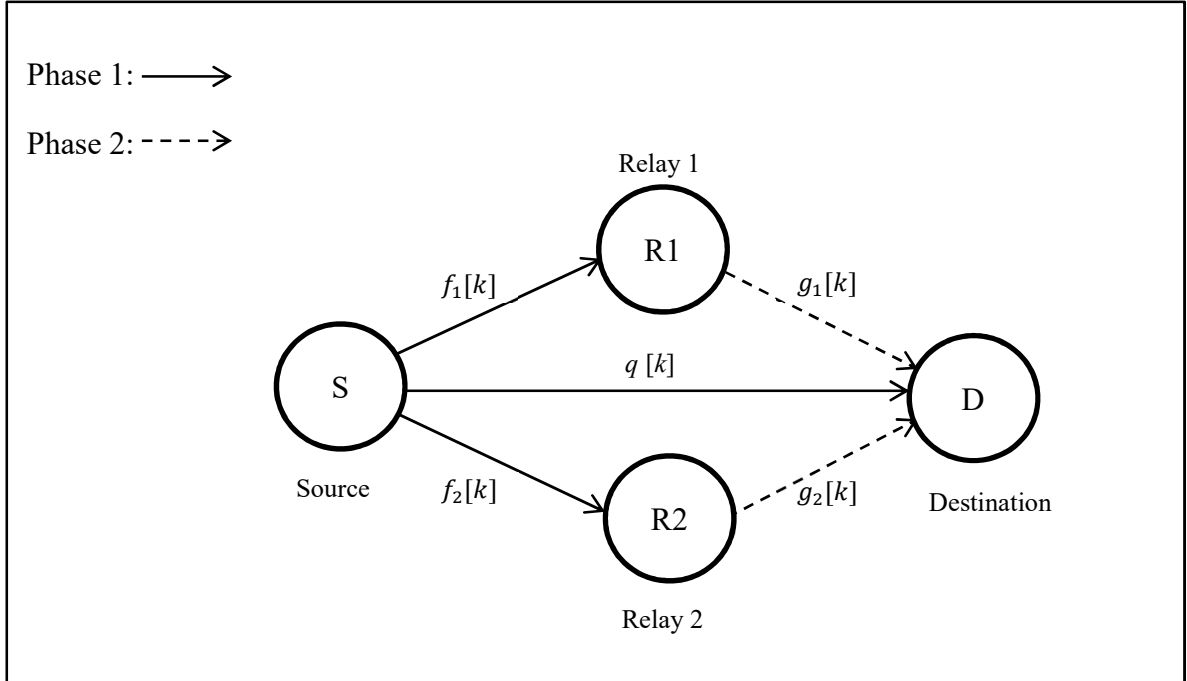


Figure 5.1: System model for distributed DD STC using AF relaying protocol

In this section, the system model from Figure 5.1 is designed following the strategy in [92] that consists of a source, two relays and a destination which allows the communicating terminals to use the direct link and relayed link via the relays to achieve a higher diversity gain. In the proposed wireless cooperative network model, a one-hop link (i.e. direct link from source to destination) together with two dual-hop links (i.e. relayed link from source to destination via relays) is considered. All the terminals are equipped with a single antenna and the communications between the terminals is in half-duplex manner. The channels of all links are assumed to be Rayleigh flat-fading (i.e. $q \sim \mathcal{CN}(0, \sigma^2)$, $f_i \sim \mathcal{CN}(0, \sigma_{s,r}^2)$, $g_i \sim \mathcal{CN}(0, \sigma_{r,d}^2)$) with $i = 1, 2$, spatially uncorrelated and changing over time due to the mobility of the terminals. To ease the notation and the analysis feasibility, it is considered that all source-relay and relay-destination channels exhibit similar channel variances, fading and frequency offsets. The channel variation is

described by the channel auto-correlation, n blocks apart between the two channels based on Jake's model [96]:

$$\varphi_{s,r}(n) = E\{f_i[k]f_i^*[k+n]\} = \sigma_{s,r}^2 J_0(2\pi f_{D_{s,r}} nR) \quad i = 1,2 \quad (5.1)$$

and

$$\varphi_{r,d}(n) = E\{g_i[k]g_i^*[k+n]\} = \sigma_{r,d}^2 J_0(2\pi f_{D_{r,d}} nR) \quad i = 1,2 \quad (5.2)$$

and

$$\varphi_{s,d}(n) = E\{q[k]q^*[k+n]\} = \sigma_{s,d}^2 J_0(2\pi f_{D_{s,d}} n) \quad (5.3)$$

where $J_0(\cdot)$ denotes the zeroth-order Bessel function of the first kind, f_i and g_i are the i th channels between the source-relay and relay-destination, correspondingly. q represents the source-destination channel. $f_{D_{s,r}}$ and $f_{D_{r,d}}$ both denotes the normalized Doppler frequency of source-relay and relay-destination channels while $f_{D_{s,d}}$ is the normalized Doppler frequency for source-destination channel. It is noted that the auto-correlation value equals 1 for fixed and static channels. On the contrary, for time-varying environment the auto-correlation value is inversely proportional with the channel fade rate (i.e. the auto-correlation value decreases in higher fade-rate channel). It is also assumed that all communication channels are perturbed with frequency offsets. Generally, the frequency offsets is caused by the mismatch between the nodes' oscillator and relative motion among the nodes. In this thesis, it is consider that the frequency offsets is due to the Doppler effects.

At the source, the information bits $\log_2 M$ are converted to symbols using M -PSK at a point that $\mathbf{v}[k] \in \mathcal{V}$ wherein $\mathcal{V}$ is $R \times R$ unitary matrices⁶ depending on the number of relays R . $\mathcal{V}$ can be written as:

⁶ A unitary matrix is an identity matrix following the multiplication process by its complex conjugate transpose matrix.

$$\mathcal{V} = \{V_l | V_l V_l^H = V_l^H V_l = I_R, l = 1, 2, \dots, L\} \quad (5.4)$$

where L represents the length of codeword.

Before the transmission process, the information symbols are transformed into codeword $v[k] \in \mathcal{V}$. The codeword are encoded with double differential modulation as follows:

$$s[k] = v[k]s[k-1]s[k-2], \quad s[0] = [1 \ 0 \ \dots \ 0]^T \quad (5.5)$$

Then the transmission is divided into two phase. For the transmission process, either codeword-by-codeword or multiple codeword can be transmitted. In both cases, the analysis is similar. During the first phase, let P_S as the average power per transmission by the source and $w \in [-\pi, \pi]$ be the frequency offset between the source and destination, which is independent of time k . It is considered that all the relays transmit at the same time during the second phase of the data transmission. The frequency offset is considered to remain static over the transmission period of at least three DD encoded matrices [20,91]. It is also considered that all users willing to cooperate without any malicious case.

The received vector at the destination and i th relays can be expressed as:

$$y_{s,d}[k] = \sqrt{P_S}s[k]e^{j2\pi f_{s,d}[k]}h_{s,d}[k] + w_{s,d}[k] \quad (5.6)$$

$$y_{s,r_i}[k] = \sqrt{P_S R}s[k]e^{j2\pi f_{s,r_i}[k]}h_{s,r_i}[k] + u_{s,r_i}[k] \quad (5.7)$$

where $w_{s,d}[k] \sim \mathcal{CN}(0, N_0 I_0)$ and $u_{s,r_i}[k] \sim \mathcal{CN}(0, N_0 I_R)$ denote the noise vector at the destination and i th relay, respectively. At the relay(s), the received vector and its conjugate is combined linearly as:

$$x_i[k] = G(A_i y_{s,r_i}[k] + B_i r_{s,r_i}^*[k]), \quad (5.8)$$

where A_i and B_i represent the combining matrices and are obtained in regards to the space-time code. Referring to Alamouti code A_i and B_i is written as:

$$\begin{aligned}
A_1 &= \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad B_1 = 0_T \\
A_2 &= 0_T, \quad B_2 = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}
\end{aligned} \tag{5.9}$$

G represents the amplifier gain at the relay that can either be constant or varies. In the environment when no channel properties and frequency offsets are known, a constant gain is employed using the variance of the source-relay channels as follows:

$$G = \sqrt{\frac{P_R}{P_S \sigma_{s,r}^2 + N_0}} \tag{5.10}$$

where P_R is the average power per transmission by the relay(s).

During the second phase of the transmission, all relay transmissions are distributed in the same channel simultaneously and are superimposed at the destination, thus the corresponding received vector at the destination is:

$$y[k] = \sum_{i=1}^R x_i[k] g_i[k] + w_i[k] \tag{5.11}$$

where $w_i[k] \sim \mathcal{CN}(0, N_0 I_R)$ is the noise vector at destination. By substituting (5.8) into (5.11) yields:

$$y[k] = G \sqrt{P_S R} S[k] e^{j2\pi f[k]} h[k] D[k] + w[k] \tag{5.12}$$

where

$$S[k] = [\hat{A}_1 \hat{s}_1 \dots \hat{A}_R \hat{s}_R] = V[k] S[k-1] S[k-2]$$

$$h[k] = [h_1[k] \dots h_R[k]]^T$$

$$D[k] = \text{diag}[1, e^{jwk}, \dots, e^{jwk[N-1]}]$$

$$w[k] = G \sum_{i=1}^R g[k] \hat{A}_i \hat{u}_i[k] + w_i[k]$$

and the distributed space-time code, the relayed channel vector and the noise vector can be expressed as:

$$\left. \begin{aligned} \hat{A}_i &= A_i, \quad h_i[k] = f_i[k]g_i[k], \\ \hat{u}_i[k] &= u_i[k], \quad \hat{s}_i[k] = s[k] \end{aligned} \right\} \text{ if } B_i = 0$$

$$\left. \begin{aligned} \hat{A}_i &= B_i, \quad h_i[k] = f_i^*[k]g_i[k], \\ \hat{u}_i[k] &= u_i^*[k], \quad \hat{s}_i[k] = s^*[k] \end{aligned} \right\} \text{ if } A_i = 0$$
(5.13)

5.2.1 Two-Codeword Double Differential Detection

In order to demonstrate the performance of the two-codeword DD detection in a time-varying environment, a relayed channel is modelled citing to the second-order autoregressive AR(2) model. In vector form, the relayed channel is written as:

$$\mathbf{h}[k] = \alpha^2 \mathbf{h}[k-2] + \alpha \mathbf{h}[k-1] + \Delta[k] \quad (5.14)$$

where

$$\begin{aligned} \Delta[k] &= \alpha_{s,r}^2 f[k-2] \sqrt{1 - \alpha_{r,d}^2} \varepsilon_{r,d}[k] + 2\alpha_{s,r} f[k-1] \sqrt{1 - \alpha_{r,d}^2} \varepsilon_{r,d}[k] \\ &+ 2\alpha_{r,d}^2 g[k-2] \sqrt{1 - \alpha_{s,r}^2} \varepsilon_{s,r}[k] + 2\alpha_{r,d} g[k-1] \sqrt{1 - \alpha_{s,r}^2} \varepsilon_{s,r}[k] \\ &+ 4\sqrt{(1 - \alpha_{s,r}^2)(1 - \alpha_{r,d}^2)} \varepsilon[k] \end{aligned} \quad (5.15)$$

$\alpha = \alpha_{s,r} \alpha_{r,d}$ is the relayed channel equivalent auto-correlation, $h[k-2]$ and $h[k-1]$ denote the relayed channel for $k-2$ and $k-1$, accordingly.

In reference to second-order autoregressive AR(2) model using the source-destination and relayed channel models developed from Section 3.2.2 and substitutes equation (3.15) and (3.19) into (5.6) and (5.12), accordingly, gives:

$$\mathbf{y}_{s,d}[k] = \alpha_{s,d}^2 y[k-1] \mathbf{y}_{s,d}[k-2] + \alpha_{s,d} \mathbf{V}[k] \mathbf{y}_{s,d}[k-1] + n_{s,d}[k] \quad (5.16)$$

$$\begin{aligned}\mathbf{n}_{s,d}[k] &= \mathbf{w}_{s,d}[k] - \alpha_{s,d}^2 \mathbf{y}[k-1] \mathbf{w}_{s,d}[k-1] - \alpha_{s,d} \mathbf{V}[k] \mathbf{w}_{s,d}[k-2] \\ &\quad + 2\sqrt{1 - \alpha_{s,d}^2} \sqrt{P_S} \mathbf{z}[k] \varepsilon_{s,d}[k]\end{aligned}$$

and

$$\mathbf{y}_{r,i,d}[k] = \alpha_{r,i,d}^2 \mathbf{y}[k-1] \mathbf{y}_{r,i,d}[k-2] + \alpha_{r,i,d} \mathbf{V}[k] \mathbf{y}_{r,i,d}[k-1] + \mathbf{n}_{r,i,d}[k], \quad i = 1, 2 \quad (5.17)$$

where

$$\begin{aligned}\mathbf{n}_{r,i,d}[k] &= \mathbf{w}_{r,i,d}[k] - \alpha_{r,i,d}^2 \mathbf{y}[k-1] \mathbf{w}_{r,i,d}[k-1] - \alpha_{r,i,d} \mathbf{V}[k] \mathbf{w}_{r,i,d}[k-2] \\ &\quad + 2G\sqrt{1 - \alpha_{r,i,d}^2} \sqrt{P_S} \mathbf{z}[k] \mathbf{h}_{r,i,d}[k] \varepsilon_{s,r,i}[k]\end{aligned}$$

It is noted from the equivalent noise power $\mathbf{n}[k]$ generated, the term is added by additional parameters, equivalent channels' autocorrelation, the source power and channel coefficient of relay-destination channel. Due to the additional parameters in the equivalent noise power, it can be observed that the system experiences larger degradation in a fast fading environment as compared to slow fading channels.

5.2.2 Multiple Codeword Double Differential Detection

Multiple codeword DD detection is proposed to circumvent the limitation of two-codeword DD detection under fast fading environments.

The received symbols at the destination can be expressed as:

$$\begin{aligned}\bar{\mathbf{y}}_{r,d}[k] &= G\sqrt{P_S R} \mathbf{s}[k] \mathbf{h}[k] + \mathbf{w}[k] \\ &= G\sqrt{P_S R} \text{diag}\{\bar{\mathbf{s}}\} \text{diag}\{\bar{\mathbf{h}}_{r,d}\} \bar{\mathbf{h}}_{s,r} + \bar{\mathbf{w}}_{r,d}\end{aligned} \quad (5.18)$$

and a one hop (i.e. direct path from source to destination):

$$\bar{\mathbf{y}}_{s,d}[k] = \sqrt{P_S} \mathbf{s}[k] \mathbf{h}[k] + \mathbf{w}[k] \quad (5.19)$$

$$= \sqrt{P_S} \text{diag}\{\bar{\mathbf{s}}\} \text{diag}\{\bar{\mathbf{h}}_{s,d}\} + \bar{\mathbf{w}}_{s,d}$$

where

$$\bar{\mathbf{s}} = \mathbf{s}[1], \mathbf{s}[2], \dots, \mathbf{s}[N]$$

$$\bar{\mathbf{y}}_{r,d} = [\mathbf{y}_{r,d}^T[1], \mathbf{y}_{r,d}^T[2], \dots, \mathbf{y}_{r,d}^T[N]]^T$$

$$\bar{\mathbf{y}}_{s,d} = [\mathbf{y}_{s,d}^T[1], \mathbf{y}_{s,d}^T[2], \dots, \mathbf{y}_{s,d}^T[N]]^T$$

$$\bar{\mathbf{h}}_{r,d} = [\mathbf{h}_{r,d}[1], \mathbf{h}_{r,d}[2], \dots, \mathbf{h}_{r,d}[N]]^T$$

$$\bar{\mathbf{h}}_{s,r} = [\mathbf{h}_{s,r}^T[1], \mathbf{h}_{s,r}^T[2], \dots, \mathbf{h}_{s,r}^T[N]]^T$$

$$\bar{\mathbf{h}}_{s,d} = [\mathbf{h}_{s,d}[1], \mathbf{h}_{s,d}[2], \dots, \mathbf{h}_{s,d}[N]]^T$$

$$\bar{\mathbf{w}}_{r,d} = [\mathbf{w}_{r,d}^T[1], \mathbf{w}_{r,d}^T[2], \dots, \mathbf{w}_{r,d}^T[N]]^T$$

$$\bar{\mathbf{w}}_{s,d} = [\mathbf{w}_{s,d}^T[1], \mathbf{w}_{s,d}^T[2], \dots, \mathbf{w}_{s,d}^T[N]]^T$$

$\mathbf{s}$ is a unitary block diagonal matrix ($\mathbf{s}^H \mathbf{s} = \mathbf{s} \mathbf{s}^H = \mathbf{I}_N$) that consists of N codewords equivalent to $N - 2$ codewords received in:

$$\bar{\mathbf{V}} = \text{diag}\{\mathbf{V}[1], \dots, \mathbf{V}[N - 1], \mathbf{V}[N - 2]\} \quad (5.20)$$

where $\mathbf{s}[n + 2] = \mathbf{s}[n - 1]\mathbf{s}[n]\mathbf{V}[n]$ and $\mathbf{s}[n] = \mathbf{I}_R$ is set as initial symbol.

Therefore, conditioned on $\bar{\mathbf{s}}$, $\bar{\mathbf{y}}_{s,d}$ is a circularly-symmetric Gaussian random variable with its PDF as:

$$P(\bar{\mathbf{y}}_{s,d} | \bar{\mathbf{s}}) = \frac{1}{\pi^N \det\{\boldsymbol{\Sigma}_{\bar{\mathbf{y}}_{s,d}}\}} \exp(-\bar{\mathbf{y}}_{s,d}^H \boldsymbol{\Sigma}_{\bar{\mathbf{y}}_{s,d}}^{-1} \bar{\mathbf{y}}_{s,d}) \quad (5.21)$$

Similarly, conditioned on $\bar{\mathbf{s}}$ and $\bar{\mathbf{h}}_{r,d}$, $\bar{\mathbf{y}}_{r,d}$ is a circularly-symmetric Gaussian random variable and its PDF can be formulated as:

$$P(\bar{\mathbf{y}}_{r,d} | \bar{\mathbf{s}}, \bar{\mathbf{h}}_{r,d}) = \frac{1}{\pi^N \det\{\boldsymbol{\Sigma}_{\bar{\mathbf{y}}_{r,d}}\}} \exp(-\bar{\mathbf{y}}_{r,d}^H \boldsymbol{\Sigma}_{\bar{\mathbf{y}}_{r,d}}^{-1} \bar{\mathbf{y}}_{r,d}) \quad (5.22)$$

In (5.21) and (5.22), the matrices $\boldsymbol{\Sigma}_{\bar{\mathbf{y}}_{s,d}}$ and $\boldsymbol{\Sigma}_{\bar{\mathbf{y}}_{r,d}}$ are conditional covariance matrices of $\bar{\mathbf{y}}_{s,d}$ and $\bar{\mathbf{y}}_{r,d}$, defined as:

$$\begin{aligned}
\mathbf{\Sigma}_{\bar{\mathbf{y}}_{s,d}} &= E\{\bar{\mathbf{y}}_{s,d} \bar{\mathbf{y}}_{s,d}^H | \bar{\mathbf{s}}\} = P_S \text{diag}\{\bar{\mathbf{s}}\} \mathbf{\Sigma}_{\bar{\mathbf{h}}_{s,d}} \text{diag}\{\bar{\mathbf{s}}^*\} + N_0 \mathbf{I}_N \\
&= \text{diag}\{\bar{\mathbf{s}}\} \left(P_S \mathbf{\Sigma}_{\bar{\mathbf{h}}_{s,d}} + N_0 \mathbf{I}_N \right) \text{diag}\{\bar{\mathbf{s}}^*\}
\end{aligned} \tag{5.23}$$

and

$$\begin{aligned}
\mathbf{\Sigma}_{\bar{\mathbf{y}}_{r,d}} &= E\{\bar{\mathbf{y}}_{r,d} \bar{\mathbf{y}}_{r,d}^H | \bar{\mathbf{s}}, \bar{\mathbf{h}}_{r,d}\} \\
&= G^2 (P_S R) \text{diag}\{\bar{\mathbf{s}}\} \text{diag}\{\bar{\mathbf{h}}_{r,d}\} \mathbf{\Sigma}_{\bar{\mathbf{h}}_{s,r}} \text{diag}\{\bar{\mathbf{h}}_{r,d}^*\} \text{diag}\{\bar{\mathbf{s}}^*\} + \mathbf{\Sigma}_{\bar{\mathbf{w}}}
\end{aligned} \tag{5.24}$$

where the covariance matrices of $\bar{\mathbf{h}}_{s,d}$, $\bar{\mathbf{h}}_{s,r}$ and $\bar{\mathbf{w}}$ is defined, accordingly as follows:

$$\mathbf{\Sigma}_{\bar{\mathbf{h}}_{s,d}} = E\{\bar{\mathbf{h}}_{s,d} \bar{\mathbf{h}}_{s,d}^H\} = \mathbf{C}_{\bar{\mathbf{h}}_{s,d}} \otimes \mathbf{I}_R \tag{5.25}$$

$$\mathbf{C}_{\bar{\mathbf{h}}_{s,d}} = \text{toeplitz}\{\varphi_{s,d}(0), \dots, \varphi_{s,d}(N-2)\} \tag{5.26}$$

and

$$\mathbf{\Sigma}_{\bar{\mathbf{h}}_{s,r}} = E\{\bar{\mathbf{h}}_{s,r} \bar{\mathbf{h}}_{s,r}^H\} = \mathbf{C}_{\bar{\mathbf{h}}_{s,r}} \otimes \mathbf{I}_R \tag{5.27}$$

$$\mathbf{C}_{\bar{\mathbf{h}}_{s,r}} = \text{toeplitz}\{\varphi_{s,r}(0), \dots, \varphi_{s,r}(N-2)\} \tag{5.28}$$

and

$$\mathbf{\Sigma}_{\bar{\mathbf{w}}} = E\{\bar{\mathbf{w}} \bar{\mathbf{w}}^H\} = \mathbf{C}_{\bar{\mathbf{w}}} \otimes \mathbf{I}_R \tag{5.29}$$

$$\mathbf{C}_{\bar{\mathbf{w}}} = N_0 \text{diag}\{(1 + G^2 \sum_{i=1}^R |\bar{\mathbf{h}}_{r,d_i}[1]|^2), \dots, (1 + G^2 \sum_{i=1}^R |\bar{\mathbf{h}}_{r,d_i}[N]|^2)\} \tag{5.30}$$

Given (5.21) and (5.22), the non-coherent ML detection can be expressed as:

$$\begin{aligned}
\hat{\mathbf{s}} &= \arg \max_{\mathbf{s} \in M^{N-2}} \left\{ \frac{1}{\pi^N \det\{\mathbf{\Sigma}_{\bar{\mathbf{y}}_{s,d}}\}} \exp(-\bar{\mathbf{y}}_{s,d}^H \mathbf{\Sigma}_{\bar{\mathbf{y}}_{s,d}}^{-1} \bar{\mathbf{y}}_{s,d}) \right\} \\
&\times E_{\bar{\mathbf{h}}_{r,d}} \left\{ \frac{1}{\pi^N \det\{\mathbf{\Sigma}_{\bar{\mathbf{y}}_{r,d}}\}} \exp(-\bar{\mathbf{y}}_{r,d}^H \mathbf{\Sigma}_{\bar{\mathbf{y}}_{r,d}}^{-1} \bar{\mathbf{y}}_{r,d}) \right\}
\end{aligned} \tag{5.31}$$

As can be seen from (5.31), the ML metric requires the expectation over the distribution of $\bar{\mathbf{h}}_{r,d}$, which does not produce a closed-form expression. Thus, it is proposed to use the following modified decision metric:

$$\begin{aligned} \hat{\mathbf{s}} = \arg \max_{\mathbf{s} \in M^{N-2}} \{ & \frac{1}{\pi^N \det\{\mathbf{\Sigma}_{\bar{\mathbf{y}}_{s,d}}\}} \exp(-\bar{\mathbf{y}}_{s,d}^H \mathbf{\Sigma}_{\bar{\mathbf{y}}_{s,d}}^{-1} \bar{\mathbf{y}}_{s,d}) \} \\ & \times \left\{ \frac{1}{\pi^N \det\{\bar{\mathbf{\Sigma}}_{\bar{\mathbf{y}}_{r,d}}\}} \exp\left(-\bar{\mathbf{y}}_{r,d}^H \bar{\mathbf{\Sigma}}_{\bar{\mathbf{y}}_{r,d}}^{-1} \bar{\mathbf{y}}_{r,d}\right) \right\} \end{aligned} \quad (5.32)$$

where

$$\begin{aligned} \bar{\mathbf{\Sigma}}_{\bar{\mathbf{y}}_{r,d}} &= \mathbb{E}_{\bar{\mathbf{h}}_{r,d}} \{ \mathbf{\Sigma}_{\bar{\mathbf{y}}_{r,d}} \} = G^2 P_S R \bar{\mathbf{S}} (\mathbf{C}_{\bar{\mathbf{h}}} \otimes \mathbf{I}_R) \bar{\mathbf{S}}^H + (1 + G^2 \sigma_{r,d}^2 R) N_0 (\mathbf{I}_N \otimes \mathbf{I}_R) \\ &= \bar{\mathbf{S}} (\mathbf{C} \otimes \mathbf{I}_R) \bar{\mathbf{S}}^H \end{aligned} \quad (5.33)$$

where

$$\mathbf{C} = G^2 P_S R \bar{\mathbf{S}} \mathbf{C}_{\bar{\mathbf{h}}} + N_0 (1 + G^2 \sigma_{r,d}^2 R) \mathbf{I}_N \quad (5.34)$$

$$\mathbf{C}_{\bar{\mathbf{h}}} = \text{toeplitz}\{\varphi_{s,r}(0)\varphi_{r,d}(0), \dots, \varphi_{s,r}(N-2)\varphi_{r,d}(N-2)\} \quad (5.35)$$

By employing the rule $\det\{\mathbf{AB}\} = \det\{\mathbf{BA}\}$, the determinant in (5.32) is no longer dependent

to $\bar{\mathbf{s}}$ and the modified decision matrix can be simplified as:

$$\begin{aligned} \hat{\mathbf{s}} &= \arg \min_{\bar{\mathbf{s}} \in M^{N-2}} \{ \bar{\mathbf{y}}_{r,d}^H \mathbf{\Sigma}_{\bar{\mathbf{y}}_{r,d}}^{-1} \bar{\mathbf{y}}_{r,d} + \bar{\mathbf{y}}_{s,d}^H \mathbf{\Sigma}_{\bar{\mathbf{y}}_{s,d}}^{-1} \bar{\mathbf{y}}_{s,d} \} \\ &= \arg \min_{\bar{\mathbf{s}} \in M^{N-2}} \{ \bar{\mathbf{y}}_{r,d}^H \text{diag}\{\bar{\mathbf{s}}\} (\mathbf{C}^{-1} \otimes \mathbf{I}_R) \text{diag}\{\bar{\mathbf{s}}\}^* \bar{\mathbf{y}}_{r,d} \\ &\quad + \bar{\mathbf{y}}_{s,d}^H \text{diag}\{\bar{\mathbf{s}}\} (\mathbf{C}_{s,d}^{-1} \otimes \mathbf{I}_R) \text{diag}\{\bar{\mathbf{s}}\}^* \bar{\mathbf{y}}_{s,d} \} \\ &= \arg \min_{\bar{\mathbf{s}} \in M^{N-2}} \{ \bar{\mathbf{y}}_{r,d}^H \text{diag}\{\bar{\mathbf{s}}\} (\mathbf{U}_{r,d}^H \otimes \mathbf{I}_R) (\mathbf{U}_{r,d} \otimes \mathbf{I}_R) \text{diag}\{\bar{\mathbf{s}}\}^* \bar{\mathbf{y}}_{r,d} \\ &\quad + \bar{\mathbf{y}}_{s,d}^H \text{diag}\{\bar{\mathbf{s}}\} (\mathbf{U}_{s,d}^H \otimes \mathbf{I}_R) (\mathbf{U}_{s,d} \otimes \mathbf{I}_R) \text{diag}\{\bar{\mathbf{s}}\}^* \bar{\mathbf{y}}_{s,d} \} \\ &= \arg \min_{\bar{\mathbf{s}} \in M^{N-2}} \{ \|\mathbf{U} \mathbf{s}_{r,d}\|^2 + \|\mathbf{U} \mathbf{s}_{s,d}\|^2 \} = \arg \min_{\bar{\mathbf{s}} \in M^{N-2}} \{ \|\mathbf{U} \bar{\mathbf{s}}\|^2 \} \end{aligned} \quad (5.36)$$

where $\mathbf{U}$ represents the upper triangular matrix obtained using the Cholesky decomposition

of $\mathbf{C}^{-1} = \mathbf{U}^H \mathbf{U}$ and thus, the new matrices $\mathbf{U}_{\bar{\mathbf{s}}}$ is defined as:

$$\mathbf{U}_{\bar{\mathbf{s}}} = \begin{bmatrix} \sum_{j=1}^N u_{\bar{\mathbf{s}}1,j} \bar{\mathbf{s}}^H[j] \bar{\mathbf{y}}[j] \\ \sum_{j=2}^N u_{\bar{\mathbf{s}}2,j} \bar{\mathbf{s}}^H[j] \bar{\mathbf{y}}[j] \\ \vdots \\ u_{N,N} \bar{\mathbf{s}}^H[N] \bar{\mathbf{y}}[N] \end{bmatrix} \quad (4.37)$$

where $\mathbf{U}_{\bar{\mathbf{s}}_{i,j}} = [\mathbf{U}_{\mathbf{s},\mathbf{d}_{i,j}}, \mathbf{U}_{\mathbf{r},\mathbf{d}_{i,j}}]^T$ is the specific vector component for i th row and j th column.

$\mathbf{U}_{\mathbf{s},\mathbf{d}_{i,j}}$ and $\mathbf{U}_{\mathbf{r},\mathbf{d}_{i,j}}$ denote the entry of $\mathbf{U}_{\mathbf{s},\mathbf{d}}$ and $\mathbf{U}_{\mathbf{r},\mathbf{d}}$ in row i and column j , respectively.

It is noted that the last variable of vector $u_{\bar{\mathbf{s}}}$ does not possess any influence on the minimization as $\mathbf{s}[N] = \mathbf{I}_R$. Thus, the variable may be neglected. By substituting $\bar{\mathbf{s}}^H[n] = \bar{\mathbf{s}}^H[n+2] \bar{\mathbf{s}}^H[n+1] \mathbf{V}_{\mathbf{n}}$ into (5.34), it follows that:

$$\begin{aligned} \hat{\mathbf{V}} = \arg \min_{\hat{\mathbf{s}} \in M^N} & \left\{ \sum_{n=1}^{N-2} \left\| u_{n,n} \mathbf{V}[n] \bar{\mathbf{y}}[n] \right. \right. \\ & \left. \left. + \bar{\mathbf{s}}[n+1] \sum_{j=n+1}^{N-1} u_{n,j} \bar{\mathbf{s}}^H[j] \bar{\mathbf{y}}[j] + \bar{\mathbf{s}}[n+2] \sum_{j=n+2}^N u_{n,j} \bar{\mathbf{s}}^H[j] \bar{\mathbf{y}}[j] \right\|^2 \right\} \end{aligned} \quad (5.38)$$

Then, the minimization can be computed with low complexity using the SD algorithm to find $N - 2$ codewords. The MCDD employing SD algorithm implemented for the wireless DSTBC cooperative communication can be written as follows:

Algorithm 2: MCDDSD

Input: $\sigma_{s,d}^2, \sigma_{s,r}^2, \sigma_{r,d}^2, f_{s,d}, f_{s,r}, f_{r,d}, R, P_S, N_0, N, \mathbf{y}$

Output: $\hat{\mathbf{V}}[k], \quad k = 1, \dots, N - 1$

1. Compute $\mathbf{C}_{\bar{\mathbf{h}}_{s,d}}$ of (5.26)
2. Compute $\mathbf{C}_{\bar{\mathbf{h}}}$ of (5.34)
3. Compute $\mathbf{U}_{\bar{\mathbf{s}}}$ of $\mathbf{C}^{-1} = \mathbf{U}_{\bar{\mathbf{s}}} \mathbf{U}_{\bar{\mathbf{s}}}^H$
4. Call function $\hat{\mathbf{V}} = MCDDSD(\mathbf{U}_{r,d}, \mathbf{U}_{s,d}, \mathbf{y})$ [30]

5.3 Discussion of Simulation Results for Multiple Codeword Double Differential Sphere Decoding under Distributed Space-Time Block Coding (DSTBC)

In this section, the performance of BER for the proposed MCDDSD detector is thoroughly evaluated and analyzed in a one-way AF relaying assisted DSTBC environment. In order to support the development and analysis of the proposed scheme, a wireless relayed network employing a source, two relays and a destination terminal is designed and analysed under various fading environments. The relays employ distributed Alamouti code and the AF relaying protocol for the transmission. To evaluate the BER of the system, the binary data at the source is encoded with M -PSK constellation and then converted to unitary matrices with Alamouti structure, that is, the data set is designed as [88]:

$$\mathbf{u} = \left\{ \frac{1}{\sqrt{|U_1|^2 + |U_2|^2}} \begin{bmatrix} U_1 & -\overline{U_2} \\ U_2 & \overline{U_1} \end{bmatrix} \middle| U_1 \in \mathcal{F}_1, U_2 \in \mathcal{F}_2 \right\} \quad (5.39)$$

$U_1 \in \mathcal{F}_1$ and $U_2 \in \mathcal{F}_2$ are information symbols modulated with PSK signals. $\mathcal{F}_1$ and $\mathcal{F}_2$ are finite sets. Then, the unitary matrix is double differentially encoded in the system to omit the needs of channel and frequency offset knowledge at the relay and destination. All the communication channels are i.i.d based on the algorithm in [94]. The algorithm uses SoS method for simulation in order to create time-correlated Rayleigh-faded channel coefficients. Following the approach in (3.7), the channels coefficients $\{f_i[k]\}, \{g_i[k], i = 1, \dots, R$ and $q[k]$ generated independently can be expressed as:

$$f(k) = f_i(k) + jf_q(k) \quad (5.40)$$

$$f_i(k) = \sqrt{\frac{2}{LM}} \sum_{l,m=1}^{L,M} \cos\{2\pi f_{s,r} \cos(\alpha_l) k + 2\pi f_{s,r} \cos(\beta_m) k + \phi_{lm}^i\} \quad (5.41)$$

$$f_q(k) = \sqrt{\frac{2}{LM}} \sum_{l,m=1}^{L,M} \cos\{2\pi f_{s,r} \cos(\alpha_l) k + 2\pi f_{s,r} \cos(\beta_m) k + \phi_{lm}^q\} \quad (5.42)$$

where

$$\alpha_l = \frac{2\pi l - \pi + \theta}{4L}$$

$$\beta_m = 2\left(\frac{2\pi m - \pi + \varphi}{4M}\right)$$

Similarly, $g_i[k]$ and $q[k]$ channel coefficients are generated except that $f_{s,r}$ from (5.40) and (5.41) are replaced with $f_{r,d}$ and $f_{s,d}$. It is noted that the channel coefficients are dependent on the channels normalized frequency Doppler, which represents the users mobility. It can be stated that the fading rate is directly proportional to the speed, such that higher speed causes higher channel fading rate. Thus, by changing the value of normalized Doppler frequency, different fading cases (i.e. slow and fast fading) can be generated and simulated. The fading channels are perturbed with frequency offsets. Throughout the simulations, all the frequency offsets are mutually independent and uniformly distributed throughout $[-\pi, \pi]$.

At the destination, conventional two and multiple codeword based detection are utilized. It is confirmed by the simulation results that the proposed multiple codeword detections is able to outperform the conventional two codeword scheme when communicating over one-way relay channels. Moreover, the system shows enhancement when the length of the transmitted block, $N_C = 10$ for multiple codeword detection schemes is used. As the length of the transmitted block increased, the search for the originally

transmitted data signals become tedious. Thus, a SD strategy is proposed so that any sophisticated searching process in the destination is no longer required. The transmission is assumed over unidirectional Rayleigh time-varying channels, whereby the magnitude of the fading channels remains near-constant within one block but varies from block to block.

5.3.1 Performance of Two-Codeword and Multiple-Codeword in Dual-Hop Transmission Only and Dual-Hop Transmission with Direct Link

Matlab simulation used the BER to examine the overall system performance for the proposed DD cooperative communications in the presence of frequency offsets. The flat fading Rayleigh channels are perturbed with frequency offsets set at $[-\pi, \pi]$.

To motivate the design of multiple-codeword over two-codeword, the performance of MCDDSD is evaluated under two scenarios. The first scenario involves the dual-hop (i.e. source-relay-destination without a direct link) while the second scenario considers both dual-hop and direct (i.e. source-destination) link. For comparison purposes, the simulation is carried out over symmetrical channel variance values. All the channels experienced slightly fast fading scenario when the normalized Doppler frequency is 0.05.

In the fairly fast fading environment, the channels are assumed to be constant for three consecutive blocks. It is worth noting that in order to make a fair evaluation, the power used for the transmission of DDBPSK and DDQPSK signals at the source terminals and the power transmission of DSTBC codewords from the relay terminals is set such that $P_S = P_T/2$ and $P_R = P_T/4$ where P_T is the total transmitted power in the system.

Figure 5.2 compares the simulated error probability between the conventional two-codeword as well as multiple codeword detection in distributed DDSTBC. The normalized Doppler frequency at each link is $f_{D_{s,r}}T_s = f_{D_{r,d}}T_s = f_{D_{s,d}}T_s = 0.05$. Also, it is considered that the channel variances are $\sigma_{s,r}^2 = 1, \sigma_{r,d}^2 = \sigma_{s,d}^2 = 1$.

It is shown that under the effect of frequency offsets, the conventional two-codeword scheme suffers severe performance degradation for both scenarios of dual-hop link network only and dual-hop link with direct link network. This degradation leads to reducible error floor when the SNR increases. On the contrary, as illustrated in Figure 5.2 the multiple codewords shares the same slope with the conventional two-codeword scheme under similar scenarios. Both two-codeword and multiple-codeword error performance in the two scenarios decreases with respect to the SNR.

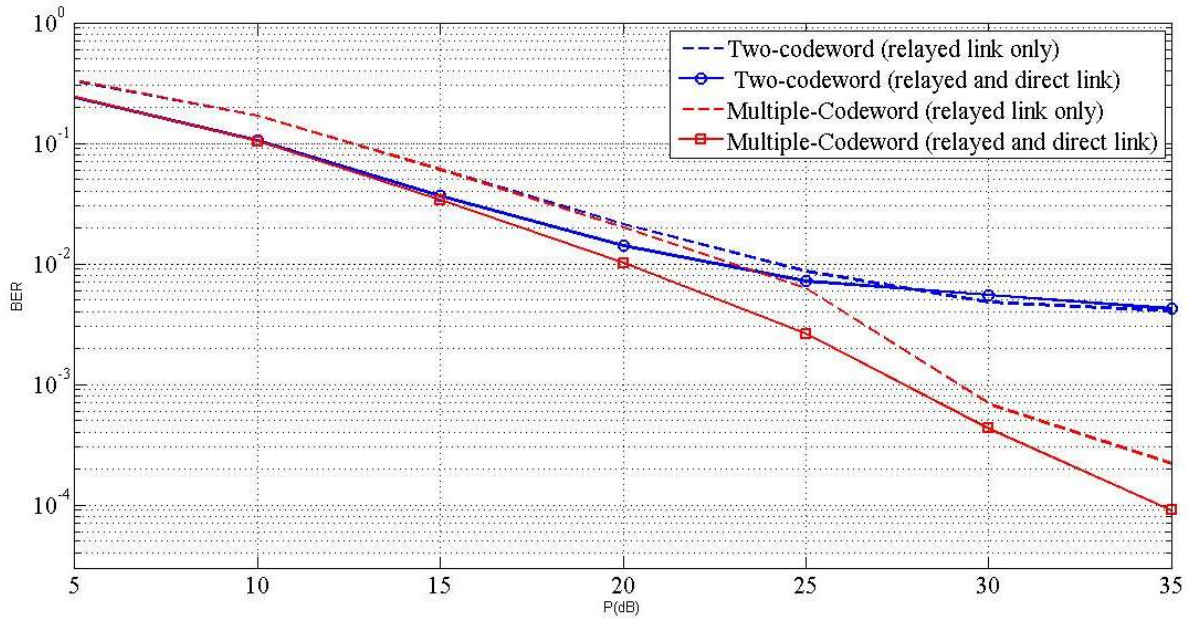


Figure 5.2: Comparison of the BER performance for the conventional DD and proposed MCDDSD scheme when $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1$

It is clearly showed that the plots of the proposed MCDDSD scheme outperform the conventional two-codeword scheme by approximately 5.0 dB at BER 10^{-3} in the case of

dual-hop link network without a direct link. Compared to the dual-hop only two-codeword network transmission, the performance of the proposed scheme in a dual-hop link including the source-destination channel shows further improvement of approximately 8.0 dB at BER 10^{-3} . This performance gain indicates that the low-complexity proposed scheme provides an efficient solution to preserving the orthogonality of the distributed space-time coded signals when transmitted by cooperating relay node and at the same enhance the error performance of the network.

Moreover, higher diversity order is obtained when a direct link is included in the relayed network, which outperforms the performance in the scenario of a dual-hop link network. Clearly, it is shown that cooperative diversity is achieved for the distributed STC and the proposed scheme allows the communicating terminals to use the direct link between them in order to produce better-relayed network performance.

Also, investigation of BER performance is made when the proposed scheme is employed in system with various channel qualities $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 10; \sigma_{r,d}^2 = 1$ with strong source-relay channel and $\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 10$ with strong relay-destination channel. The simulation results are depicted in Figure 5.3 and 5.4, respectively. It is shown in Figure 5.3 that the proposed MCDDSD scheme under fairly fast fading channels is also able of achieving second order transmit diversity. It can be observed that, to be specific, the performance of MCDDSD is approximately 3.0 dB better than in the previous scenario at BER 10^{-3} in strong source-relay channel. Likewise, it is also worth noting that, the proposed MCDDSD scheme is about 3.0 dB superior than the conventional two-codeword scheme at BER 10^{-3} under strong source-relay channel. The BER performance of the proposed scheme becomes significantly improved and achieve excellent performance in high SNR regime.

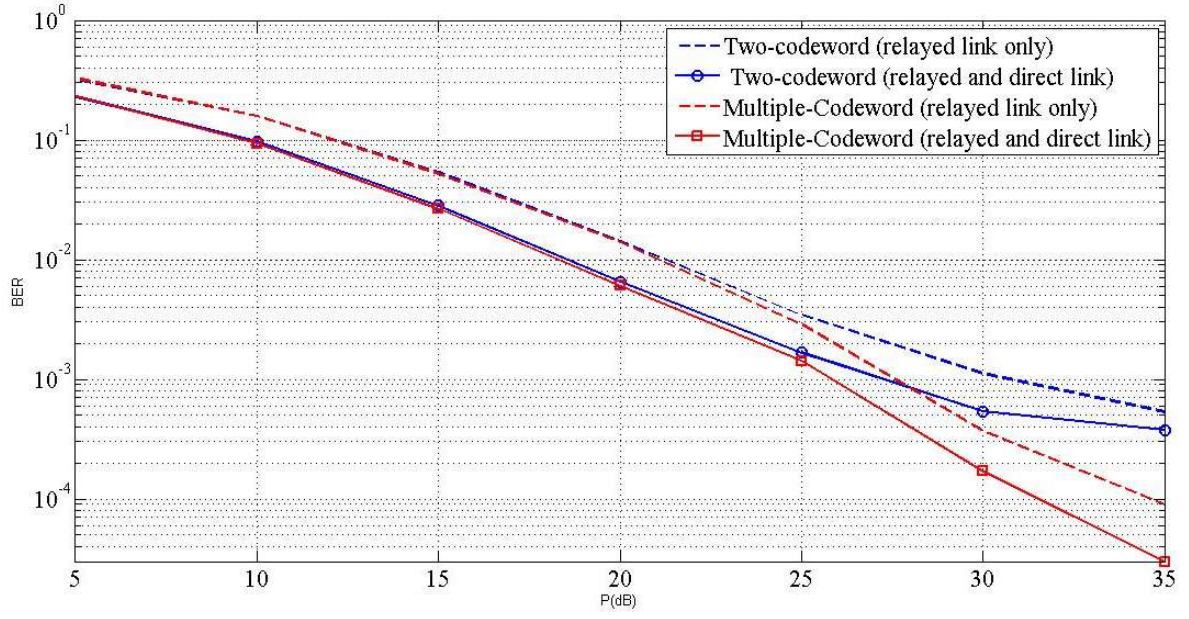


Figure 5.3: Comparison of the BER performance for the conventional DD and proposed MCDDSD scheme when $\sigma_{s,d}^2 = 1$; $\sigma_{s,r}^2 = 10$; $\sigma_{r,d}^2 = 1$

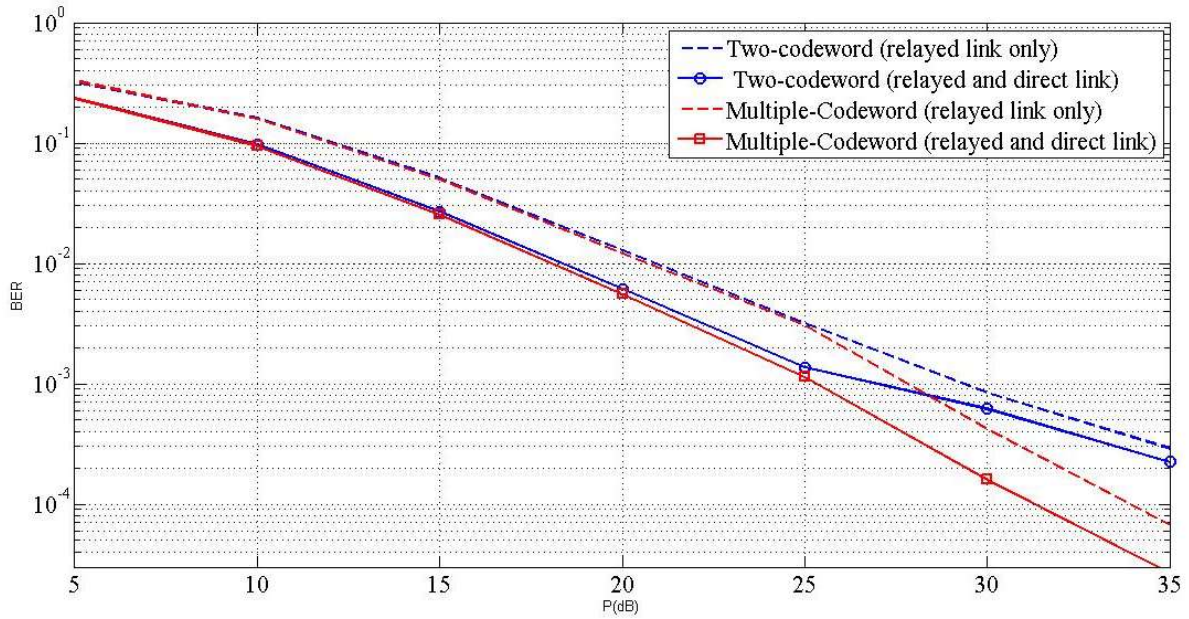


Figure 5.4: Comparison of the BER performance for the conventional DD and proposed MCDDSD scheme when $\sigma_{s,d}^2 = 1$; $\sigma_{s,r}^2 = 1$; $\sigma_{r,d}^2 = 10$

5.3.2 Performance of MCDDSD Under Time-Varying Channels

The proposed scheme is further implemented and simulated for relayed and direct link system in various channels fading scenario. The results are depicted in Figure 5.5. Following Table 5.1, it is considered that the source and destination terminals are moving in slow velocity (i.e. $f_{D_{s,r}}T_s = f_{D_{r,d}}T_s = f_{D_{s,d}}T_s = 0.001$) for case I. The normalized frequency of each channel shows that all the channels are under slow faded. In case II, both source and destination are moving with fairly fast fading channels (i.e. $f_{D_{s,r}}T_s = 0.05$; $f_{D_{r,d}}T_s = 0.05$; $f_{D_{s,d}}T_s = 0.05$). In case III, the source and destination is considered to have high mobility with fast fading channels (i.e. $f_{D_{s,r}}T_s = 0.01$; $f_{D_{r,d}}T_s = 0.01$; $f_{D_{s,d}}T_s = 0.01$).

Table 5.1: Different fading channels

Case	Normalized frequency ($f_{D_{s,d}}T_s, f_{D_{s,r}}T_s, f_{D_{r,d}}T_s$)	Channel condition
I	0.001, 0.001, 0.001	SD, SR, RD slow fading
II	0.05, 0.05, 0.05	SD, SR, RD fairly slow fading
III	0.01, 0.01, 0.01	SD, SR, RD fast fading

It can be observed from Figure 5.5 that for the two-codeword scheme in case I, the BER is monotonically decreasing when SNR increases. As the concentration is shifted to fast fading scenario (i.e. in case II), the curve shows deviation from the case I result at approximately at 20dB and experiences error floor at 35dB. A further performance loss of around 2dB is observed if compared with the slow fading case at BER 10^{-2} . In case III, the two-codeword scheme performance degrades more severe as compared to case II. The degradation starts earlier at around 15 dB and reaches error floor at 30dB.

Due to the poor performance of the two-codeword scheme in both case II and III, multiple-codeword SD based on DD modulation with $N_c = 10$ is employed to compensate the performance degradation in both cases for BPSK and QPSK constellations. The simulation results are plotted in Figure 5.5 and Figure 5.6. It can be seen that the MCDDSD scheme significantly outperform the two-codeword at all SNR in both cases II and III.

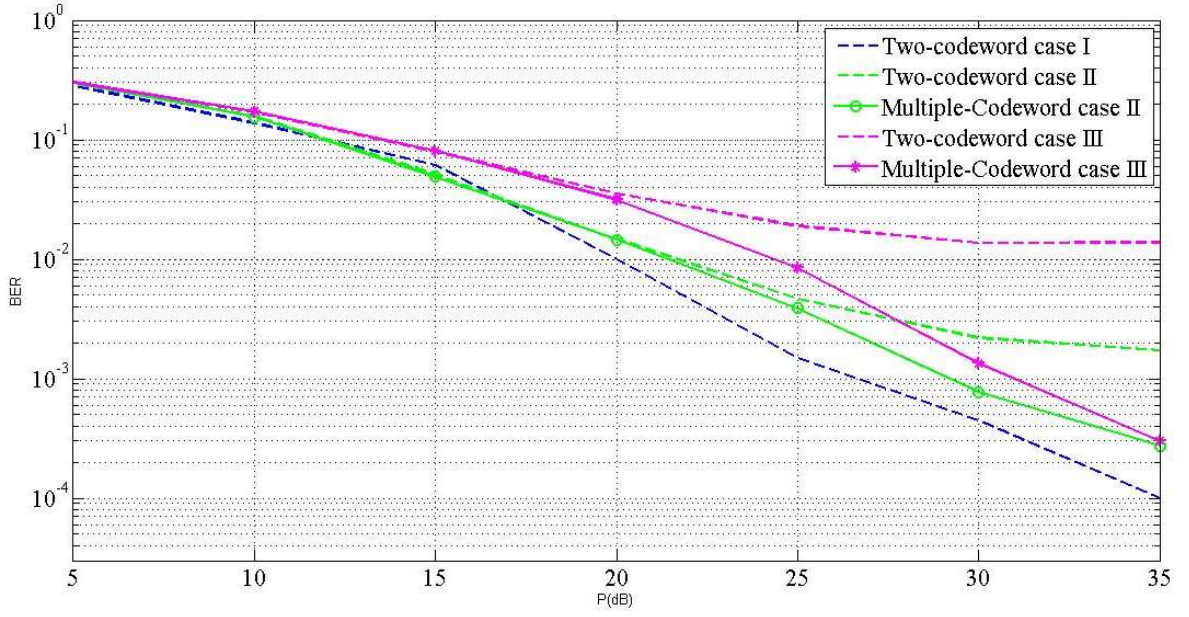


Figure 5.5: Performance comparison of two-codeword and MCDDSD under different fading channels and $(\sigma_{s,d}^2 = 1; \sigma_{s,r}^2 = 1; \sigma_{r,d}^2 = 1)$ for BPSK constellations

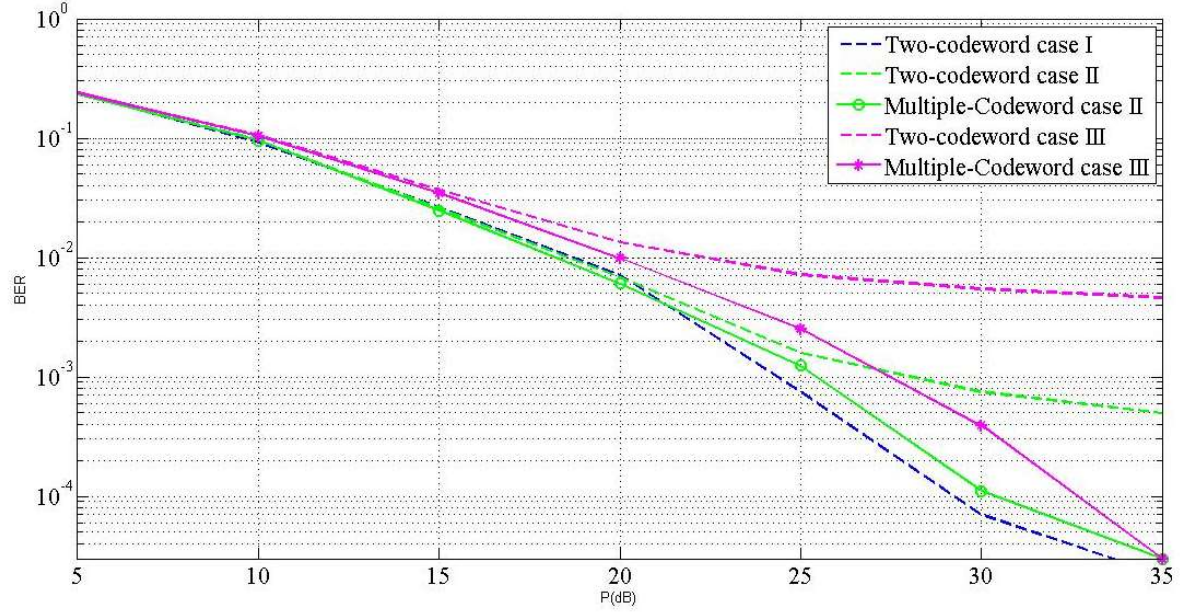


Figure 5.6: Performance comparison of two-codeword and MCDDSD under different fading channels and $\sigma_{s,d}^2 = 1$; $\sigma_{s,r}^2 = 10$; $\sigma_{r,d}^2 = 1$ for QPSK constellations

As demonstrated in Figure 5.5 and 5.6, as expected, the BER curve of case I show the best performance achieved in both system employing BPSK and QPSK constellations since it is simulated in a slow fading environment. Furthermore, in this case the error floor basically does not occur. Hence, the plot is served as a benchmark with the purpose to present the effectiveness of the suggested scheme in the time-varying channels.

From the plots, it can be observed that the proposed method is able to reduce the error floors in fast fading environment and simultaneously increases the performance by narrowing down the performance gap caused by the frequency effects to that of case I.

5.4 Summary

A multiple-codeword based DD transmission was studied over the time-varying channels of the DSTBC strategy. It was revealed that the two-codeword DD detection experience huge degradation, particularly in fast fading channels. Thus, a multiple-codeword

was developed to alleviate the two-codeword performance loss. The results showed that the multiple-codeword was able to gain better performance as compared to its counterpart, two-codeword detection with higher complicated detection process. However, the complexity problem can be solved by the SD algorithm. The theoretical numerical analysis and simulations results were analysed and would become the valuable design criterias to design practical system in corresponding to the users' mobility. The next chapter will conclude the research which highlights the contributions presented in the thesis and the directions of future research.

CHAPTER 6

CONCLUSION AND RECOMMENDATIONS FOR FUTURE WORKS

This chapter presents the discussion of the results and summarizes the main contributions to conclude the thesis. Also, in this chapter, some issues and suggestions that are interesting for the future recommendations are presented.

6.1 Conclusion

Cooperative communication is a promising technique to provide spatial diversity, increase the coverage of the cell and improve the error performance in wireless networks. Most existing research has intensively studied and focused on the coherent detection where full knowledge of the channel coefficient is required. Despite the fact that it is easy to obtain the channel knowledge for a point-to-point wireless network under slow fading channels, it is a challenging task to estimate the channel knowledge and frequency offsets accurately in cooperative relayed network for fast fading environment. In order to alleviate the challenges, this thesis emphasis on efficient detection designs for AF relaying cooperative diversity networks. The primarily focus and aims in this thesis is to design and analysis low-complexity DD detector that require no channel and frequency offsets knowledge in repetition-based and DSTBC transmission networks for one-way relay wireless cooperative system using M -PSK constellations. Eliminating high-complexity channel and frequency offsets estimation is especially vital at the light-weight shirt-pocket-sized mobile handset based relays, due to the reason that it is unrealistic to perform estimation in associate with the mobile-to-mobile channels. Moreover, the extra pilot signals included would also pose a security threat and impose a Doppler-dependent pilot overhead. Therefore DD encoding

scheme and non-coherent detection is proposed to eliminate the estimation of the channel in the presence of frequency offsets.

However, since the DD scheme generally lead to error rate degradation and perform poorly in high mobility applications, numerous detection and decoding schemes are devised and studied to efficiently enhance the performance of DD PSK over fast fading channels. It is demonstrated that the proposed schemes are able to perform effectively and provide positive solutions in the transmission of data for cooperative diversity networks. A brief summary of the contents and major contributions are listed as follows:

1. To study and analyse the effect of the mobility by the users in the performance of the differential AF cooperative relay networks. A dual-hop relayed network incorporating single direct link using the DD M -PSK and symbol based approach is considered. At the destination, a direct combiner with a fixed combining gain is chosen. The conventional DD detector error performances are evaluated against channel variations and verified with the derived theoretical BER expression for the system performance. It is shown that the scheme is able to achieve better performance with diversity order of two. However at the same time, the results show that the two-symbol DD scheme is subjected to a large loss especially in time-varying channels due to the effect of frequency offsets that occur in high mobility applications. Since the differential scheme generally lead to error rate degradation and poor performance (error floor), DD detector utilizing multiple symbols is devised and theoretically analysed to effectively bypass the needs for frequency offset estimation, when neither the relay(s) nor the destination has any knowledge of the channel. It is shown that proposed scheme is able to extend the coverage service of wireless mobile communication at the edge of the cell by employing relay(s) in the

system. The theoretical and simulation performance demonstrate that the proposed multiple-symbol differential detection significantly improve the performance of the system under fast fading channels.

2. In Chapter 3, DDPSK modulated repetition-based transmission using AF relaying protocol for single relay aided one-way half duplex cooperative communication system is presented. The network topology is where a source transmits information signals towards a desired destination via a single relay based on the multiple symbol approach. The computational complexity of the proposed scheme is further simplified by utilizing the SD in which, the search space area of the transmitted symbol are reduced depending on the sphere radius. The simulation result demonstrated that the proposed scheme is able to mitigate the error floor problem inflicted from the conventional DD transmission under fast-fading environment. In addition, the error floor can be mitigated by increasing the spatial diversity of the system. In other words, by increasing the diversity can be achieved by increasing the number of relays in the network to obtain desired performance.

In order to further minimize the BER of the proposed scheme, a numerical power allocation is formed to determine the suitable power to be distributed among the source and relay terminals based on the faded channel. First, the equal power strategy is good, but in general does not perform optimally in the cooperative systems and it is noted that the optimum power allocation is subjected to the quality of the channel link. Secondly, in case that all channel links are available, the optimum power allocation depends only on the channel link between the source and relay as well as that between relay and destination. The optimum power allocation is independent on the source-destination channel. The power allocation results show

that it can achieve better performance as compared to the uniformly distributed power at the source and relays. Specifically, if the link quality between source and relay is better than that between relay and destination, then more power is allocated at the source rather than the relay. Similarly, if the link quality between the relay and destination is much larger than the source-relay link, more power is allocated at the source. In contrast, if the relayed channel (i.e. source-relay-destination) quality is larger than the direct link transmission, then equally distributed power strategy tends to provide optimized system performance. Third, it is observed that the performance gain depend on the different modulation types. Specifically, larger signal constellation size produces less performance gain. Therefore, for high data-rate cooperative communications (with large signal constellation size), AF cooperative protocol can be employed to reduce system complexity while maintains a comparable performance. Theoretical end-to-end BER performance of the proposed multiple symbol SD based on DD modulation is investigated using the union bound of the PEP analysis approach.

Next, a simple SC is appropriately designed for a single and multiple relays to effectively enhanced the BER performance. Compared to the direct combiner, MRC and semi MRC combiner, the SC schemes does not requires any channel and frequency offset information. A theoretical BER expression using DDBPSK is derived for time-varying channels. Both simulation and theoretical analysis results show that the SC scheme achieve diversity order of two. Furthermore its performance is very near to the semi MRC scheme, which requires the second-order statistics of the channel. Since the channel and frequency offsets knowledge are not available in

the fast fading channels, the SC scheme is utilized as an alternative for semi MRC scheme, without degrading the performance.

This thesis also focuses and provides an insight of the implementation for multiple codeword in a DSTBC cooperative networks. By utilizing the multiple codeword based, the performance enhancement of a relayed communication link which includes the direct link without requiring the channel and frequency offsets estimation are presented. The simulation results also show that the conventional two-codeword transmission for DD distributed STC in wireless cooperative diversity networks perform poorly under fast fading environment. Thus, a suboptimal MCDDSD detection is designed based on ML and implemented into the network. Since the complexity of the designed detection scheme increases with the window size N_c , SD is integrated to reduce the search complexity. Simulation results demonstrate that the proposed scheme significantly improve the poor performance of two-codeword scheme under fast fading environments.

As a conclusion, the objectives of this thesis has been achieved. this thesis has studied and analysed the cooperative diversity schemes in a time-varying channel. The study showed that conventional DD transmission degrades substantially under fast fading channel, which reduces the network performances. The MSDDSD and differential SC have been proven to be effective to improve the AF MSDD transmission for cooperative wireless communication with different mobility. The results are simulate in simplified form which enable the analysis of MSDD detection without the need of exhaustive search at the destination using SD. Furthermore, closed-form expressions of BER and outage probability analysis have been derived to forecast the behaviour of the cooperative system. The thesis revealed that the Matlab simulations are verified with numerical analysis results. The thesis also compared

the proposed schemes with the existing conventional schemes which reveals that the proposed schemes outperform the conventional schemes (i.e. DD and MRC) in repetition and DSTC strategy. It is demonstrated that the proposed MSDDSD together with SC can overcome the typical irreducible error floor of two-symbols differential detectors particularly under fast fading environments in terms of BER and outage probability. In addition, a numerical power allocation based on the derived numerical expression has been performed to further enhance the system performance.

This thesis mainly focused on developing and analysing non-coherent detection and combiner schemes for cooperative network. The schemes are effective wherein communication overhead required for channel and frequency estimation is not wasted. This is especially very useful for the emerging next-generation broadband technologies, such as LTE wherein these technologies are estimated to support high data rates performance with high mobility. As inaccurate channel and frequency estimation is experienced especially in high mobility environment, it is essential to adopt non-coherent detection and combiner schemes for the future technologies. Furthermore, as the network increases, whereby more relays are implemented, the developed schemes are able to transmit the information effectively over wireless relay networks.

6.2 Future Recommendations

This thesis aim is to examine some important issues for cooperative communication employing DD scheme. In order to achieve this aim, other concerns arised and it is worthy to be investigated in details. The issues are listed as follows:

- MSDDSD is proposed and analysed for a relayed and direct link. This detection is attractive due to its ability to mitigate the error floor suffered in conventional DD

transmission, especially over fast Rayleigh fading environment. For a multi-hop differential cooperative network, the propagated signals experienced higher channel fading. Although it seems challenging, it will be motivating to simulate and analyse the MSDDSD performance for multiple hop differential cooperative network to reinforce the reliability of cooperating link, particularly when fast fading channel is considered.

- In this thesis, the relay(s) is assumed to employ SISO antenna. Investigate the MIMO relay networks for distributed STC where the multiple antennas at the relay terminal introduces additional dimension for designing distributed coding. Furthermore, there are also some practical issues that should be considered in the implementation of MIMO relay into the wireless cooperative networks, such as interference cancellation design at relays and destination, optimum vector relay protocols design, coding structures and code design criteria.
- In this thesis, it is assumed that frequency offset exists in the relay networks. However, in real practical environment, several interference such as Co-Channel Interference (CCI) and ISI would arise because of the frequency re-used and synchronised transmission of multiple users over the similar frequency-band as well as imperfect synchronization among multiple relays. Thus, it is interesting to take into the consideration of both CCI and ISI for time-varying channels and develop detection schemes to overcome the effects of CCI and ISI.

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