UNIVERSITI SAINS MALAYSIA

Laporan Akhir Projek Penyelidikan Jangka Pendek

Network Coding Techniques Using Novel Fast 2-Bit Error Correction Codes For Wireless Communications

by

Assoc. Prof. Dr. Mohd Fadzli Mohd Salleh

Assoc. Prof. Dr. Bakhtiar Afendi Rosdi

2015

RU GRANT
FINAL REPORT CHECKLIST

Please use this checklist to self-assess your report before submitting to RCMO.
Checklist should accompany the report.

NO.	ITEM	PLEASE CHECK (✓)		
		PI	JKPTJ	RCMO
1	Completed Final Report Form	✓		
2	Project Financial Account Statement (e-Statement)	✓		
3	Asset/Inventory Return Form *(Borang Penyerahan Aset/Inventori)*	✓		
4	A copy of the publications/proceedings listed in Section D(ii) (Research Output)	✓		
5	Comprehensive Technical Report	✓		
6	Other supporting documents, if any			
7	Project Leader's Signature	✓		
8	Endorsement of PTJ's Evaluation Committee	✓	✓	
9	Endorsement of Dean/ Director of PTJ's	✓	✓	

Project Code :
(for RCMO use only)

UNIVERSITI SAINS MALAYSIA APEX	**RU GRANT FINAL REPORT FORM**

Please email a softcopy of this report to rcmo@usm.my

A	**PROJECT DETAILS**
i	**Title of Research:** Network Coding Techniques Using Novel Fast 2-Bit Error Correction Codes For Wireless Communications
ii	**Account Number:** 1001/PELECT/814178
iii	**Name of Research Leader:** Mohd Fadzli Bin Mohd Salleh
iv	**Name of Co-Researcher:** 1. PM Dr. Bakhtiar Afendi Bin Rosdi
v	**Duration of this research:** **a) Start Date** : 15 December 2012 **b) Completion Date** : 14 December 2015 **c) Duration** : 36 months **d) Revised Date** (*if any*) : ....
B	**ABSTRACT OF RESEARCH**
	(An abstract of between 100 and 200 words must be prepared in Bahasa Malaysia and in English. This abstract will be included in the Report of the Research and Innovation Section at a later date as a means of presenting the project findings of the researcher/s to the University and the community at large) *VERSI BAHASA MALAYSIA* Pada masa ini, teknik-teknik pengesanan kesilapan dan pembetulan pengekodan adalah sangat penting dalam sistem penghantaran data terutamanya dalam penghantaran kadar data yang tinggi rangkaian aplikasi tanpa wayar. Data yang dihantar melalui saluran komunikasi tanpa wayar mungkin rosak atau terherot kerana hingar atau gangguan. Beberapa kod mudah boleh mengesan kesilapan tetapi tidak boleh membetulkannya,

seperti kod kitaran lebihan semak (CRC) dan bingkai urutan semak (FCS) yang biasa digunakan dalam lapisan pautan data (contohnya, bingkai pengepala) dan lapisan rangkaian (contohnya, paket pengepala) dalam rangkaian komputer. kod lain boleh mengesan dan membetulkan satu kesilapan seperti kod Hamming. Projek ini mencadangkan pengesanan kod blok yang baharu dalam pengesanan dan pembetulan ralat bit satu dan dua yang boleh menyelesaikan kesilapan berturut-turut dan tidak berturut-turut. Di samping itu, ia boleh mengesan kesilapan lebih daripada dua bit (yang bermaksud ralat pecah). Manfaat kaedah yang dicadangkan ini akan meningkatkan prestasi penghantaran data dalam rangkaian seperti meningkatkan daya pemprosesan, mengurangkan kelewatan hujung-ke-akhir dan kesilapan kadar bit.

ENGLISH VERSION

At present, errors detection and coding correction techniques are very important part of data transmission especially in high data rate wireless networks applications. Data being transmitted over a wireless communications channel may be damaged or distorted due to noise or interference. Some simple codes can detect error but cannot correct it, such as cyclic redundancy check (CRC) and frame check sequence (FCS) which commonly used in data link layer (e.g., frame header) and network layer (e.g., packet header) in computer networks. Other codes can detect and correct one bit errors such as Hamming code. This project proposes a novel block code detection and correction for single and double bit error that supports consecutive and non-consecutive bit errors. In addition, it can detect more than double bit errors (which means burst error). The benefits of this proposed method will improve network transmission performance such as increasing throughput, reducing end-to-end delay and bit error rate.

C BUDGET & EXPENDITURE

i

Total Approved Budget : RM 248,000.00

Yearly Budget Distributed

Year 1 : RM 141,000.00
Year 2 : RM 63,000.00
Year 3 : RM 00,000.00

Total Expenditure : RM 154,951.02

Balance : RM 49,048.98

Percentage of Amount Spent (%) : 76

 # Please attach final account statement (eStatement) to indicate the project expenditure

ii	Equipment Purchased Under Vot 35000

No.	Name of Equipment	Amount (RM)	Location	Status
1	WARP MIMO KIT (V3) – VIRTEX IV (2 UNITS)	70,840.00	DSP Lab	available

 # Please attach the Asset/Inventory Return Form (Borang Penyerahan Aset/Inventori) – Appendix 1

D RESEARCH ACHIEVEMENTS

i Project Objectives *(as stated/approved in the project proposal)*

No.	Project Objectives	Achievement
1	To investigate the diversity improvement of network coding techniques in wireless communication networks.	Achieved
2	To develop novel fast 2-bit error corection codes for network coding.	Achieved
3	To develop new data link and network layers network coding techniques using novel fast 2-bit error corection codes.	Achieved
4	To implement the hardware prototype of the novel physical layer network coing techniques in wireless communication networks using Software Define Radio (SDR) environment.	Achieved

ii Research Output

a) Publications in ISI Web of Science/Scopus

No.	**Publication** (authors,title,journal,year,volume,pages,etc.)	**Status of Publication** (published/accepted/ under review)
1	S. A. Alabady, and M. F. M. Salleh, "Analysis of the Effect of Binary OR Operation Property in Linear Network Coding Transmission Performance," *IETE Journal of Research*, vol. 62, no. 2, pp. 164-178, 2016.	Published
2	J. Zakaria and M. F. M. Salleh, "A New PAPR Reduction Scheme: Wavelet Packet-based PTS with Embedded Side Information Data Scheme," *IET Communications*	Conditional acceptance (Revision)

3	S. A. Alabady, and M. F. M. Salleh, "Wireless network coding for multiradio multichannel mesh networks," *Malaysian Journal of Computer Science.*	Under review
4	S. A. Alabady, and M. F. M. Salleh, "Low Complexity Parity Check (LCPC) Codes," *AEU Intemational Joumal of Electronics and Communications-ELSEVIER*	Under review
5	S. A. Alabady, and M. F. M. Salleh, "Saturation Throughput and Delay Performance Evaluation of IEEE 802.11g/n for a Wireless Lossy Channel," *KSII Transactions on Internet and Information Systems*	Under review
6	M. A. Sidig and M. F. M. Salleh, "Buffer-aided Relay Selection for Cooperative Diversity Systems without Delay Constraints," *IEEE Transactions on Vehicular Technology.*	Under review

b) Publications in Other Journals

No.	Publication (authors,title,journal,year,volume,pages,etc.)	Status of Publication (published/accepted/ under review)

c) Other Publications
(book,chapters in book,monograph,magazine,etc.)

No.	Publication (authors,title,journal,year,volume,pages,etc.)	Status of Publication (published/accepted/ under review)

d) Conference Proceeding

No.	Conference (conference name,date,place)	Title of Abstract/Article	Level (International/National)
1	IEEE Region 10 Conference (TENCON 2015), 1-4 Nov 2016, Macau China	Three Description Lattice Vector Quantization for Efficient Data Transmission	International

 # Please attach a full copy of the publication/proceeding listed above

iii	**Other Research Ouput/Impact From This Project** *(patent, products, awards, copyright, external grant, networking, etc.)*

E HUMAN CAPITAL DEVELOPMENT

a) Graduated Human Capital

Student	Nationality (No.)		Name
	National	International	
PhD			1. Salah A. Alabady
MSc			1. Jamaludin Bin Zakaria
Undergraduate			1. Ooi Tzy Way 2. Chew Wei Loon 3. Goh Shi Zhao

b) On-going Human Capital

Student	Nationality (No.)		Name
	National	International	
PhD			1. 2.
MSc			1. 2.
Undergraduate			1. 2.

c) Others Human Capital

Student	Nationality (No.)		Name
	National	International	
Post Doctoral Fellow			1. 2.
Research Officer			1. Jamaluddin Bin Zakaria (after MSc) 2.
Research Assistant			1. Teo Hui Ting 2.
Others (.....................)			1. 2.

F COMPREHENSIVE TECHNICAL REPORT

Applicants are required to prepare a comprehensive technical report explaining the project. The following format should be used (this report must be attached separately):

- Introduction
- Objectives
- Methods
- Results
- Discussion
- Conclusion and Suggestion
- Acknowledgements
- References

G	PROBLEMS/CONSTRAINTS/CHALLENGES IF ANY
	(Please provide issues arising from the project and how they were resolved) Support software package (such as Xilinx, for FPGA development) is required for the SDR hardware. However the license is very expensive (not affordable by the amount of grant).
H	**RECOMMENDATION**
	(Please provide recommendations that can be used to improve the delivery of information, grant management, guidelines and policy, etc.) Please make managing the grant be more flexible to the principle researcher. There are too many layers of approval which produce the unneccessary delay.

Project Leader's Signature:

Dr. Mohd Fadzli Bin Mohd Salleh
Associate Professor
School of Electrical and Electronic Engineering
Universiti Sains Malaysia, Seri Ampangan,
14300 Nibong Tebal, Pulau Pinang.

Name : MOHD FADZLI BIN MOHD SALLEH

Date : 20/7/2016

I	COMMENTS, IF ANY/ENDORSEMENT BY PTJ'S RESEARCH COMMITTEE

Acceptable research output but financial achievement are not as budgeted.

Signature and Stamp of Chairperson of PTJ's Evaluation Committee

Name : PROFESSOR DR. MOHD FADZIL BIN AIN
Deputy Dean
(Research, Postgraduate & Networking)
School of Electrical And Electronic Engineering
Engineering Campus
Universiti Sains Malaysia

Date : 26/7/16

Signature and Stamp of Dean/ Director of PTJ

Name : PROFESOR DR. MOHD RIZAL ARSHAD
Dean
School of Electrical & Electronic Engineering
Engineering Campus
Universiti Sains Malaysia

Date : 27-7-16

Appendix 1

UNIVERSITI SAINS MALAYSIA

BORANG PENYERAHAN ASET / INVENTORI

==

A. BUTIR PENYELIDIK

1. NAMA PENYELIDIK : Mohd Fadzli Bin Mohd Salleh
2. NO STAF : AE50304
3. PTJ : ELEKTRIK DAN ELEKTRONIK
4. KOD PROJEK : 814178
5. TARIKH TAMAT PENYELIDIKAN : 14/12/2015

B. MAKLUMAT ASET / INVENTORI

BIL	KETERANGAN ASET	NO HARTA	NO. SIRI	HARGA (RM)
1	RU WARP V3 FPGA KIT	AK00007477	W3-A-00442	35,420.00
2	RU WARP V3 FPGA KIT	AK00007478	W3-A-00308	35,420.00

C. PERAKUAN PENYERAHAN

Saya dengan ini menyerahkan aset/ inventori seperti butiran B di atas kepada pihak Universiti:

....

(Mohd Fadzli Bin Mohd Salleh) Tarikh: 19/07/2016

D. PERAKUAN PENERIMAAN

Saya telah memeriksa dan menyemak setiap alatan dan didapati :

- [x] Lengkap
- [ ] Rosak
- [ ] Hilang : Nyatakan..
- [x] Lain-lain : Nyatakan masih digunakan oleh pelajar sisuazah lanjut.

Diperakukan Oleh :

....
Tandatangan
Pegawai Aset PTJ

Nama : KHAIRUL ANUAR BIN AB. RAZAK
Penolong Jurutera
Pusat Pengajian Kejuruteraan Elektrik & Elektronik
Universiti Sains Malaysia
Kampus Kejuruteraan

Tarikh : 19/07/2016

***Nota :** Sesalinan borang yang telah lengkap perlulah dikemukakan kepada Unit Pengurusan Harta, Jabatan Bendahari dan Pejabat RCMO untuk tujuan rekod.

PENYATA PERUNTUKAN DAN PERBELANJAAN SEHINGGA 19 JULAI 2016

Kwg	Akaun	PTJ	Projek	Peruntukan	Bel.T.kumpul	Baki Peruntukan Tahun Lalu	Peruntukan Tahun Semasa	Jumlah Peruntukan Tahun Semasa	Tanggungan Semasa	Bayaran Tahun Semasa	Jumlah Belanja Tahun Semasa	Baki Projek
				RM	RM	RM	RM	RM	RM	RM	RM	RM
1001	111	PELECT	814178	84,000.00	43,237.38	40,762.62	-	40,762.62	-	-	-	40,762.62
1001	221	PELECT	814178	5,000.00	6,323.25	(1,323.25)	-	(1,323.25)	-	-	-	(1,323.25)
1001	227	PELECT	814178	12,000.00	5,834.90	6,165.10	-	6,165.10	-	-	-	6,165.10
1001	228	PELECT	814178	3,000.00	-	3,000.00	-	3,000.00	-	-	-	3,000.00
1001	229	PELECT	814178	6,000.00	34,197.49	(28,197.49)	-	(28,197.49)	-	-	-	(28,197.49)
1001	335	PELECT	814178	94,000.00	65,358.00	28,642.00	-	28,642.00	-	-	-	28,642.00
				204,000.00	154,951.02	49,048.98	-	49,048.98	-	-	-	49,048.98

Unit Perngurusan Perakaunan

Jabatan Bendahari,Kampus Kejuruteraan

PH2PA2P020/07/2016

Title of Research :

Network Coding Techniques Using Novel Fast 2-Bit Error Correction Codes For Wireless Communications

Research Leader :

Associate Prof. Dr. Mohd Fadzli Bin Mohd Salleh

Co-Researcher:

Associate Prof. Dr. Bakhtiar Afendi Bin Rosdi

Grant Number: 1001/PELECT/814178

1. INTRODUCTION

1.1 Motivation

During the past decade, many schemes have been proposed to increase ther eliability of the wireless network systems, in order to fulfill the quality of the data in a high data rate wireless network applications, such as telephone conversations, video conference and television cameras. In wireless networks and wireless digital communication systems, channel coding have received considerable attention, since some data bits might be exposed to attenuation or distortion due to interferences, channel noise and multi-path fading. In addition, both random and burst errors occur during transmission in noisy communication channel. These types of errors increase the bit-error-rate (BER) which results in bad quality transmission. Channel coding is considered as one of the major boosts, which enhances the performance of wireless networks at higher data rates. In wireless network applications as well as realtime application systems, low complexity and shorter codeword length in channel coding scheme are preferred.

In a wireless network,the packet is dropped if there transmitted information is still erroneous and this is a waste of bandwidth and time, which leads to bad performance fornetworks (Mahmoud and Shen, 2011; Alayed and Rikli, 2013; Esmaeilzadeh et. al., 2014). In addition, packets can be lost due to errors, congestionina queuedueto high traffic, collision, hidden nodes, and the link failures. The problem of packets dropped becomes more critical in real-time applications such as video conference and remote control in a wireless sensor network (WSN) (Kobbane et al., 2013). In addition to the channel coding, a wireless network coding(WNC) has been considered as a promising solution with its capability in improving the quality of service (QoS) performance of wireless networks by retrieving the packet lost. In a traditional store-and forward network, packets are forwarded hop-by-hop along the intermediate routers from a source to a receiver. An intermediate node forwards the packets as it receives through a predefined path. WNC is a recent field of information theory that allows the intermediate nodes,instead of simply forwarding data, to generate new packets by combining packets received on their incoming ports before sending the combined data on its output links to increase the capacity and throughput of each link (path), and increase the packet delivery ratio (PDR).

The aims of channel coding and network coding techniques are retrieving the original information. Channel coding is used for point-to-point communication over a single channel, and it uses the error correction codes to improve the error performance of the wireless link. It is implemented at the physical layer to recover erroneous bits through redundant parity check bits added inside a packet. The error retrieval capability for channel coding depends on the specific coding and the amount of redundant bits. An erasure correction coding can be used to handle lost packets on the end-to end connection level that uses redundant packets to recover the original information at the network layer for end-to-end communication. Traditional network coding allows the intermediate nodes to generate a redundant network-coded packet (Fragouli et al.,2006; Matsudaetal.,2011). Joint channel-network coding provides reliable communication and achieves a better performance for a network communication system instead of the case when channel and network codes are designed separately (Tranetal.,2008; Hausl and Dupraz, 2006; Qiang et al., 2009). Recently, many authors proposed design for channel network codes to exploit the benefits of channel and network codes in recovering the original information.

This project presents a new channel coding scheme, namely low complexity parity check (LCPC) code. LCPC code is a linear block code for error detection and correction in wireless communication networks. The LCPC codes have the capability to detect and correct consecutive and non-consecutive bit errors. Also, this project presents new wireless network correcting codes (WNC) with a router detection packet loss (WNC-RDPL) algorithm for recovering lost packets. WNC increases the band width efficiency of reliable broadcast in a wireless network. WNC code employs network coding to reduce the number of retransmission as a result of packet losses due to the link failures. In addition, this project presents a novel binary joint network-channel coding (B-JNCC) scheme for reliable multi-radio multi-channel multi-hop wireless networks. B-JNCC combines LCPC and WNC-RDPL codes in order to reduce the BER and increases the PDR over a wireless channel.

1.2 ProblemStatement

Transmissions in the wireless networks are jeopardized by the injection of errors or erasure of symbols data. Errors are caused by the channel noise or by the multi-path fading channel. Recently, Low Density Parity-Check (LDPC) codes (Gallager, 1963) have been extensively developed and regarded as the best channel coding schemes. The error correction capability of the LDPC codes depend on the codeword length and the characteristic of the parity check matrix H (Gallager,1963;Carrasco and Johnston,2009). The decoder gives a better performance with a larger codeword (i.e. large size of G matrix) and with good parity-check matrix H (Carrasco and Johnston, 2009). In practice, to achieve a better BER performance with LDPC codes close to the channel capacity, the length of the LDPC codeword used should be in the order of thousands of bits,(CarrascoandJohnston,2009). The LDPC decoding is effective only when the parity-check matrix has a relatively large column weight (Bhargava and Bose, 2013). The matrix multiplication for this big codeword size demands huge memory, computational requirements and more complex decoding (Kou et al., 2001; Voicila et al.,2007; Carrasco and Johnston, 2009). Consequently, the existing decoding algorithms are either too costly to implement (Wang et al., 2013).

Furthermore, LDPC codes require iteration in the detection and correction error processes around 10 to 50 times of iteration (SalbiyonoandAdiono,2010). For example, the average number of iterations for iterative decoding of the LDPC (1008, 504) code with belief propagation (BP) and uniformly most powerful (UMP) BP-based decoding algorithms, is 50 and 200 (Fossorier et al., 1999). Besides, the decoder fails to correct errors if the number of errors occurred is greater than the error correction capability of the decoder regardless of the number of iterations (Carrasco and Johnston, 2009). For practical applications, these codes are inappropriate to be used since they involve high encoding-decoding complexity (Yahya, 2010). Therefore, the need of efficient channel codes with lower encoding and decoding complexity, and lower memory size requirement, which do not require any iteration in the decoding process, is quite obvious. As the solution to the above problems, this project proposes Low Complexity Parity Check (LCPC) codes that can offer several different code rates.

LCPC codes provide better BER performance. LCPC codes have low complexity (low O(n)) and small memory size requirement. Unlike LDPC, the proposed LCPC codes do not need any iteration during the decoding. The short codeword length of LCPC codes reduce the possibility of error and decrease decoding delay compared to the big codeword length in LDPC codes. LCPC codes are particularly appropriate for real time multimedia applications through the use of a small codeword length to decrease decoding delay, and are able to further increase network throughput. On the other hand, the demand for real time applications and multimedia applications has increased due to the rapid increase in the number of wireless network users.

In addition, providing high-speed and reliable services with the ability to access video over the Internet and share large files in wireless networks is the fundamental challenge due to the interference and unreliable nature of wireless link (i.e., variable link qualities), which causes packet losses and link failures. Moreover, some applications cannot use automatic repeat-request (ARQ) when the packet is lost, because the original packets are no longer available in the source. For example, most of the television cameras immediately for get the information as soon as they are sent, and cannot resend the original packets.

Recently, network coding (NC) has been proposed as an effective technique to in crease the network bandwidth-efficiency (Fragouli et al., 2006) and throughput (Kattiet al., 2008). NC gives intermediate nodes the ability of randomly encoding different packets received previously into one output packet. Linear Network coding (LNC) is one research area that able to enhance the reliability and efficiency of network communications systems. In LNC, packets transferred through a network are viewed as symbols in finite field GF (q) on which their arithmetic operations are defined, where q denotes the size of finite field. Each coding node has two or three input links and one or two output links. Coefficients are assigned to inputlink so f coding nodes. Each input link has one coefficient. These coefficients are randomly generated and selected from the Galois Field GF(q). These coefficients are grouped together which refer to as the coding vector. This coding vector was used to encode the native packets at the source node and to retrieve native packets at the receiver nodes.

In order to retrieve native packets, receiver nodes need to know the coding vector of the coded packets they received. Chou et al. (2003) proposed a method of embedding the coding vector in the header of the coded packet in order to deliver coding vectors to the receiver nodes. The main problem in LNC is, if some error occurs in the coding vector (coefficients) through the transmission or it is lost due to the unreliable nature of wireless link, interference or link failure. In this case, the receiver cannot retrieve the native packets as it does not have the correct coefficients. In LNC system, the transmission was assumed to be in ideal channel with no error, no packet loss and no link failure.

Motivated by the above problems, a new Wireless Network Coding (WNC) architecture and its router detection packet loss (RDPL) algorithm are proposed in this project with assumption of having packet loss and link failure. In WNC-RDPL, there is not any need for coding vector (coefficients). The proposed scheme focuses on exploiting NC in multi-hop multi-radio multi-channel wireless networks. The new Wireless Network Coding (WNC) architecture and its router detection packet loss (RDPL) algorithm has the capability to retrieve the lost packets in wireless network. Overall, the proposed architecture is called WNC-RDPL. WNC-RDPL scheme employs network coding to reduce the number of retransmission as a result of packet losses due to the link failures.

A data transmission over wireless networks and communications frequently suffers from delay and packet loss due to the several reasons such as unreliable nature of wireless channel, interference, link failure, fading and shadowing wireless link. The high and variable BER in the wireless channel considers the main problem to implement a reliable wireless network. For real time applications, packet loss probability is the most important performance measures. In order to address the problem of packet losses and to improve the performance of wireless network, a binary joint network coding with channel coding (B-JNCC) scheme is proposed in this research. B-JNCC combines low complexity parity check (LCPC) channel coding and wireless
network coding without errror detection packets loss (WNC-RDPL) scheme to reduce the chance of error propagation and improve the error correcting performance for multi radio multi-channel wireless networks.

Channel coding schemes are implemented at the physical layer to recover erroneous bits through redundant parity check bits added inside a packet. Whereas, wireless network correction code used to handle lost packets on the end-to-end connection level that uses redundant packets to recover the original information at the network layer for end-to-end communication. Joint channel-network coding provides reliable communication and achieves a better performance in a network communication instead of the case when channel and network codes are designed separately (Tranetal.,2008; Qiang et al., 2009; Guo et al., 2012; Yu and Zhaoyang, 2013; Hernaez et al., 2013).

1.3 Project Objectives

The main aim of this research is to introduce a novel channel coding scheme as a useful and usable for wireless communications networks. In addition to propose a new wireless network coding (WNC) scheme for wireless networks. The new schemes have many advantages such as reduces the complexity, decrease the BER and increase the PDR in the wireless networks. The research objectives are given as follows:

1. To investigate the diversity improvement of network coding techniques in wireless communication networks.
2. To develop novel fast 2-bit error corection codes for network coding.
3. To develop new data link and network layers network coding techniques using novel fast 2-bit error corection codes.
4. To implement the hardware prototype of the novel physical layer network coing techniques in wireless communication networks using Software Define Radio (SDR) environment.

A step-by-step approach is taken in this research to achieve the above objectives. The research methodology is explained in the next section

2. METHODS

2.1 Overview of Methodology

Figure 2.1 depicts an overview of this research. The scope of this research is focused on the channel coding and network coding,which are an important part of the wireless communications networks for decrease the BER and increase the PDR as explained in

the section 1.1. In this research, a novel linear block channel coding for wireless communication networks is proposed. The design includes different values of code rates

(i.e.,0.444,0.375and0.428)that implemented using the MATLAB programming. The proposed codes are very effective in correcting random bit errors and random symbol errors over wireless fading channels and Additive White Gaussian Noise (AWGN) channels at Binary Phase-Shift Keying (BPSK) and 16-Quadrature Amplitude Modulation (QAM) modulations. The advantages of the proposed codes are lowc omplexity, low memory size requirement and on iteration in the decoding process.

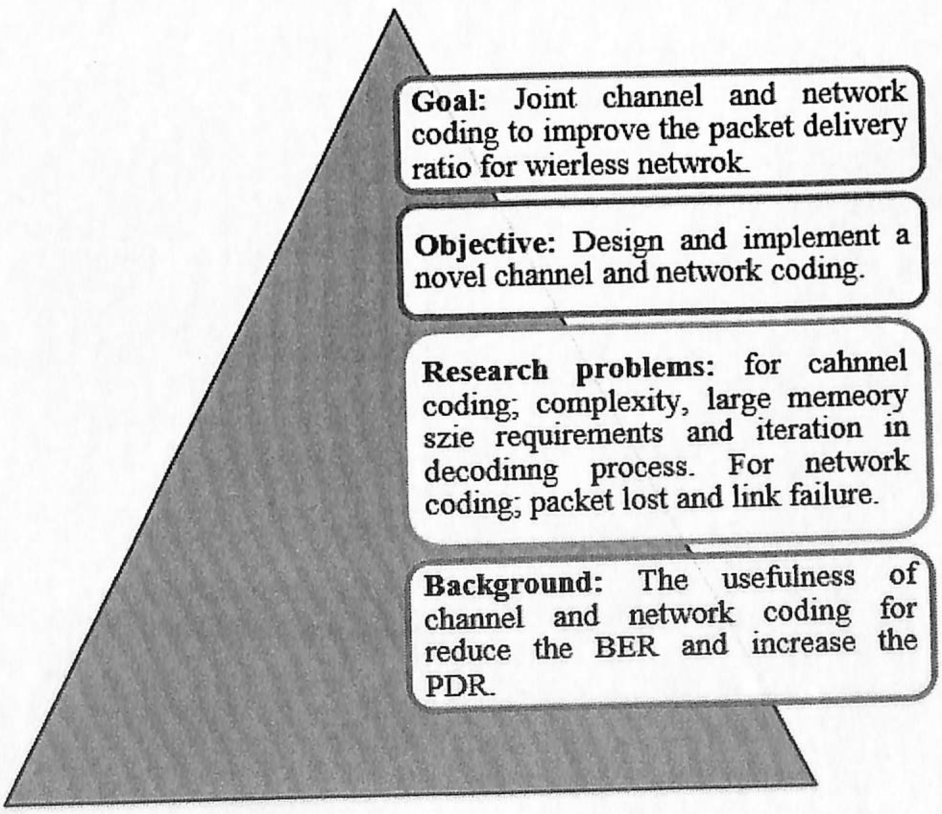


Figure 2.1: Research flow

In addition, in this research a novel wireless network coding (WNC) is proposed to recovering lost packets due to the link failure in wireless networks. The proposed code is implemented and evaluated using the MATLAB programming for four different topologies scenarios. The WNC is evaluated for different possibilities of links failure at four different network topologies. On the other hand, the analytical analysis on the effect of using the binary filed (OR operation) property, to generate the coding matrix elements, instead of using the Galois Field (GF) operation property to generate the coding matrix elements in Linear Network Coding (LNC) is presented. The LNC with binary OR operation (LNC_OR) is simulated using MATLAB programming under various multi-hop wireless network conditions of coefficients and coding router for three different network topologies. Performances of the LNC with the different probability of the coefficients of coding matrix base on packet delivery ratio (PDR) are analyzed.

2.2 Low Complexity Parity Check Codes (LCPC)

In information theory, error detection and correction are techniques that facilitate reliable delivery of digital data over unreliable communication channels by decrease the BER and increase the PDR. Many communication channels are subjected to channel noise, and thus errors may be introduced during transmission from source to receiver nodes. This section presents the first contribution of this research,i.e.,the novel linear block codes for error detection and correction called Low Complexity Parity Check (LCPC) codes. LCPC codes are efficient that provide reliable transmission in wireless channel. The LCPC (9, 4), LCPC (8, 3) and LCPC (7, 3) codes with different code rates (i.e., 0.444, 0.375 and 0.428) respectively are proposed. The codes are short in length, which makes the proposedcodes attractive forlow-latencyandreal time applications. Table 2.1 shows the specification of the proposed LCPC codes, where k is the number of information bits, n number of codeword bits, dmin minimum distance (n-k), and R is the code rate (k/n).

Table 2.1: LCPC code specification

Type of code	Parameters			
	n	k	d_{min}	R
LCPC (9, 4)	9	4	5	0.444
LCPC (8, 3)	8	3	5	0.375
LCPC (7, 3)	7	3	4	0.428

2.2.1 The Proposed LCPC (9,4) code

Low Complexity Parity Check (LCPC) code is a one type of block linear code that used for error detection and correction. The main difference between our proposed LCPC code and the other codes such as RS, and LDPC is that it is a simple encoding and decoding method with low complexity. Besides, it does not require any iteration process in the decoding and can work with very low memory requirement. These advantages make the LCPC code a good choice for real time application. In this research, design a novel linear block channel coding scheme with four algorithms, to detect and correct single, double and more than double bit errors for wireless communication networks is proposed. The new channel coding The design includes different values of code rates (i.e., 0.444, 0.375 and 0.428). The new channel coding

schemes with different code rates are implemented using MATLAB program.

To validate the performance of the proposed LCPC code, it is investigated at different data transmission values using the Binary Phase-Shift Keying (BPSK) and 4-Quadrature Amplitude Modulation (QAM) and 16-QAM modulations over the Additive White Gaussian Noise (AWGN) and wireless fading channels. The proposed codes are very effective in correcting random bit errors and random symbol errors over wireless fading channels and AWGN channels at BPSK, 4-QAM and 16-QAM modulations. In this work, four types of LCPC codes as, LCPC-a (9, 4), LCPC-b (9, 4), LCPC (8, 3) and LCPC (7, 3) are proposed. LCPC-a (9, 4) and LCPC-b (9, 4) codes have the same code rate (i.e., 0.444). They have the same BER performance and the same capability for error correction, but the G and

H matrices are different (number of one in Gand H matrices for LCPC-b (9,4) is less than in LCPC-a(9,4)).

This section presents the G and H matrices with error patterns and syndrome vector Tables for LCPC-a (9, 4) codes. The G and H matrices with error patterns and syndrome vector Tables for LCPC-b (9, 4), LCPC (8, 3) and LCPC (7, 3) codes are presented in Appendix B. The encoding and decoding processes are the same for all LCPC codes.

Figure 2.1 shows the components of the LCPC codes. The LCPC code is defined as
block code (n,k) where the codeword length n=9 and the data (sample) length k=4 for the LCPC-a (9,4) and LCPC-b (9,4). The LCPC codes has the capability to detect and correct consecutive and non-consecutive bit errors. In addition to the main units (i.e., encoding, error detection, error correction and recovery), other unites such as XOR, lookup tables for error pattern (EP) and syndrome vector (SY) are also considered as important parts of the LCPC codes. Matrices algebra and polynomial functions describe the implementation of the LCPC codes. Usually, the decoder contains two parts: one for error detection (for syndrome vector calculation), and the other for error correction (assign error pattern). The general idea for achieving error detection and correction is by adding some redundancy bits to a source data (Carrasco and Johnston, 2009). After encoding the original data with the redundant bits at the source, a new data is formed which is termed as a codeword. When a destination receives the codeword, the decoder in the receiver can use the redundancy bits or parity check matrix (as proposed in this work) to check the reliability of the delivered data, and to recover the
original data from erroneous data.

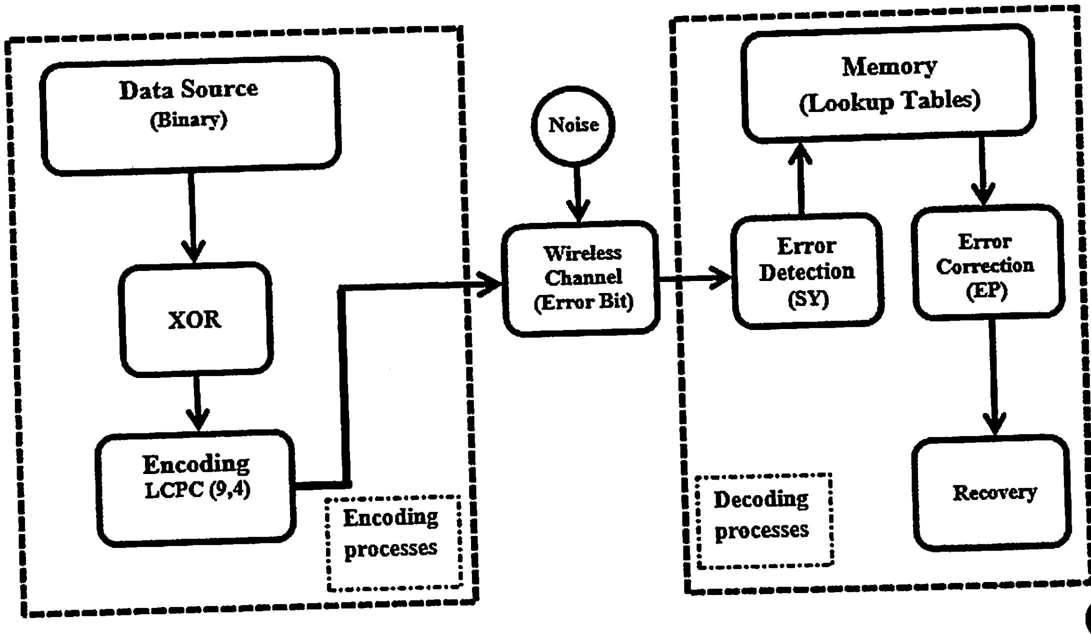


Figure 2.1: General block diagram of encoding/decoding LCPC

2.2.2 LCPC Encoding

Figure 2.1 shows the block diagram of encoding and decoding process of LCPC codes. To encode source data, firstly, divide the source data sequence into symbols of equal length k bits. Subsequently, each symbol (k bits) is taken and map it into a codeword c of n bits, where n > k. The n - k additional parity-check bits are the redundancies added, which used for error detection and correction. The LCPC code is a block code which takes the data stream from the source encoder, divides it into four-bit symbol, and then encodes each four bit symbol (depending on the number of rows in G matrix) into a nine-bit codeword (depending on the number of columns in G matrix), before the transmission. The symbol of source data is denoted as SDi = (v1,v2...vk), where 1 ≤i≤j, and j is the number of symbols of the source data, v is a binary bit, and k = 4 is the length of the symbol. Each k-bit symbol is then encoded into an n-bit codeword before the transmission, where the values of the n bits depend only on the values of the k bits in the corresponding symbol source block. In order to start the transmission and encoding of the source data, as shown in Figure 3.2, two samples from the source data are taken, for example (SD1 and SD2). The XOR operation is used to create SD3 symbol from SD1 and SD2. After the XOR, the encoding process is implemented using the generator matrixG, as represented by Eq. (2.1).

$$
\mathbf{G} = \begin{bmatrix} 1 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 \\ 0 & 1 & 0 & 0 & 1 & 1 & 1 & 0 & 1 \\ 0 & 0 & 1 & 0 & 1 & 1 & 0 & 1 & 1 \\ 0 & 0 & 0 & 1 & 1 & 0 & 1 & 1 & 1 \end{bmatrix} \qquad (2.1)
$$

In the encoding unit the redundant bits r is then added to each symbol to make the length of the codeword equal to n, where n = k + r, and r = 5. The codeword of the symbol corresponds to CDTi = (β1, β2 ... βn), where n = 9, and βi is a binary bit. A codeword CDTi that defined as a multiplication between SDi and G is given by Eq. (2.2).

$$
CD_{Ti} = SD_i \times G \qquad (2.2)
$$

SDi is an information symbol and CDTi is the transmitted codeword and G is the proposed generator matrix. The Generator matrixG also can be expressed in the form of a bipartite graph known as Tannergraph (Tanner,1981) as shown in Figure 2.3. The Tanner graph contains two sets of nodes: the first set of nodes represents the bits of a codeword (code nodes), and the second set of nodes implements the symbol bits (data nodes) in G matrix or implements the check bits (check nodes) in H matrix. The graph presents an edge between the code nodes and the data nodes if and only if the bit is involved in the parity-check equation. One benefit of Tanner graphs is used to construct and create a longer error correcting codes from a smaller ones (Tanner, 1981).

The data and code nodes in Figure 2.3 denote the rows and columns of the G matrix

respectively. An edge connects a data node to a code node if a non-zero entry exists at the intersection of the corresponding row and column. Each time, three codewords (CDT1, CDT2 and CDT3) for the three symbols (SD1, SD2 and SD3) respectively are sent. Figure 3.4 shows the encoded codeword construction of the

LCPC (9, 4) code. The five bits ($\gamma 1, \gamma 2, \ldots \gamma 5$) from the right side is the parity bits, while the four bits ($v1, v2, \ldots v4$) from the left side is the symbol bits. Eq.(3.3) and Eq.(3.4) show the symbol information bits and parity bits respectively for LCPC-a (9, 4) code.

$$
\begin{aligned}
\beta_1 &= v_1 \\
\beta_2 &= v_2 \\
\beta_3 &= v_3 \\
\beta_4 &= v_4
\end{aligned}
\tag{2.3}
$$

$$
\begin{aligned}
\beta_5 = \gamma_1 &= v_1 \oplus v_2 \oplus v_3 \oplus v_4 \\
\beta_6 = \gamma_2 &= v_1 \oplus v_2 \oplus v_3 \\
\beta_7 = \gamma_3 &= v_1 \oplus v_2 \oplus v_4 \\
\beta_8 = \gamma_4 &= v_1 \oplus v_3 \oplus v_4 \\
\beta_9 = \gamma_5 &= v_2 \oplus v_3 \oplus v_4
\end{aligned}
\tag{2.4}
$$

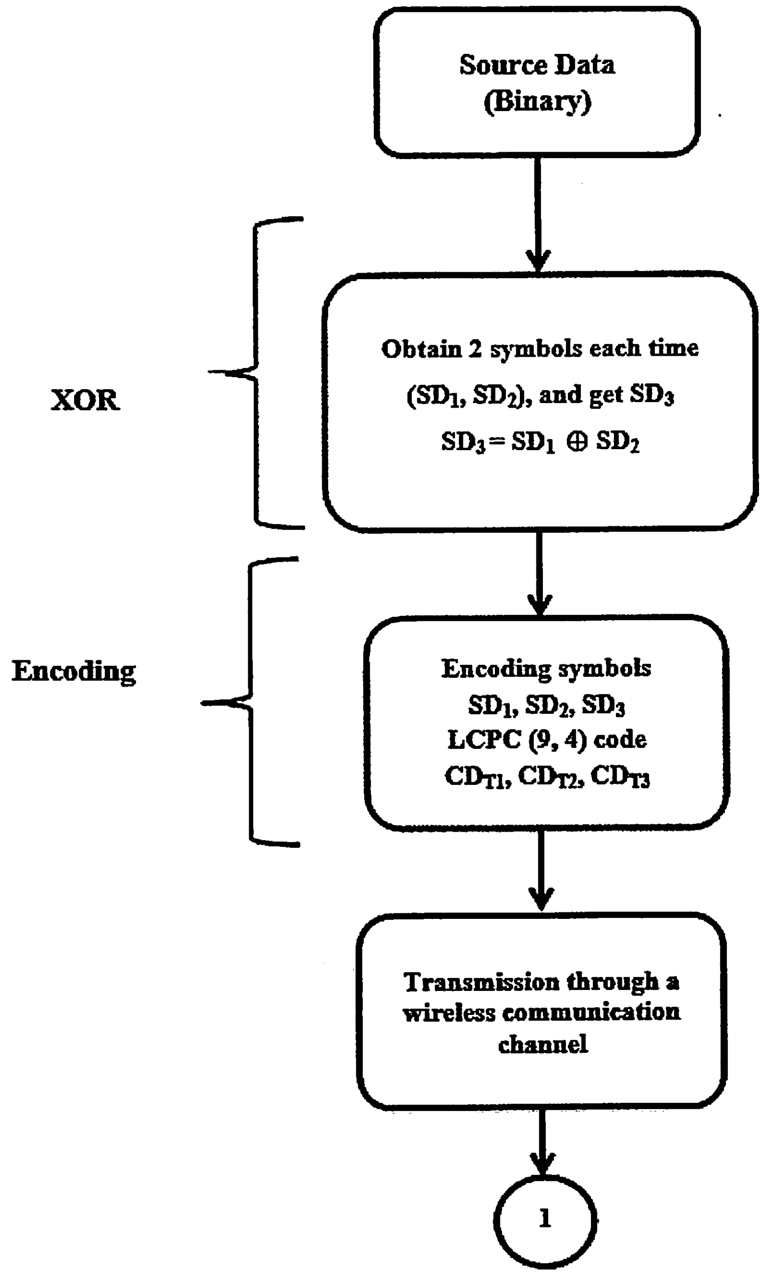


Figure 2.2 Flowchart of LCPC (9,4) encoding stage at the sender

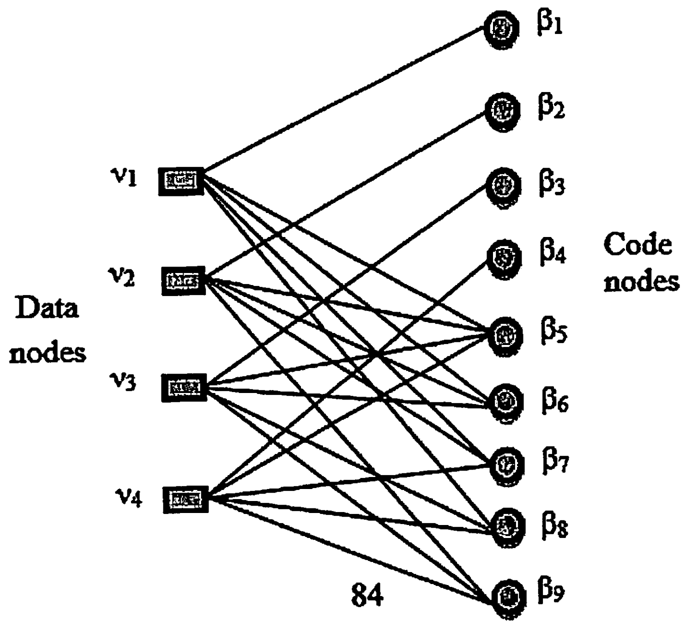


Figure 2.3: Tanner graph for LCPC (9,4)

Encoded Codeword

β_1	β_2	β_3	β_4	β_5	β_6	β_7	β_8	β_9
v_1	v_2	v_3	v_4	γ_1	γ_2	γ_3	γ_4	γ_5

Data

Parity

Figure 2.4: Codeword construction for LCPC-1(9,4) code

Table 2.2: Codewords of LCPC-a (9,4) code

Message Binary SD_i	Codeword Binary CD_{Ti}	Codeword Hex CD_{Ti}
0000	0 0000 0000	0 00
0001	0 0011 0111	0 37
0010	0 0101 1011	0 5B
0011	0 0110 1100	0 6C
0100	0 1001 1101	0 9D
0101	0 1010 1010	0 AA
0110	0 1100 0110	0 C6
0111	0 1111 0001	0 F1
1000	1 0001 1110	1 1E
1001	1 0010 1001	1 29
1010	1 0100 0101	1 45
1011	1 0111 0010	1 72
1100	1 1000 0011	1 83
1101	1 1011 0100	1 B4
1110	1 1101 1000	1 D8
1111	1 1110 1111	1 EF

The benefit of CDT3 is to use in the correction and recovery of the CDT1 and CDT2 if there are more than double bit errors in one of them, or there are two error patterns with the same value of syndrome vector.

The proposed parity check matrixHof the LCPC (9, 4) code is shown in Eq. (2.5)

that is used for error detection.

$$\mathbf{H}=\begin{bmatrix} 1&1&1&1&1&0&0&0&0\\ 1&1&1&0&0&1&0&0&0\\ 1&1&0&1&0&0&1&0&0\\ 1&0&1&1&0&0&0&1&0\\ 0&1&1&1&0&0&0&0&1 \end{bmatrix} \qquad (2.5)$$

The Tanner graph of H matrix is shown in Figure 2.5. The check and code nodes in Figure 2.5 denote the rows and columns of the H matrix respectively. An edge connects a check node to a codenode if a nonzero entry exists at the inter section of the corresponding row and column.

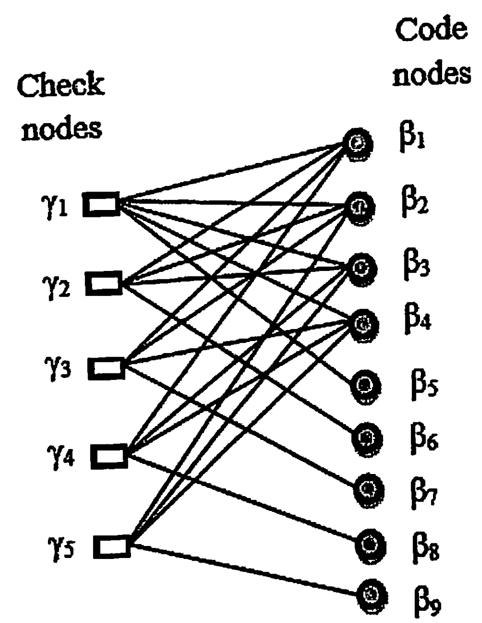


Figure 2.5: Tanner graph of a check matrix H for the LCPC-a (9,4) code

2.2.3 LCPC Decoding

The decoding algorithm consists of three parts. The first is to compute the syndrome for error detection. The second is determining the error pattern, and the third is the error correction and recovery of the native data. The aim of this part is to detect and correct the errors in the received codeword. The proposed LCPC codes have the capability to detect and correct single, double bit errors, and higher bit errors with some impose conditions. The main difference among the proposed LCPC codes and Hamming codes is that, the latter can detect and correct only single bit error,or it can detect double bit error without correction capability.

Table 2.3 shows the capability and limitations of the proposed LCPC (9, 4) code for errors detection and correction. LCPC (9, 4) code has the capability to detect and correct the bit errors when there are 1,2,7,8 and 9 bit errors in the code word received. Inaddition, it can correct 13 cases from the 84 cases for 3 and 6 bit errors, while being able to detect the error for 3, 4, 5 and 6 bit errors without correction. The reason of LCPC (9,4) being in capable of correction is that the number of possibility error pattern (NoEP) is more than the number of syndrome vector (SY) that the LCPC (9,4) do have (i.e., 32); the NoEP in cases of 3 and 6 bits error is 84, and it is 128 in 4 and 5 bits error.

The number of possibility error pattern (NoEP) and number of syndrome vector (NoSY) for LCPC (9,4) code is shown in Table 2.4. The NoEP is obtained from using Eq. (2.6), whereas the NoSY is obtained from Eq. (2.12) by taking into consideration all the possibilities of error pattern.

$$NoEP = \frac{n!}{e!(n-e)!} \tag{2.6}$$

Table 2.3: Capability of error detection and correction in LCPC (9,4) code

Number of bit error in the codeword	Error Detection	Error Correction
1	✓	✓
2	✓	✓
3	✓	×
4	✓	×
5	✓	×
6	✓	×
7	✓	✓
8	✓	✓
9	✓	✓

where n is the codeword length (in our proposed code n = 9), and e∈(1, 9) is the number of it errors that may occurred in the codewords. The proposed code assumes the look up tables that include the Epof each SY for1,2,7,8 and 9 bit errors are stored in the memory.

2.2.2 Binary Joint Network and Channel Coding (B-JNCC)

In this section, coding and decoding procedures of the proposed binary joint network and channel coding model (B-JNCC) are presented. Specifically, B-JNCC combines binary low complexity parity check (LCPC) channel coding and wireless network erasure correcting with router detect packets lost (WNC-RDPL) code. This combines assisted to exploit completely the spatial variability and redundancy in both channel and network codes. To understand the rationale behind the joint treatment of the channel and network coding, a simple example as shown in Figure 2.6 is proposed. In this simple topology, there are five routers distributed as two sources, two relays, and one destination. The two sources routers (i.e., *S*1 and *S*2) broadcast two independent packets P_1 and P_2, after encoding them using LCPC code, to the two relay routers *R*1 and *R*2 and to common destination router *D*. The functions of *R*1 and *R*2 are; encoded the packets that received using WNC-RDPL code, store the coded packets temporary, and then forward the packets to the destination router *D*. *R*1 and *R*2 generates $P_3 = (P_1 \oplus P_2)$ and forward it to the *D*. In Figure 2.6, we assume all routers operate in multi radio multi-channel (MRMC) techniques and that all communications link takes place over orthogonal channels such that mutual interferences can be neglected.

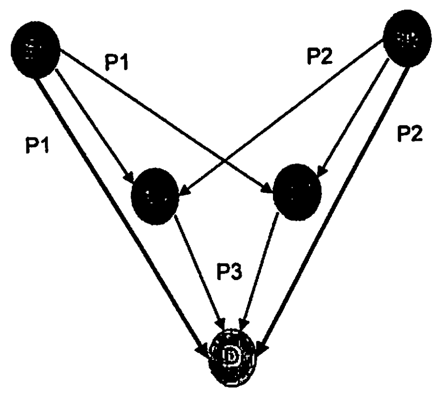


Figure 2.6: Simple example of two sources two relay one destination network topology

Accordingly, in the normal case (i.e., no link failure) the destination router D will receive four packets, two packets from the $S1$ and $S2$ and two packets from $R1$ and $R2$. On one hand, D received P_1 and P_2 from $S1$ and $S2$ respectively. On the other hand, D received double P_3 from $R1$ and $R2$. The B-JNCC will be implemented at the destination router D. At the D, the LCPC codes can detect and correct signal, double and more than double bit errors with some impose conditions. Whereas, WNC-RDPL code has the capability to recovery the packet that the LCPC code cannot corrected or the lost packets due to a link failure. The capability of detection and correction error packets when the B-JNCC used base in the topology that shown in Figure 2.6 is shown in Table 2.4. The desirable from the destination router D is received P_1 and P_2 correctly. If the D could not correct bit errors in P_1 or P_2 directly, the LCPC tries to correct P_3 that received it from $R1$ and $R2$. Then, D can use P_3 to retrieve the packet that cannot correct (i.e., P_1 or P_2).

In the worst-case, if LCPC cannot correct P_1, P_2 and P_3, in this case WNC-RDPL send NAK to $R1$ and $R2$, then the latter send the packets that received from $S1$ and $S2$ respectively (i.e., P_1 and P_2 without P_3) to the D. In this case, D will receive 2 packets of P_1 and 2 packets of P_2 from $R1$ and $R2$ respectively. Again, LCPC code detects and corrects those new packets that received to obtain the P_1 and P_2.

Figure 2.7 shows more details about the architecture and the functions of each router for our proposed B-JNCC that applied in the topology that shown in Figure 2.6. We assume all channels are lossy, and this is one difference between our work and other works that assume the channels between the source router and the relay router is lossless [6]. $S1$ generates CD_{T1} from encoding the symbol SD_1 (4 bits) using the

LCPC code, in the same case $S2$ generates CD_{T2} from encoding SD_2. Both $S1$ and $S2$ broadcast their encoded data to $R1$, $R2$ and D, over the lossy wireless channel. Due to the lossy channel, $R1$, $R2$ and D received CD_{Ri} that considered as;

$CD_{Ri} = CD_{Ti} + E$,where, CD_{Ti} transmitted codeword and E is the error pattern.

Table 2.4: Capability of Error Detection and Correction for B-JNCC at the destination

Index of Possibility	Possibilities received packets at the destination router (D)				States correction error packets at D
	P_1 from S1	P_2 from S2	P_3 from R1	P_3 from R2	
1	✓	✓	✓	✓	✓
2	×	✓	✓	✓	✓
3	✓	×	✓	✓	✓
4	✓	✓	×	✓	✓
5	✓	✓	✓	×	✓
6	×	×	✓	✓	✓ (NAK)
7	×	✓	×	✓	✓
8	×	✓	✓	×	✓
9	✓	×	×	✓	✓
10	✓	×	✓	×	✓
11	✓	✓	×	×	✓

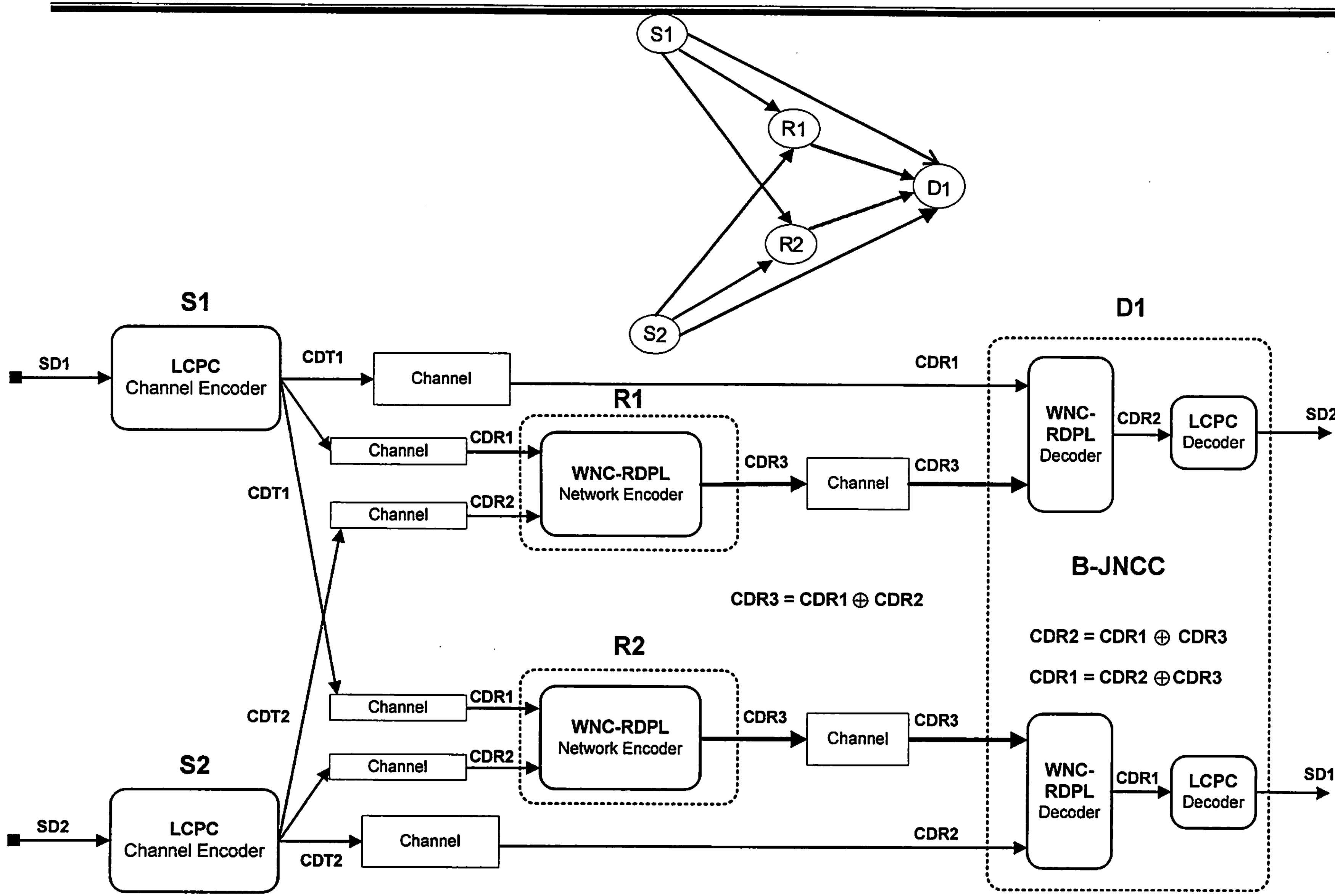


Figure 2.7: Architectural of B-JNCC for a simple example of two sources, two relay, one destination network topology

3. RESULTS AND DISCUSSION

3.1 LCPC (9,4) codes

Simulations are carried out to validate the performance of the proposed LCPC (9,4) code. The codeword generated from the LCPC (9, 4) code is modulated using BPSK, 4-QAM and 16-QAM modulation schemes before being transmitted via wireless channels such as additive white Gaussian noise (AWGN) channel and Rayleigh fading channel. The performance of the LCPC(9,4) code is compared with Hamming (Moon, 2005), RS (Wicker and Bhargava, 1999), and LDPC (Gallager, 1963) codes using various values of codeword length. The Hamming, RS, Golay, BCH Soft and BCH Hard codes are modulated in MATLAB. The comparison is implemented at different codeword length to investigate the BER performance of proposed LCPC codes at different cases. Some decoding algorithms of LDPC (8, 4) code such as bit flip, log domain and log domain simple are also are modulated using MATLAB program.

Figure 3.1 presents the simulated BER versus SNR plots of LCPC (9, 4), Hamming (15, 11) and RS (15, 11) codes, with the Shannon limit over AWGN channels and BPSK modulation. The results clearly show that the performance of the LCPC (9,1324) code is much better than Hamming and RS codes, because the proposed LCPC code has the capability to corrects double and more than double bit errors. It is also worth noting that, for the LCPC (9, 4) code, BER value is 10−6 at SNR of 6 dB, while for RS and Hamming codes, SNRs are 8.2 dB and 9.1 dB respectively, to obtain BER of 10^{-6}. This shows that LCPC (9, 4) code reduces the transmission power requirement to obtain a BER 10^{-6} and therefore showcases significant advantage over other codes. Besides improving the system performance, the proposed code reduces the complexity in the decoder compared to the LDPC codes.

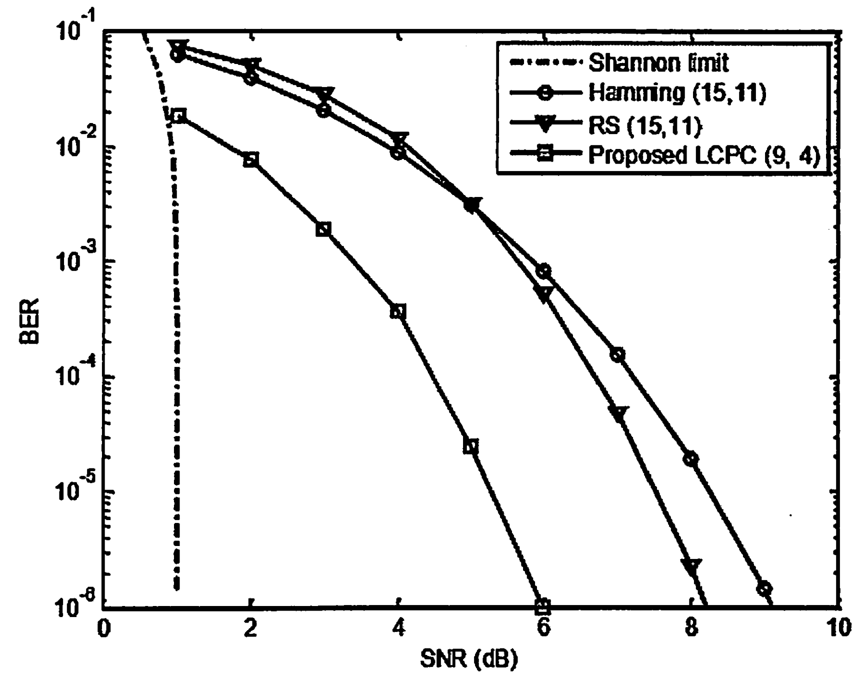


Figure 3.1: Comparison between LCPC (9,4) code with Hamming and RS codes in AWGN channel.

Figure 3.2 presents the comparison between LCPC, Hamming and RS codes at codeword lengths (7, 4), over AWGN channels using BPSK modulation. The Hamming (7, 4) and RS (7, 4) are modulated using MATLAB program when the n = 7, k = 4, and the code rate R = 0.571. The Figure 3.2 shows that the LCPC code improves the BER performance when compared with the Hamming and RS codes. The LCPC (9,4) code have the capability to correct two and more than two bit errors, whereas the Hamming (7, 4) has the capability to correct one bit error and RS (7, 4) codes has the capability to correct one bit error in each symbol.

Figure 3.3 shows more accurate results of the performance of BER for the LCPC (9, 4) code stay better than the RS codes, especially in the parity code (6, 8, 10, 16 And 32) bytes. The performance of the proposed LCPC (9,4) code is investigated with different codeword length of RS codes over AWGN channels using BPSK modulation.

As the graph shows, the BER performance of RS code is becoming better when the amount of parity code increases. For illustration, RS (255,223) at parity code 32 bytes are better than RS (255, 2249) at parity code 6 bytes. Nevertheless, the LCPC (9, 4) code is better than RS codes. Further, increasing the parity code increases the complexity of detection and correction of errors, which in turn increases the time delay. The LCPC (9, 4) code has a good BER performance, because the LCPC has the capability to correction double bits and more than double bits. In addition,the LCPC codes have the capability to correct the burst error.

The capability of the proposed code (LCPC) in error detection and correction is also studied. The minimum Hamming distance defined as dmin = n - k, where m $\geq$ 3 is a positive integer. The number of errors that a block code can detect and correct is determined by its minimum Hamming distance dmin. This is defined as the minimum number of places where any two codewords differ. Ingeneral, the number of errors (u) that can be detected for a block code is u = d - 1. For example at m = 3, the codeword length n = 7, message length k = 4 and dmin = 3, where t is the number of errors that a block code can correct, $t = \left[\frac{n-k}{2}\right] = 1$. Since the Hamming code has a minimum Hamming distance dmin = 3, it can only correct 1 bit error for each 7 bits transmitted.

Therefore, the error correction is 1/7 = 14.285 %.

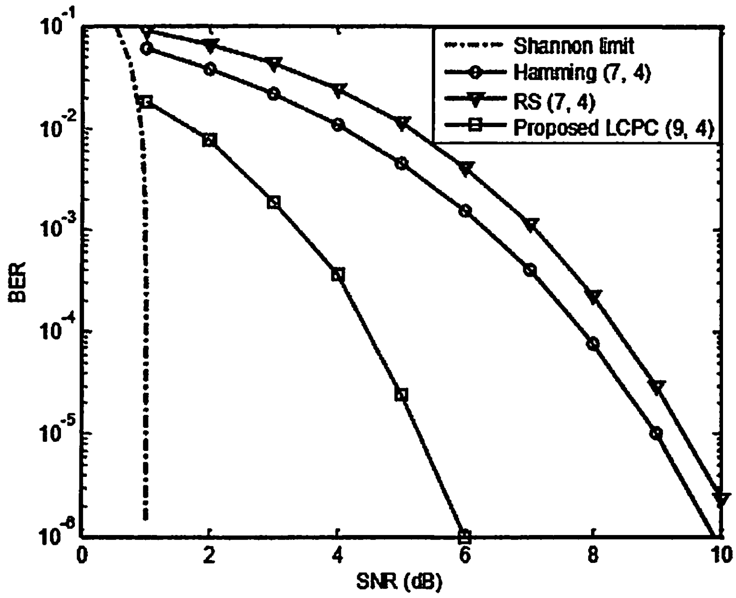


Figure 3.2: Comparison between LCPC (9,4) code with Hamming (7,4) and RS (7.4) codes in AWGN channel

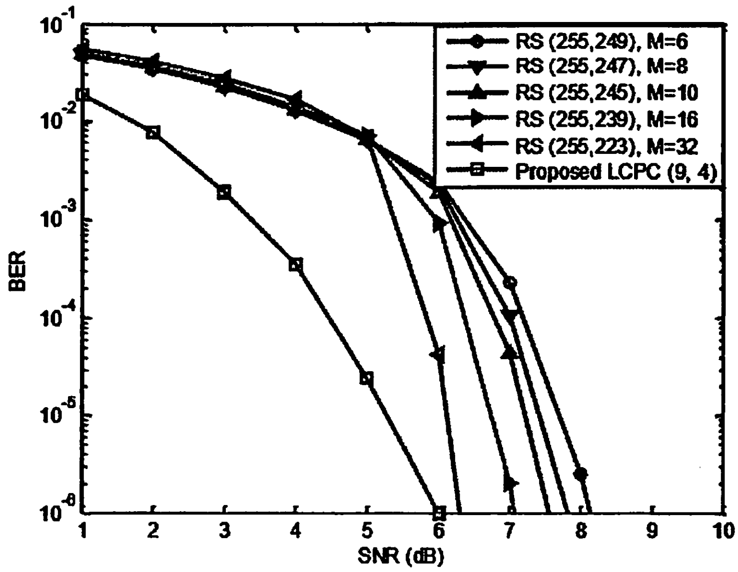


Figure 3.3: Comparison between LCPC (9,4) code with different RS codes over AWGN channel

Likewise, in the case of Reed-Solomon (RS) codes, the number and type of errors that can be corrected depends on the characteristics of the RS code. RS codes are a subset of Bose-Chaudhuri-Hocquenghem (BCH) codes and are linear block codes (Moon, 2005). RS codes are burst error-correction codes. A Reed-Solomon code is specified as RS (n, k) with s-bit symbols. This mean that the encoder takes k data symbols of s bits each and adds parity symbols to make an n=2s-1,symbol codeword. There are n-k parity symbols of each s bits. An RS decoder can correct up to t symbols that contain errors in a error codeword, where 2t = n - k. If s = 3 bits, n = 7, and when the number of parity is 3, k=4. The number of symbols containin g errors that RS code can correct is t, where, $t = \left\lceil \frac{n-k}{2} \right\rceil =$ 1. So, based on t value the RS (7, 4) code can only correct one symbol error from the 7 codeword symbols that are transmitted. If the symbol sizes are 3 bits, the worst case happens only when a one-bit error occurs in separate symbols. In this case the error correction is 1/21 = 4.7619 %, which is small compared with the percentage of error correction in Hamming (7, 4) code, and this explains the reason why the Hamming (7, 4) code has a better BER performance when compared to the RS (7,4) code, as shown in Figure 3.4.

The best case for RS (7, 4) code occurs when all bits are errors in a one symbol. This mean that the percentage of error correction is the number of errors in one symbol over the total number of symbols bit transmitted, i.e. (3/21 = 14.285 %). Compared with our proposed code LCPC the worst case happens only when there is a one bit error in each codeword (9 bits). According to this approach, there are three codewords transmitted each time. Two codewords contain the original source data (SD1 and SD2) and the other codeword (SD3) is the outcome of XOR operation involving the two original source data. The benefit of sending the third codeword (CDT3) is to use it for correcting the error when there are more than two bit errors in one codeword that contains the original source data (CDT1 orCDT2). In this case, when there is a single bit error the error correction obtained is 2/18 = 11.11 %. On the other hand the error correction is 22.22 % when there are 2 bit errors in each codeword (CDT1 and CDT2). The best percentage of error correction is achieved when there are two bit errors in one codeword and 9 bit errors in another codeword (i.e.,11/18=61.11%). This clearly shows that the LCPC code is better than Hamming, BCH, Golay and RS codes.

To increase the capability of error correction, the number of the parity codes must increase. This mean that the value of t in RS code must be large. A popular Reed Solomon code is RS (255, 223) with 8-bit symbols (Carrasco and Johnston, 2009; Jiang, 2010). Each block contains 255 codeword bytes, of which 223 bytes are data and 32 bytes are parity. For this code: $n = 255$, $k = 223$, $s = 8$, $2t = 32$, and $t = 16$. A large value of t mean that a large number of errors can be corrected but requires more computing power than a small value of t. One symbol error occurs when a 1 bit in a symbol is wrong or when all the bits in a symbol are wrong. RS (255,223) decoder can correct 16 symbol errors. The worst case occurs when a 16 bit error occurs, and each one of these bits is in a separate symbol (byte), so that the decoder can be corrected as maximum 16 bit errors from the 2040 bits (255×8); the error correction is 16/2040 = 0.784 %. The best case occurs when a 16 complete byte error happens so that the decoder can correct a 128 bits (16×8) error from the 2040 bits; the error correction is 128/2040=6.274%. There are clear restrictions and limitation in the number of errors that the decoder can correct in each block, and this limitation depends on the number of parity byte.

The difference between the LCPC code and RS code is that the latter has the ability To correct a specific number of bit errors based on the parity number,whereas the LCPC code cannot do so. As an example in LCPC code, if 2040 bits (255 byte) are needed to be sent, these 2040 bits are segmented to nearly 227 blocks, each block containing 9 bits. So, the worst case is that the LCPC code can correct 1 bit in each one block (9 bits). Since there are 227 blocks, the error correction is 227/2040 = 11.127 %, and if there are two bit errors in each block the error correction is 454/2040 = 22.25 %. The best case is when half of 227 blocks have two bit errors, and another half of the 227 block have 9 bits error as the maximum. Therefore, the total number of errors that LCPC code can correct is: (227/2)×2 = 227 bits (for 2 bits errors) (227/2)×9 = 1021 bits (for 9 bits errors) The error correction is (227+1021)/2040 = 61.176 %. Thus, the proposed code LCPC outperforms the renowned codes. Moreover, Figure 3.4 presents the comparison between LCPC (9, 4) code and Hamming, Golay, BCH (Soft, Hard) and RS codes which modulated using MATLAB program with the Shannon limit over AWGN channels on BPSK modulation. It is obvious that the performance of LCPC (9, 4) code is better than the other codes. As shown in Figure 3.4 to get BER 10^{-5}, needs 5.6 dB for a LCPC (9, 4) code. The code gain is 1.9, 2.5, 3.4, 3.6, 3.9 dB for Golay, BCHsoft, Hamming, BCHsoft and RS codes respectively. The code gain defined as the amount of additional SNR or Eb/No that would be required to provide the same BER performance for an uncoded signal. Figure 3.5 presents the coding gain for LCPC (9, 4) code compared with the Hamming and RS codes in AWGN channel. It is evident that the coding gain for LCPC (9, 4) code increases when the codeword length of Hamming and RS codes decrease. The minimum code gain is equal to 1.4 dB at codeword length 255, whereas it is equal to 3.5 dB and 4 dB for Hamming and RS codes respectively at codeword length 7 when the BER is 10^{-5}. The saving power is one of the benefits of LCPC(9,4) code. Therefore, the proposed code can be effectively used in wireless sensor network (WSN) because of its huge reduction in the power consumption, and saves the battery life to a large extent which is a very important consideration in WSN.

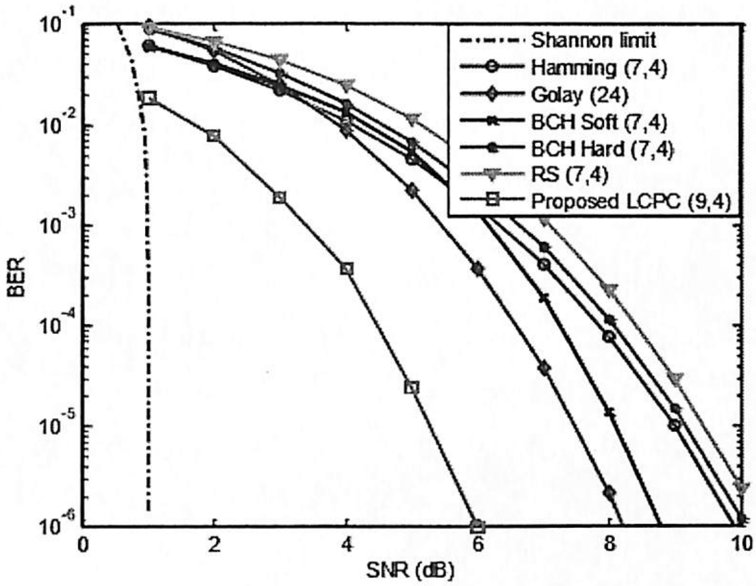


Figure 3.4: Comparison between LCPC (9,4) code and other codes with Shannon limit over AWGN channel

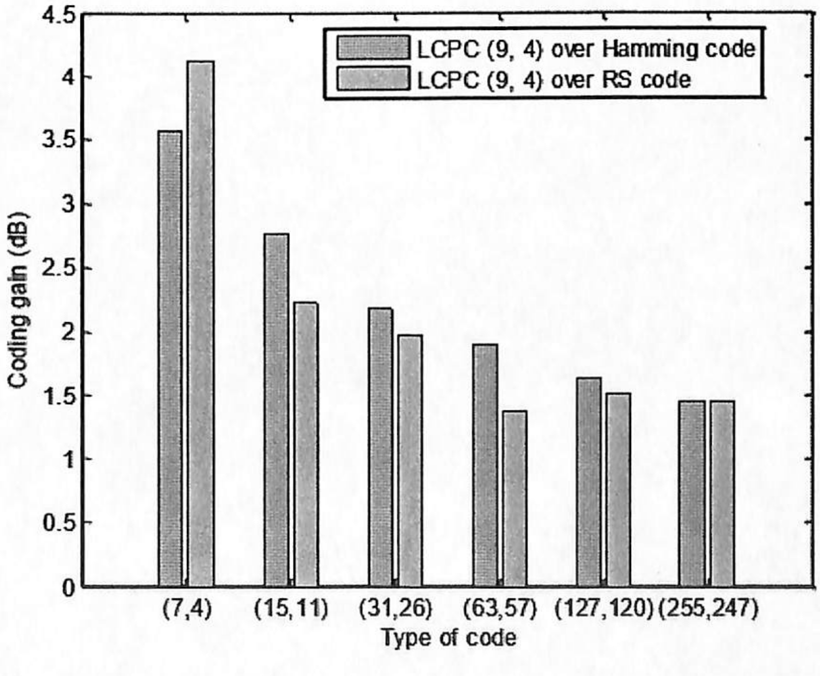


Figure 3.5: Coding gain for LCPC (9,4) code over Hamming and RS codes at BER 10^{-5}

Similarly, Figure 3.6 presents the BER performance comparison between LCPC (9, 4) code and LDPC (8, 4) code by various decoding algorithms over AWGN channels using BPSK modulation. The LDPC (8, 4) code is modulated using MATLAB program with the code rate is 0.5 for three different types of decoding algorithms (i.e., bit flip, log domain, and log domain simple). The performance of LCPC (9, 4) code is better than LDPC code when the bit flipping decoding method (BF) is used, and the coding gain is around 5dB at BER equal10^{-5}. However, the BER performance of the LCPC (9, 4) code becomes same to the LDPC code that uses log domain and log domain simple decoding methods at 3 dB SNR. The error correction capability of LDPC code depends on the codeword length and the characteristic of the parity check matrix. The decoder gives a better performance with a larger codeword (i.e., big size of G and H matrices), The matrix multiplication for that big codeword size demands huge memory, computational requirements and more complex decoding(Kou et al., 2001; Carrasco and Johnston, 2009).

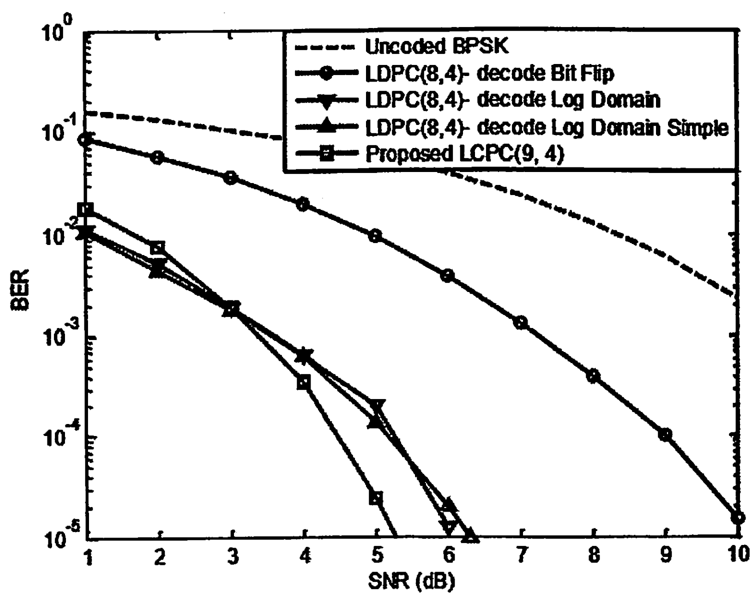


Figure 3.6: Comparison between LCPC (9,4) code and LCPC (8,4) at different types of decoding over BPSK AWGN channel.

The BER performance of the LCPC (9, 4) code is also compared with Euclidean Geometry-Low Density Parity Check (EG-LDPC) (255,175) code (Kou et al., 2001) by various decoding methods over AWGN channels using BPSK modulation as shown in Figure 3.7. The different types of decoding methods used are bit flipping decoding (BF), one-step majority-logic (MLG) decoding, weighted MLG decoding, and weighted BF decoding (Kouetal.,2001). Figure 3.7 shows that the BER performance of LCPC (9, 4) code is better than that of (EG-LDPC) (255,175) code at low SNR. On the other hand, at SNR 4.5 dB the BER performance of the LCPC is same as the performance of the EG-LDPC code when the EG-LDPC weighted BF decoding method is employed.

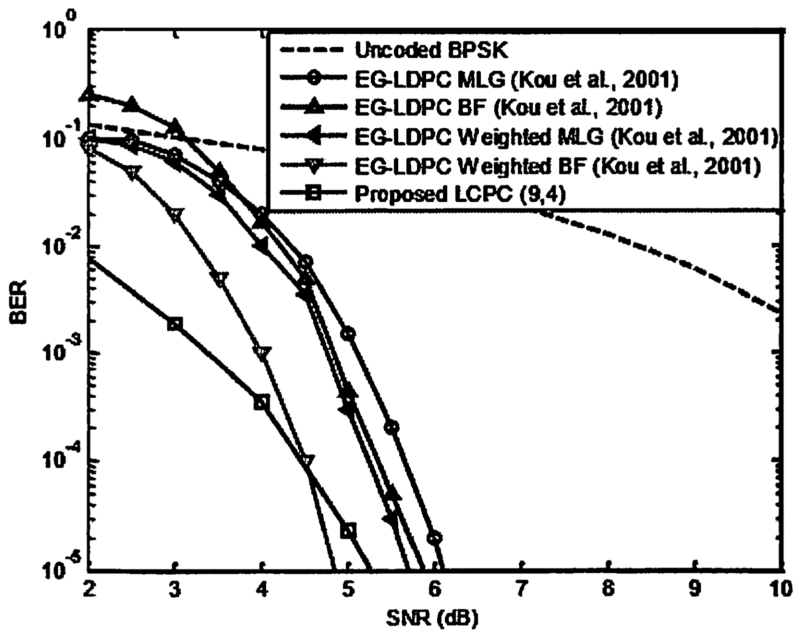


Figure 3.7: Comparison between LCPC (9,4) code and LDPC (255,175) at different types of decoding over BPSK AWGN channel

Moreover, the LCPC (9, 4) code is investigated when BPSK and M-QAM modulations are used over AWGN channel. Figure 3.8 presents the BER performance of the LCPC(9, 4) code with BPSK,4-QAM and 16-QAM modulation respectively. The results show that the performance of LCPC (9,4) code in the 4-QAM is better than 16-

QAM modulation, and the BPSK modulation is better than the 4-QAM and 16-QAM modulations. The reason of the BPSK modulation is better than QAM modulation is that in BPSK each one bit send per one symbol, whereas in 4-QAM each two bits send per one symbol and in 16-QAM each four bits send per one symbol. This increase the probability of error and number of bits error.

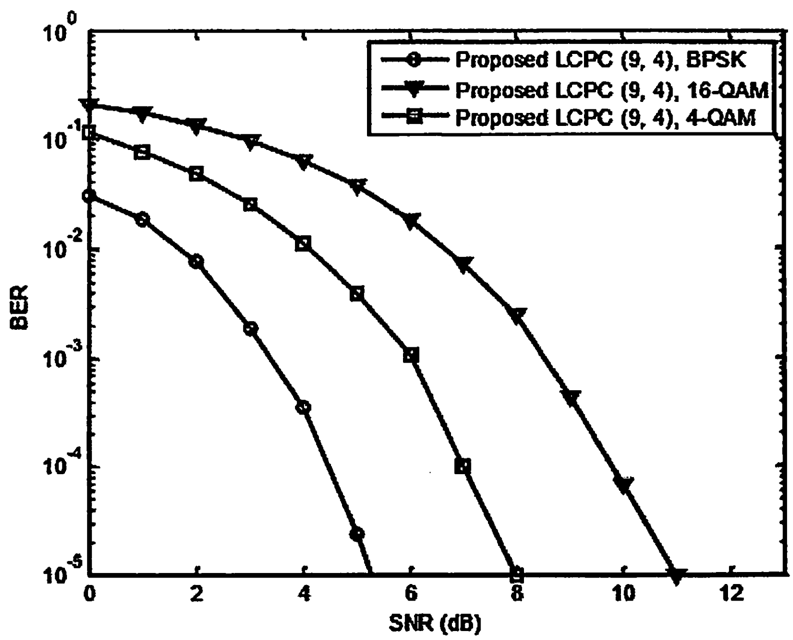


Figure 3.8: BER versus SNR for LCPC (9,4) code with BPSK, 4-QAM, and 16-QAM modulations over AWGN channel

Similarly, the BER performance of the LCPC (9, 4) code is studied when BPSK modulation is used over Rayleigh fading channel for data transmission 1024 bits. Figure 3.9 presents the BER performance of the LCPC(9,4) code for BPSK modulation over AWGN and Rayleigh fading channels for data transmission 1024 bits. It can be seen clearly that the LCPC code has better performance in both AWGN and Rayleigh fading channels. In addition to this, Figure 3.10 shows the BER performance comparison between the LCPC (9, 4) code and LDPC (576, 288) code in BPSK modulation over

Rayleigh fading channel. It is observed that the LCPC (9, 4) code outperforms the LDPC code (576, 288) over Rayleigh fading channel.

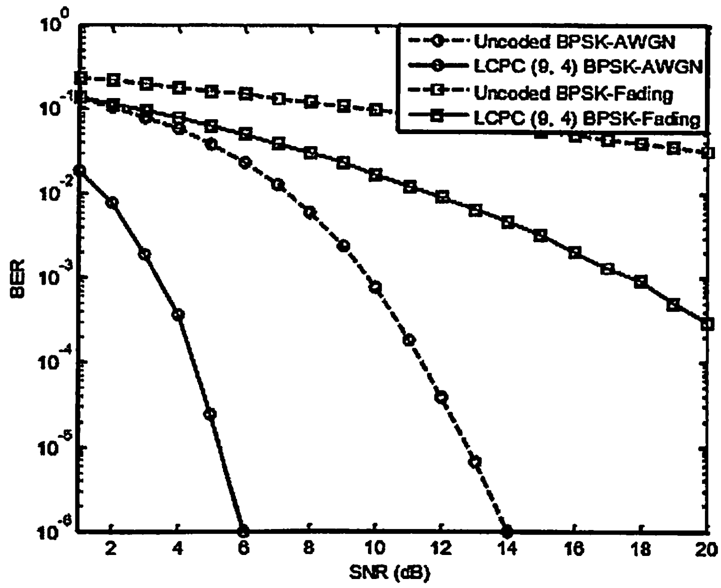


Figure 3.9: BER versus SNR for LCPC (9,4) code with BPSK modulation over AWGN and Rayleigh fading channels

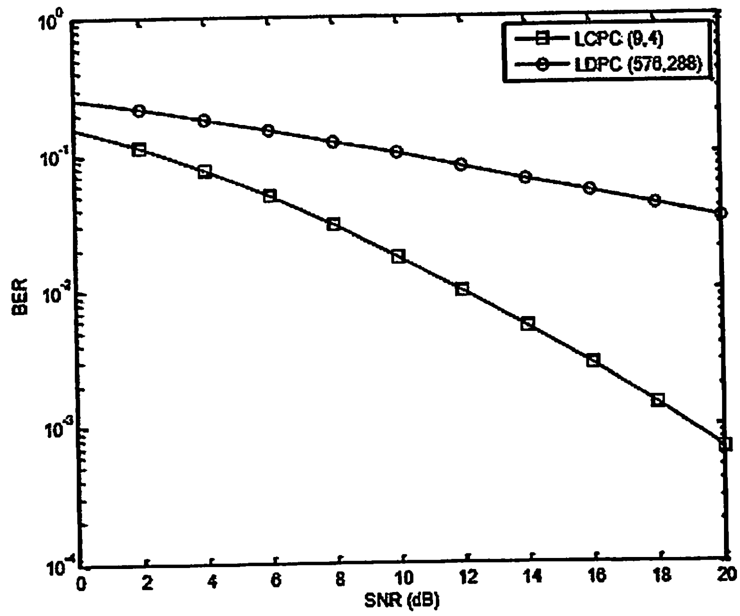


Figure 3.10: Comparison between LCPC (9,4) and LDPC (576, 288) codes in BPSK over Rayleigh fading channel

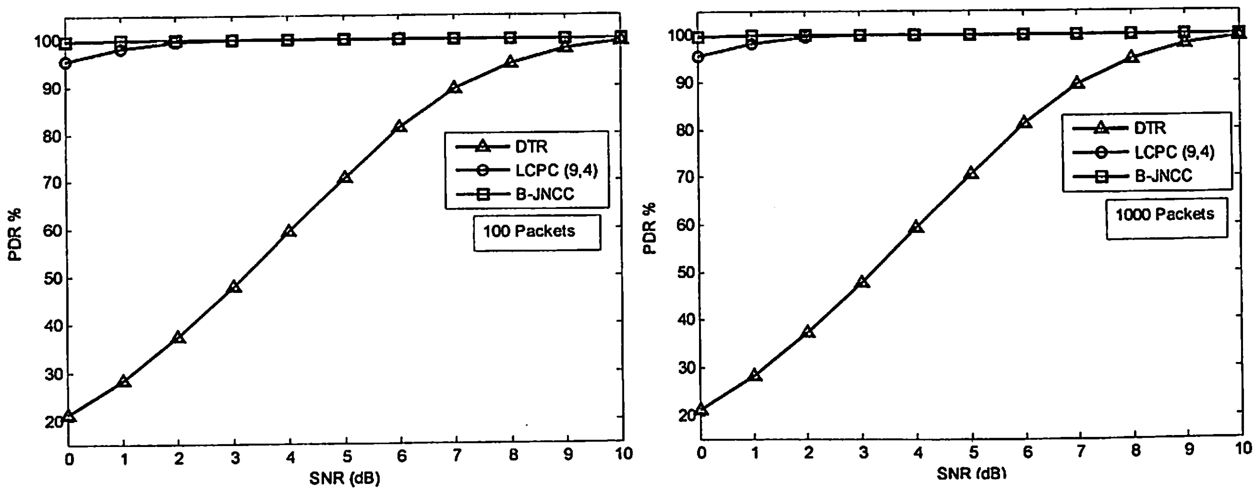
3.2. Wireless Network Coding

In this section, based on the topology that shown in Figure 5.4, the performance of B-JNCC is evaluated under AWGN with BPSK modulation. The proposed B-JNCC is investigated by computing the packet delivery ratio (PDR), bit error rate (BER) and block error rate (BLER). The BLER here means the number of blocks that has at least one error bit. The length of the block is 9 bits (i.e., codeword length) based on the LCPC (9, 4) code.

The performance of the B-JNCC is compared with two schemas; first, is direct transmission with relays (DTR), DTR here means transmission without coding. Second is LCPC channel coding. In the DTR case, the source directly transmits packets to the destination, and in addition relay routers forwards the packets received from source to the destination. Verification is done whenever the proposed systems are simulated using computer programs.

3.2.1 Packet Delivery Ratio (PDR)

The PDR defined as a ratio between the numbers of correct packets received at the destination to the total number of sent packets. The performance of B-JNCC is investigated in two cases; one at low traffic (send 100 packets), and the second at high traffic (send 1000 packets) at different values of SNR in AWGN channel with BPSK modulation. Figures 3.11(a) and 3.11(b) show the PDR-SNR curve for DTR (without coding), LCPC code, and B-JNCC in case a 100 and 1000 packets sent respectively. It can be seen that along with the increment of SNR, the PDR of the three systems increment to different extents. However, the PDR reached to around 100 % in case LCPC and B-JNCC at low SNR values (3 dB), whereas in case DTR the PDR becomes 100 % at SNR > 11 dB. The Figures also demonstrate that the values of PDR in case B-JNCC is better than LCPC at low values of SNR (< 2 dB).

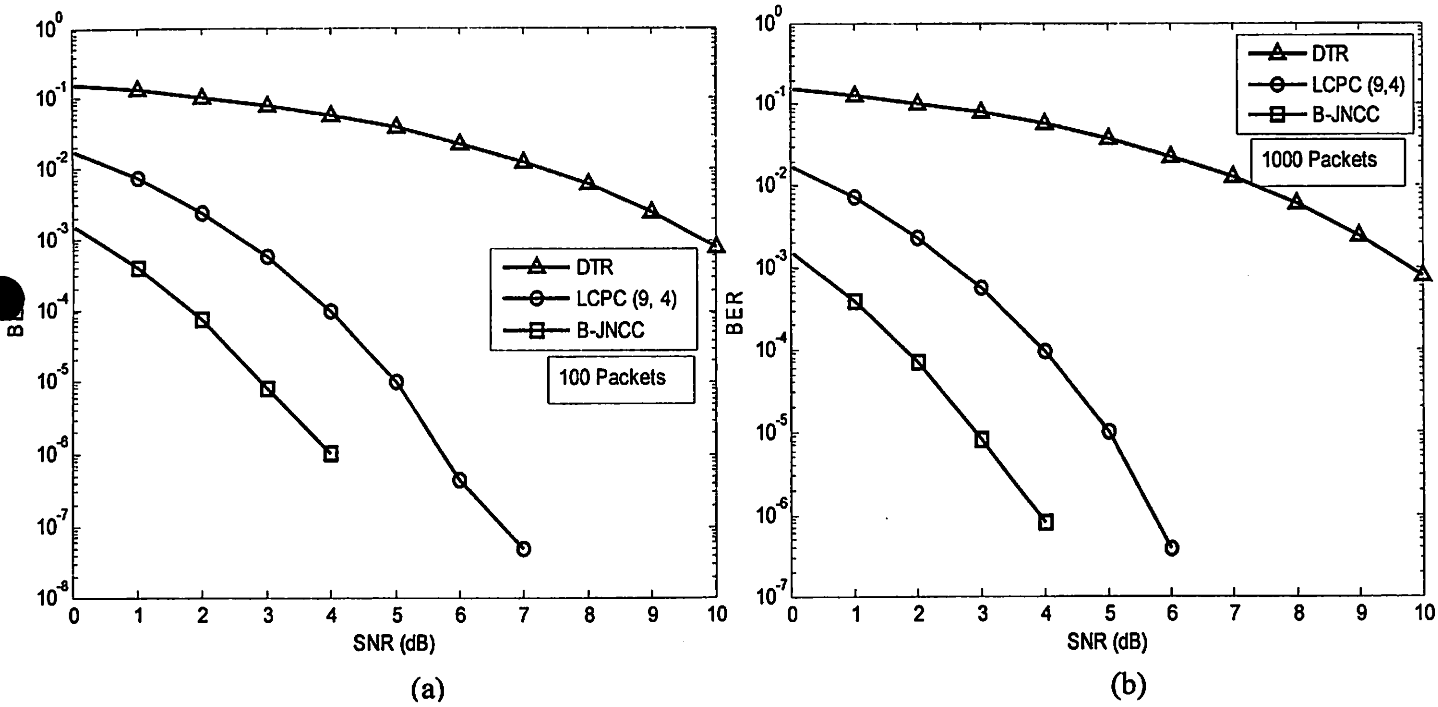


(a) (b)

Figures 3.11: PDR versus SNR at AWGN with BPSK, (a) 100 packets, (b) 1000 packets

3.2.2 Bit Error Rate (BER) and Block Error Rate (BLER)

In addition to the PDR performance evaluation, the performance of BER and BLER are investigated for the B-JNCC at 100 and 1000 packets send at different values of SNR under AWGN channel with BPSK modulation. The Figures 3.12 (a, b) illustration BER versus SNR for DTR (without coding), LCPC code, and B-JNCC in case a 100 and 1000 packets sent respectively. Figures 3.12 (a,b)clearly show the BER of the three systems reduce to different extents along with the increment of SNR. However, the performance of BER in case B-JNCC is better than DTR and LCPC, also the BER of B-JNCC and LCPC code dropped rapidly when the SNR is getting larger . On the other hand, the code gain that obtains from using B-JNCC is around 2 dB compare with the LCPC code at BER 10^{-5} in case 100 packets sent.



Figures 3.12: BER versus SNR at AWGN with BPSK, (a) 100 packets, (b) 1000 packets

3.2.3 Performance evaluation comparison with various schemes

The performance of B-JNCC over the two-source two-relay topology using MATLAB simulation is also evaluated. Comparison is made between the B-JNCC with other schemes; binary symbol-wise joint network-channel coding (BSJNCC) (Guo, et. al., 2012). Non-binary joint network-channel coding (NB-JNCC) (Guo, et. al., 2012). Non-binary separate network-channel coding (NB-SNCC). NB-SNCC schemes treat channel codes and network codes separately, where channel decoding is followed by network decoding with no iteration (Tran, et. al. 2008), (Berger, et. al., 2018). Direct transmissions with relays (DTR), where the sources directly transmit packets to the sink, and relay forwards the packets received from source to the sink. Direct transmissions without relays (DT), where the sources directly transmit packets to the sink, and the relays do not forward any packets.

To demonstrate the benefits of B-JNCC, extensive simulation results are presented, base in the network topology. In our simulation, the LCPC (9, 4) code and WNC-RDPL code are joint together. Simulation results obtained using both BPSK and 16QAM modulation under fading slow channel. Figure 5.10 shows compare the packet error rate (PER) of the various schemes when varying the average received SNR under BPSK modulation. Figure 3.13 shows that B-JNCC outperforms other schemes such as BS-JNCC, NB-SNCC and NB-JNCC (Guo, et. al., 2012). For example at PER of 10^{-3}, B-JNCC outperforms BS-JNCC by about 10 dB, and outperforms the NB-SNCC by about 9 dB and NB-JNCC by about 7 dB.

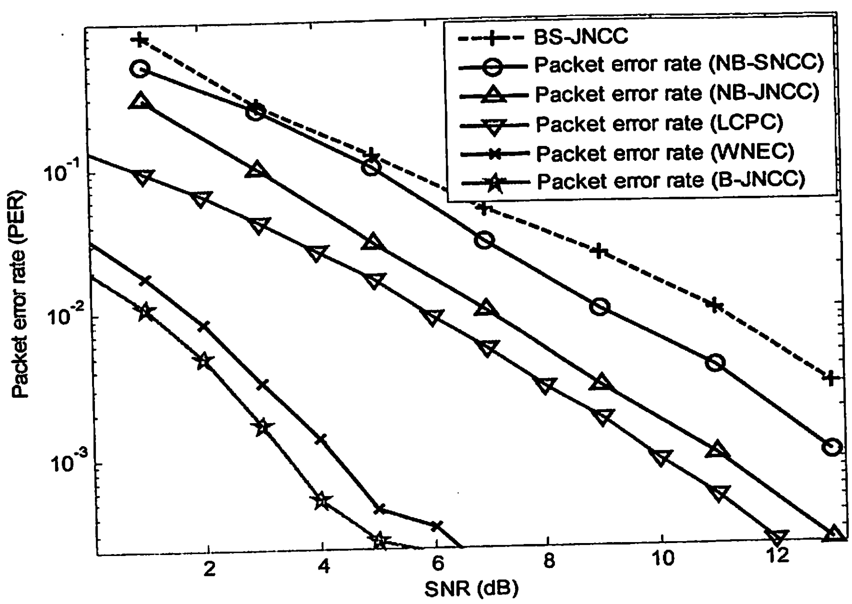
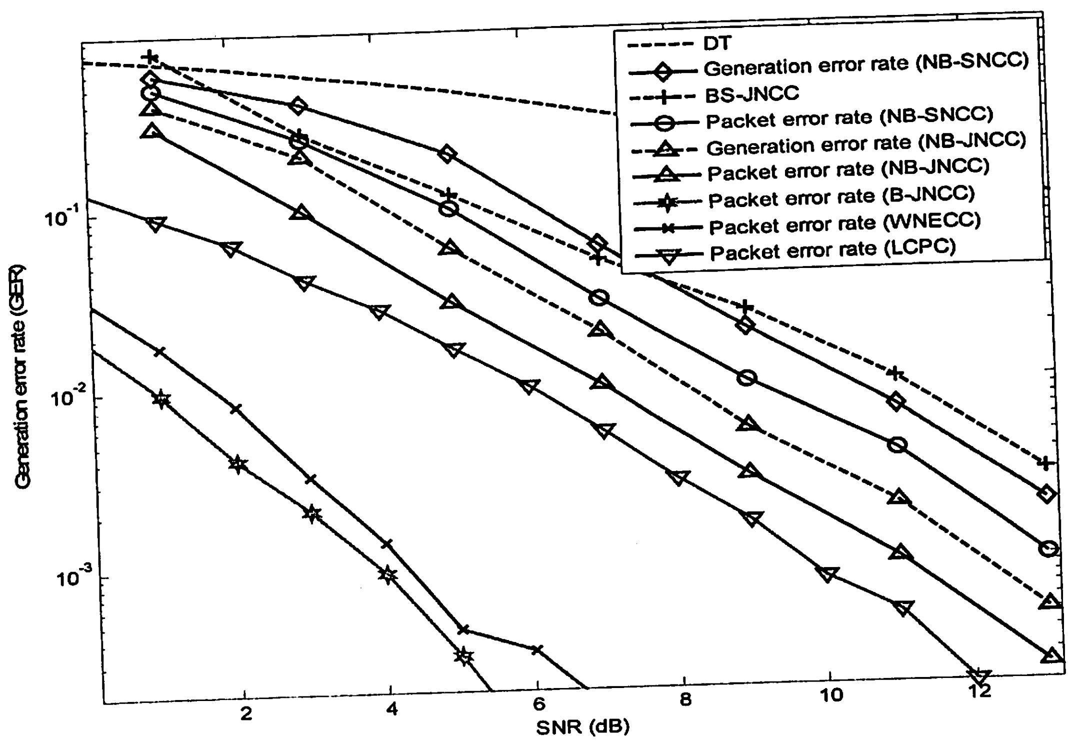
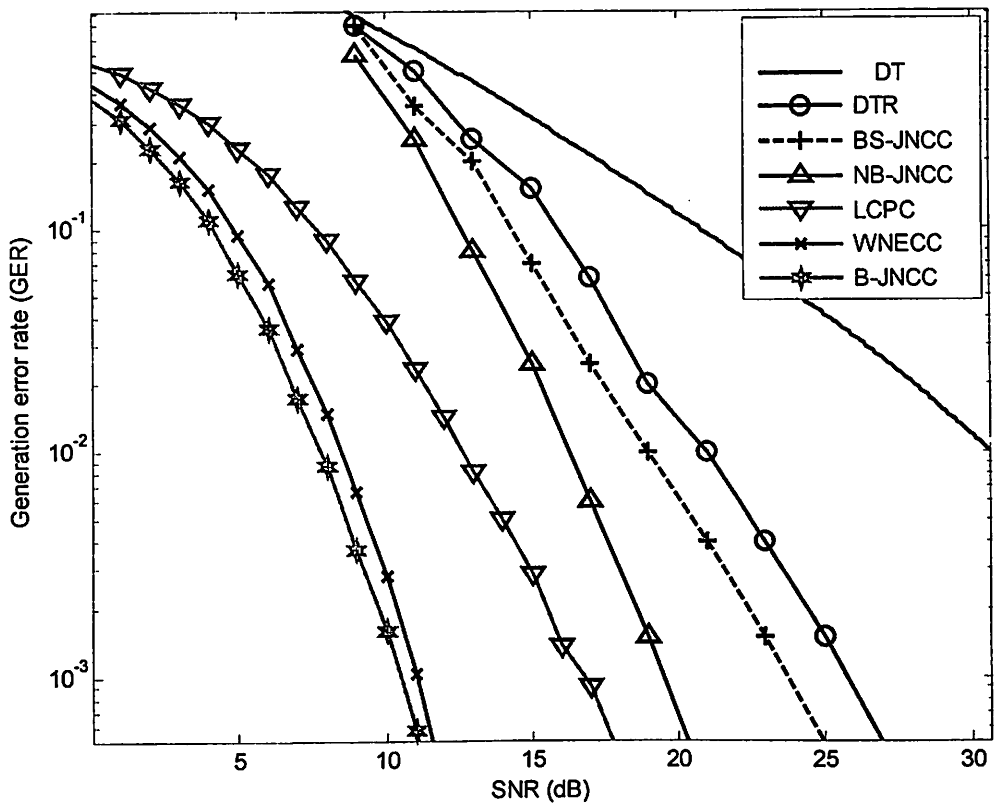


Figure 3.12: Packet error rate of B-JNCC and other schemes under fading channel, BPSK

Figure 3.13(a) and (b) compare the GER of the B-JNCC with various schemes at slow fading channel under BPSK and 16QAM modulations. Figure 3.13(a) presents the results with BPSK modulation. The Figure shows that B-JNCC outperforms with other schemes under BPSK modulation. For example at GER of 10^{-3}, B-JNCC outperforms BS-JNCC by about 11 dB, NB-JNCC by about 9 dB and LCPC by about 7 dB. Figure 3.13(b) presents the results with 16QAM modulation. Also, under 16QAM modulation B-JNCC is better than other schemes. Figure 3.13(b) shows that B-JNCC outperforms with other schemes. For example at GER of 10^{-3}, B-JNCC outperforms DT by about 25 dB, DTR BY 15 dB, BS-JNCC by about 14 dB, NB-JNCC by about 7 dB and LCPC by about 6 dB. Simulation results have demonstrated that the significant benefits of B-JNCC compared to other schemes which present in [6].



(a) GER under BPSK



(b) GER under 16QAM

Figure 3.13: Generation error rate of B-JNCC and other schemes under BPSK and 16QAM modulations

4 CONCLUSION AND SUGGESTION

4.1 Conclusion

In this project,a block linear code for error detection and correction has been developed identified as low complexity parity check (LCPC) code with different values of code rates. In addition to detecting and correcting single and double bit errors, the LCPC code can correct more than double bit errors in one codeword if the other two codewords have one or two bit errors. The performance of the proposed LCPC code in two types of communication channels (AWGNand Rayleigh fadingchannels)and threetypes of modulations (BPSK, 4-QAM and 16-QAM) are investigated. BER performance comparisons are made between the LCPC code and other codes such as Hamming, RS, BCH, Golay and LDPC codes. The simulation results are based on the RS and LDPC codes, at different transmission data and codeword lengths. The results indicate that the BER performance of LCPC code outperforms the renowned codes. Besides, a good value of coding gain is determined when comparing the BER performance of LCPC code with Hamming and RS codes. The results indicate that the coding gain decreases when the codeword length increases. The maximum coding gain is 3.5 dB and 4 dB for Hamming and RS codes respectively at codeword length 7 when the BER is 10−5 whereas the minimum coding gain is 1.4 dB at codeword length 255. However, some types of decoding for LDPC code have a better BER performance compared to the proposed code over AWGN channels such as EG-LDPC weighted BF decoding method. On the other hand, the LCPC code has characteristics that distinguishit from LDPC codes, such as low complexity encoding and decoding O(n), low demand for memory for matrix multiplication in case of big codeword size, and non requirement of any iterations in decoder when compared with LDPC codes that need more than 40 time of iterations to correct the error codeword. Furthermore, the proposed LCPC code has a better BER performance when compared with the LDPC code over Rayleigh fading channel at different block code length and at different values of data transmitting. The benefits of the proposed code is expected to improve the network performance such as increasing throughput, reducing end-to-end delay and bit error rate for real time applications.

Additionally, this project also developed a wireless network coding (WNC) technique and router detection packet loss (RDPL) algorithm. WNC-RDPL technique allows retrieving the lost packets in multi radio multi-channel wireless networks. WNC-RDPL technique reduces the retransmission packets, thereby increasing the PDR. In addition, WNC-RDPL allows mixing multi packets in to one coded packet, which leads to reduce the capacity requirement of multiple unicast sessions and increases the throughput. The performance of WNC-RDPL in multiple cases of link failures is investigated. The simulation results indicate that WNC-RDPL significantly increase the PDR and reduce the number of retransmissions due to packet loss, thus improving the transmission bandwidth efficiency.

4.2 Suggestion

In this dissertation, the error correction codes named low complexity parity check (LCPC). Wireless network erasure correction (WNC) code with a router detect packet loss (RDPL) algorithm, for multi radio multi-channel wireless mesh networks (MRMC WMN). Binary joint network with channel coding (B-JNCC) scheme for reliable multi-hop multi-radio multi-channel wireless mesh networks (MHMRMC-WMN) are have been successfully simulation based on MATLAB. However, this research can be significantly enhanced further through the following:

- In error correction codes, the project proposes to implement an adaptive error detection and correction codes base SNR. The receiver estimates the value of SNR and based in this it decided which correction algorithm used. From the SNR value, one can estimate the type of bit error that happened in received codeword. For example, if the SNR $\geq$ 5 dB, in this case the type of error in received codeword is one and double bit error. Whereas, if the SNR = 0 dB, in this case the type of error in a received codeword is one, two, three and four bit error when the length of received codeword is 9 bits.

- Find another solution for the problem of two error patterns for one syndrome vector. This project already solved the problem by send CD_{T3}.
- In addition to the link failure problem, the project takes in consider the lost packets in wireless network due to the packets collision and hidden node problem.
- Study the effects of the proactive routing protocol (OLSR) and the dynamic channel assignment in the performance of proposed WNC-RDPL.

5 ACKNOWLEDGEMENT

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Analysis of the Effect of Binary OR Operation Property in Linear Network Coding Transmission Performance

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KEYWORDS
Linear network coding; Multi hop wireless network; Network coding; Wireless network coding

ABSTRACT
In all of the works related to linear network coding (LNC), researchers use the Galois field (GF) operation property to generate the coding matrix elements. The elements of the coding matrix produced from the coefficients are randomly generated and selected from the GF(q). In this paper, we present an analytical analysis on the effect of using the binary field (OR operation) property to generate the coding matrix elements in LNC instead of using the customary GF. The LNC with binary OR operation (LNC_OR) is simulated under various multihop wireless network conditions of coefficients and coding router for three different network topologies. Each of the three network topologies has different coefficients since different network coding router is used. Performances of the LNC with the different probability of the coefficients of coding matrix base on packet delivery ratio (PDR) are analysed. Simulation results for three different network topologies confirm that LNC_OR reduces the number of lost packets received and improves PDR performance as compared to LNC with GF (LNC_GF) operation property.

1. INTRODUCTION

Network coding is a new approach that aims to increase transmission capacity and improves throughput of a network. The basic concept of the network coding has been introduced in [1]. Unlike the traditional store and forward method, this new technique prevents the network from traffic congestion. Network coding technique increases the throughput, robustness of the transmission and packet delivery ratio (PDR) that define as the ratio of the total number of packet received correctly to the total number of packets transmitted from the source.

There are two main research domains in network coding. First domain is the inter-session network coding method where the technique combines packets from different sessions or flows in the coding node (relay node) such as COPE [2]. The second type is the intra-session network coding method, where the technique encodes data packets strictly belonging to the same session or flow such as MORE [3]. In this paper, the intra-session network coding method is used. The concept of linear network coding (LNC) has been introduced in [4] where an intermediate router combines data from different input links prior sending the combined data to the next node. The LNC technique improves the overall network's performance.

The authors in [4] and [5] propose their LNC where the data (symbols) transferred are viewed based on Galois field(q) [GF(q)], where q denotes the size of the finite field. The arithmetic operations of those symbols follow the GF(q) property for generating the coding matrix elements [4].

The GF(q)-based coding matrix coefficients are randomly generated for each network topology. These coefficients represent the data due to the existence of the network coding routers. Each network coding router produces two randomly GF(q) coefficients. Thus, the number of coefficients increases with the increase in the number of network coding router in a particular network topology. These coefficients are grouped together which refer to as the coding vector. These coding matrix coefficients are used to encode the native packets at the source node and to retrieve native packets at the receiver nodes.

This paper analytically analyses LNC that utilizes the binary OR operation instead of GF(q) to generate the coding matrix elements. We investigate the PDR performance of the LNC_OR for three different network topologies. Extensive analysis has been carried out to demonstrate the number of error packets received for all possible coding matrix configurations with one, two, three, and four coefficients equal zero. The transmission performance of LNC_GF and LNC_OR is obtained based on the PDR. In this analysis, it is assumed that all the nodes activities during the transmission of data packets from source to destination to happen within one session period. Simulation results are compared with the renown LNC works that utilizes GF(q) (LNC_GF) such as in [4, 6–8] which show improvement in terms of PDR almost 20%, 4%, and 15% for one, two, and three coefficients equal zero, respectively.

The rest of the paper is organized as follows. Section 2 presents a related work. Background of network coding is shown in Section 3. Analysis of LNC with binary OR operation is shown in Section 4. Simulation results are presented in Section 5. Section 6 concludes the paper.

2. RELATED WORK

Network coding is first introduced by Ahlswede et al. [1] which shows by having a router to encode several received packets achieve multicast capacity. Since then, there have been many works produced such as in [2], [8], [9], [10]. The authors in [4] and [5] use LNC based on GF(q), where q denotes the size of a finite field. Most network coding applications [11] are developed by the concept of LNC. The non-linear network coding on GF(q) also has been studied where the works are readily available in [12–16].

The survey of network coding and its applications is found in [7] where the authors present the details of LNC. Sanna and Izquierdo [17] present a survey of LNC and network error correction code constructions and algorithms. Li et al. [18] review the theory and algorithms of LNC, which carries applications to many established fields, including coding theory, computer science, distributed data storage, information security, information theory, network tomography, optical networks, optimization theory, peer-to-peer (P2P) content delivery, switching theory, and wireless/satellite communications.

The authors in [19] investigate the throughput and decoding-delay performance of random LNC as a function of the coding window size and the network size. In [20], the segment linear network coding scheme has been proposed. In that work, the decoding complexity dramatically reduced by adding some constraints to the encoding coefficients. The work in [21] presents analytical results of decoding probability for random LNC in acyclic networks.

In [22], the authors study the linear network error correction code on a multicast network. They show that the linear network error correction codes have different error correcting capacity at different destination.

One property of GF operation is $1+1 = 0$. This leads to create a problem when generating the coding matrix elements from the coefficients of coding that are randomly generated and selected from the GF. For example, if there are two coefficients with the value 1, this leads to make at least one element of the coding matrix equal to zero, which means the packet multiplied by this coding matrix element, when encoding the native packet, will be lost. This is similar to the link failure problem in the network. The worst case occurs if there are other elements of coding matrix equal to zero, in which case it is impossible to retrieve the packets transmitted. The aim of using the binary OR operation in this research (instead of GF) is to prevent the zero element produced from the summation of two coefficients having value 1.

3. BACKGROUND OF NETWORK CODING

A typical network coding can be represented as the butterfly topology shown in Figure 1 [4, 23]. This topology consists of two source routers, S_1 and S_2, two relay routers, R_1 and R_2, and two destination routers, D_1 and D_2. Each source S_1 and S_2 sends packets P_1 and P_2 to R_1. In addition, these sources send both packets at the same time slot to D_1 and D_2.

Assuming that each link can transmit only one packet during each slot time, using the traditional routing, the central link (between R_1 and R_2) will be the bottleneck of the transmission. This is because the central link R_1 and R_2 is only able to carry either P_1 or P_2 but not both during the same time slot. If P_1 is send through the central link, then D_1 will receive P_1 twice and there is no P_2 at

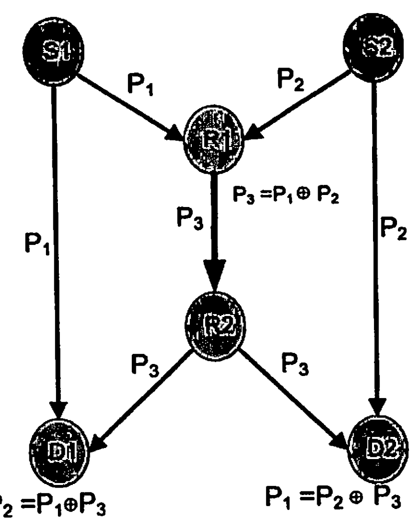


Figure 1. Butterfly network, a simple scenario showing how network coding improves throughput

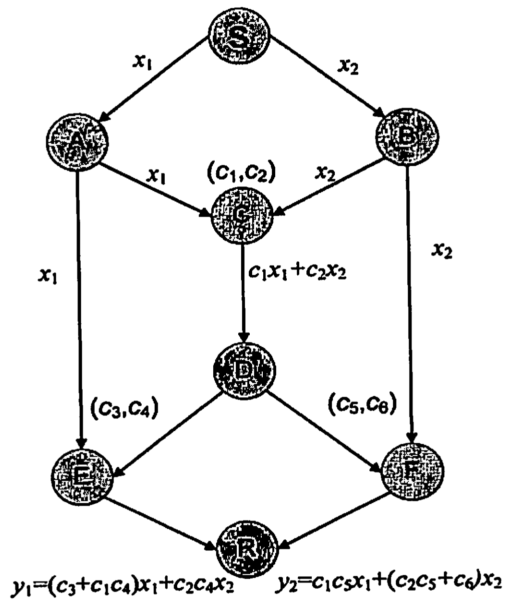


Figure 2. First network topology scenario of linear network coding [7].

all. The same issue is true, when sending P_2 to D_2 via the central link.

Network coding solves the above problem. In this technique, both D_1 and D_2 can receive packets P_1 and P_2 simultaneously. The R_1 encodes P_1 and P_2 into P_3 which is sent through the central link to R_2. P_1 is recovered at D_2 by taking the binary exclusive OR ($\oplus$) of P_2 and P_3. Thus, the network coding method allows to obtain multicast throughput of two packets per time slot, which is better than the traditional routing approach which can at best achieve 1.5 packets per slot time.

In this paper, we proposed used three network topologies to investigate the performance of the proposed method of LNC_OR on different number of coding matrix coefficients. Figure 2 shows the first network topology scenario for a typical LNC in order to use for the analysis. This figure shows the example of one source one destination (1S_1D) of the network scenario.

4. ANALYSIS OF LNC

In this paper, we have chosen the network topology with one source one destination (1S_1D) scenario shown in Figure 2 for the LNC analysis. Routers S and R correspond to the source and receiver, respectively. S generates N native packets and sends these packets to the receiver router R through a set of intermediate routers (A, B, C, D, E, and F).

In LNC, the variables, X, Y, and C are the native packets, the received coded packets at receiver R, and the coding matrix, respectively [7]. They are expressed as follows:

$$X=\begin{pmatrix} x_1 \\ x_2 \\ \vdots \\ x_N \end{pmatrix},\quad Y=\begin{pmatrix} y_1 \\ y_2 \\ \vdots \\ y_{M^{[r]}} \end{pmatrix},\quad C=\begin{pmatrix} c_1 \\ c_2 \\ \vdots \\ c_{M^{[r]}} \end{pmatrix},$$

where, $x_1 \ldots x_N$, are the components of a particular native packet, $y_1 \ldots\ldots y_{M^{[r]}}$ are the components of the coded packet and $c_1 \ldots . c_{M^{[r]}}$ are the coefficients of the coding matrix as indicated in Figure 2.

It then follows that the received coded packets Y can be expressed as $Y = CX$, and thus

$$\begin{pmatrix} y_1 \\ y_2 \\ \vdots \\ y_{M^{[r]}} \end{pmatrix} = \begin{pmatrix} c_{1,1} & c_{1,2} & \cdots & c_{1,N} \\ c_{2,1} & c_{2,2} & \cdots & c_{2,N} \\ \vdots & \vdots & \ddots & \vdots \\ c_{M^{[r]},1} & c_{M^{[r]},2} & \cdots & c_{M^{[r]},N} \end{pmatrix} \begin{pmatrix} x_1 \\ x_2 \\ \vdots \\ x_N \end{pmatrix}. \tag{1}$$

Note that the native packets elements of X can only be retrieved from Y when $M^{[r]} \geq N$ and rank $C = N$ [7].

As shown in Figure 2, each of the intermediate routers (coding routers) C, E, and F (pink colour) has two input links and one output link. The input link consists of two coefficients. For example, (c_1, c_2) are the related input link coding coefficients at router C. The (c_3, c_4) denotes the coefficients for the input link coding at router E. The (c_5, c_6) denotes the coefficients for the input link coding at router F. Thus, these coefficients (c_1, c_2, c_3, c_4, c_5, c_6) are the component of a coding matrix C for a particular LNC. Each coefficient will have the value of either 0 or 1 generated randomly based on GF(2).

In order to maintain the correctness of our proposed technique, we assume that the coefficients in the coding matrix will never be all equal zeroes or ones for a particular vector. A mechanism is proposed whereby the two randomly coefficients produced by each network coding router must be either (0, 1) or (1, 0). Therefore, the (0, 0) or (1, 1) coefficients configurations are avoided. Figure 3 shows a practical checking mechanism within a particular network coding router. If the coefficients happen to be (0, 0) or (1, 1), the scheme regenerates new set of coefficients, and this process continues until the condition is satisfied.

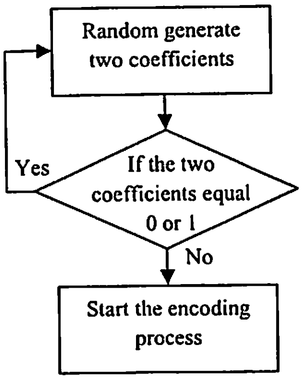


Figure 3. Flowchart for prevents all the coefficients of coding matrix equal to zero or one in network coding router.

The intermediate routers A, B, and D in Figure 2 (green colour) dedicate themselves to forward the received packets without any encoding process.

As shown in Figure 2, the source S transmits its native packets components x_1 and x_2 to the intermediate routers A and B. Each of these routers produces two output links. Then, the packets are transmitting to receiver router R via the coding routers (C, E, and F). The receiver router R receives two coded packets components y_1 and y_2, which are expressed as [7]

$$y_1 = (c_3 + c_1c_4)\ x_1 + c_2c_4x_2 \tag{2}$$

$$y_2 = c_1\ c_5x_1 + (c_2c_5 + c_6)x_2. \tag{3}$$

Since the received packet is expressed as $Y = CX$, the linear equations for LNC is given by

$$\begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = \begin{pmatrix} c_3 + c_1c_4 & c_2c_4 \\ c_1c_5 & c_2c_5 + c_6 \end{pmatrix} \begin{pmatrix} x_1 \\ x_2 \end{pmatrix} \tag{4}$$

In order to retrieve the native packet components, a receiver router needs to know the variable coding matrix of the encoded packets that they received [7].

If no packet is lost at receiver, from Equation (4), the coding matrix C is given as

$$C = \begin{pmatrix} c_3 + c_1c_4 & c_2c_4 \\ c_1c_5 & c_2c_5 + c_6 \end{pmatrix} \tag{5}$$

or it can be further simplified as

$$C = \begin{pmatrix} c_{1,1} & c_{1,2} \\ c_{2,1} & c_{2,2} \end{pmatrix}, \tag{6}$$

where

$c_{1,1} = c_3 + c_1c_4$, $c_{1,2} = c_2c_4$, $c_{2,1} = c_1c_5$, $c_{2,2} = c_2c_5 + c_6$.

4.1 LNC with one-coefficient equal zero

GF(2) property: suppose that $q = 2$, $c_1 = 0$, and c_2, c_3, c_4, c_5, and c_6 are equal to one. The following steps show the analytical analysis of the LNC. From Equation (5), the coding matrix C for the LNC_GF is given as

$$C = \begin{pmatrix} c_3 + 0 & c_2c_4 \\ 0 & c_2c_5 + c_6 \end{pmatrix}. \tag{7}$$

In this case, since the received packet is expressed as $Y = CX$ and from Equation (7), the received coded packets y_1 can be expressed as

$$y_1 = c_3\ x_1 + c_2c_4x_2 \tag{8}$$

and

$$y_2 = 0.x_1 + (c_2c_5 + c_6)\ x_2 = (1.1 + 1)\ \ x_2 = \ (1 + 1)x_2.$$

Since in the GF (2) property, $(1 + 1) = 0$, thus

$$y_2 = (0).\ \ x_2 = \ 0.$$

From Equation (8) and $c_2 = c_3 = c_4 = 1$,

$$y_1 = \ 1.x_1 + 1.1.x_2 = x_1\ +\ x_2 = x_3. \tag{9}$$

The variable x_3 is the GF(2) additional operation of $(x_1 + x_2)$, which is equivalent to binary exclusive OR $(x_1 \oplus x_2)$.

As shown above that $y_2 = 0$ and $y_1 = x_3$. Therefore, x_1 and x_2 cannot be recovered from y_1 and y_2 when the LNC's coding matrix components C are viewed based on GF(2).

Binary OR property: suppose a particular LNC uses the same setup as above, when the coding matrix C components now are viewed based on the binary OR property. The following steps show the analytical analysis.

From Equation (2) and $c_1 = 0$, $c_2 = c_3 = c_4 = 1$,

$$y_1 = (c_3 + c_1c_4)\ x_1 + c_2c_4\ x_2 = (1 + 0)\ x_1 + (1.1)x_2.$$

Therefore,

$$y_1 = \ 1x_1 + 1x_2 = x_1\ +\ x_2\ =\ x_3. \tag{10}$$

From Equation (3),

$$y_2 = c_1c_5x_1 + (c_2c_5 + c_6)\ x_2 = \ 0 + (c_2c_5 + c_6)x_2 \\ = (1.1 + 1)\ x_2. \tag{11}$$

Since in the binary OR property, (1 + 1) = 1, thus Equation (11) becomes

$$y_2 = (1 + 1)\ \ x_2 = x_2. \tag{12}$$

Consequently, from the above analytical analysis, x_2 can be obtained from y_2 as shown in Equation (12).

As shown in Figure 2 at the destination node, the component x_1 is obtained by summing operation between two receiving terms (i.e., Equation (10) and Equation (12)).

$$y_1 + \ y_2 \ = \ (x_1 \ + \ x_2) + \ x_2 = x_1. \tag{13}$$

Thus, after this analysis, the component x_1 and x_2 can be recovered.

4.2 LNC with two coefficients equal zero

GF(2) property: suppose c1 = 0 and c4 = 0 and c2 = c3 = c5 = c6 = 1. The following steps show the analytical analysis of LNC viewing the coding matrix C coefficients in GF(2) reveals the following structure:

$$C = \begin{pmatrix} c_3 & 0 \\ 0 & c_2c_5 + c_6 \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ 0 & 0 \end{pmatrix}. \tag{14}$$

From Equation (2) and $c_1 = c_4 = 0$ and $c_2 = c_3 = c_5 = c_6 = 1$,

$$y_1 = (c_3 + c_1c_4)\ x_1 + c_2c_4\ x_2 = (1 + 0)\ x_1 + 0.x_2.$$

Since (1+ 0 = 1) in GF(2), thus

$$y_1 = x_1.$$

From Equation (3),

$$y_2 = c_1\ c_5x_1 + (c_2c_5 + c_6)\ x_2 = \ 0.x_1 + (1 + 1)\ x_2 = \ 0.$$

Since (1+ 1 = 0) in GF(2), thus

$$y_2 = 0.$$

Thus, $y_1 = x_1$ and $y_2 = 0$. The native packet element x_2 cannot be retrieved since the rank $C < 2$ [7].

Binary OR property: suppose a particular LNC uses the same setup as above, when the coding matrix C components now are viewed based on the binary OR property. The following steps show the analytical analysis.

From Equation (14), and $c_1 = c_4 = 0$ and $c_2 = c_3 = c_5 = c_6 = 1$. The right column and bottom row element of C (and after using binary OR property),

$$c_2c_5 + \ c_6 \ = \ 1.1 + 1 \ = \ 1.$$

Thus, the coding matrix C has the following structure:

$$C = \begin{pmatrix} c_3 & 0 \\ 0 & c_2c_5 + c_6 \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}. \tag{15}$$

Using Equation (15) and from the received packets that are expressed as $Y = CX$, the native component can be recovered as the steps are shown below:

$$y_1 = (1 + 0)\ x_1 + 0x_2.$$

Since in binary OR, (1 + 0) = 1, thus,

$$y_1 = \ x_1.$$

Similarly from Equation (15),

$$y_2 = (1 + 1)\ \ x_2 = x_2.$$

Since in binary OR, (1 + 1) = 1, thus,

$$y_2 = \ x_2.$$

In this case, there is no problem to retrieve the native packet components x_1 and x_2 at the receiver R.

Thus, based on the above analytical analysis, the GF(2) property leads LNC to errors in recovering the received packets. The addition of two "1"s coefficients equal zero (1+1 = 0), which make the rank of $C < m$, where m is the number of rows in coding matrix C. This suggests that viewing the coding matrix C coefficients as the binary OR property in the analysis solves the problem.

5. RESULTS

In addition to the network topology shown in Figure 2, this work presents the results for another two different network topologies, one with 4 coefficients coding matrix and another with 10 coefficients coding matrix which will be shown later in Figures 6 and 8, respectively. The analysis for these network topologies is the same as the analysis for the network topology shown in Figure 2, thus, we demonstrate only the tables and figures that show the performance of that network topologies due to a limited number of pages.

5.1 Analytical analysis

First, we use the network as shown in Figure 2 for this analysis. The network produces the coding matrix coefficients C (c_1, c_2, c_3, c_4, c_5, and c_6). In particular, we analyse the LNC based on four different cases, i.e., 1, 2, 3, and 4 coefficients of coding matrix. Each of the coefficients has a random value (0, 1). As shown in Figure 2, if $c_1 = 0$, this indicates that node C does not receive packet x_1 from node A via a link (A, C) (i.e., packet x_1 is lost).

5.1.1 Case 1: One coefficient is zero

Table 1 shows that the results obtained for LNC with one coefficient is zero. The coding matrix coefficients column (2nd column) in Table 1 shows all the possible combinations with one coefficient is zero that can happen. There are six coefficients of the coding matrix (c_1, c_2, c_3, c_4, c_5, c_6), which produces six possible of one link failures. Table 1 also shows the corresponding number of received packets with errors when the LNC_GF (4th column) and LNC_OR (6th column) are analysed.

Assuming that S sends packet x_1 to the intermediate router A and packet x_2 to the intermediate router B, then both routers forward x_1 and x_2 packets to the next router (next hop), i.e., routers C, E, and F, respectively. As shown in Figure 2, routers C, E, and F have two input links and one output link. Hence, these routers are considered as the coding router (coding nodes). Each coding router has two variable coefficients on the input links.

The total number of errors (which is obtained by summing all the values in the 4th column) is 6 for the LNC_GF and the total number of errors (summing all the values of the 6th column) is 4 for the LNC_OR as shown in Table 1.

Another important quantity is the total number of possible configuration with errors (the number of shaded cells on the 4th and 6th columns). In case LNC_GF, the total number of possible configurations with errors is 4 (the total number of shaded cell on the 4th columns) in Table 1. Whereas, in LNC_OR, the total number of possible configuration with errors is 2 (the total number of shaded cell on the 6th columns) in Table 1. This suggests that the LNC_OR is more powerful which reduces the total number of errors and reduces the total number of possible configuration with errors.

Take, for example, at index 2 in Table 1, to calculate the number of packets received with errors for LNC_GF and LNC_OR. $c_2 = 0$ and $c_1, c_3, c_4, c_5, c_6 = 1$ as shown in Table 1. The coded packets received in LNC_GF can be analysed as follows:

$$y_1 = (c_3 + c_1c_4)\ x_1 + c_2c_4\ x_2 = (1+1)\ x_1 + 0$$
$$= (0)\ x_1 + 0\ =\ 0$$
$$y_2 = c_1\ c_5x_1 + (c_2c_5 + c_6)\ x_2 = x_1\ + (0+1)x_2$$
$$y_2 = x_1 +\ x_2 = x_3.$$

Thus, the native packets x_1 and x_2 cannot be retrieved, because $y_1 = 0$ and $y_2 = x_3$. Therefore, the number of packet error received is 2 as indicated in Table 1.

In LNC_OR with the same condition $c_2 = 0$ and $c_1 = c_3 = c_4 = c_5 = c_6 = 1$, the coded packets received can be analysed as follows. From Equations (2) and (3),

$$y_1 = (c_3 + c_1c_4)\ x_1 + c_2c_4\ x_2 = (1+1)\ x_1 + 0$$
$$= (1)\ x_1 + 0\ = x_1$$
$$y_2 = c_1\ c_5x_1 + (c_2c_5 + c_6)x_2$$
$$y_2 = x_1\ + (0+1)\ x_2 = x_1\ +\ x_2 = x_3.$$

Table 1. Number of error packets received for LNC_GF and LNC_OR in case one coefficient equal zero for the first network topology scenario

	Coding matrix coefficients						Packets received of LNC_GF			Packets received of LNC_OR		
Index	c_1	c_2	c_3	c_4	c_5	c_6	y_1	y_2	No. of errors	y_1	y_2	No. of errors
1	0	1	1	1	1	1	x_3	0	2	x_3	x_2	0
2	1	0	1	1	1	1	0	x_3	2	x_1	x_3	0
3	1	1	0	1	1	1	x_3	x_1	0	x_3	x_3	2
4	1	1	1	0	1	1	x_1	x_1	1	x_1	x_3	0
5	1	1	1	1	0	1	x_2	x_2	1	x_3	x_2	0
6	1	1	1	1	1	0	x_2	x_3	0	x_3	x_3	2
							Total number of errors		6			4
							Total number of possible configurations with errors		4			2

Thus, the native packet x_1 can be retrieved from y_1, and the native packet x_2 can be retrieved from the summation between y_1 and y_2. As mentioned (summation operation in GF is the same as XOR operation in binary field). Therefore,

$$y_1 + y_2 = x_1 + x_3 = x_1 + x_1 + x_2 = x_2.$$

The number of error packets received in this case is 0 for the LNC_OR (index 2). Similarly, analysis can be applied for other indexes in Table 1.

5.1.2 Case 2: Two coefficients are zero

Table 2 shows that the results obtained for the network topology scenario with two coefficients are zero. There are 15 possible configurations of two coefficients equal zero for the LNC_GF and LNC_OR. The total number of errors (summing all the values of the 4th column) is 8 for both LNC_GF and LNC_OR (6th column), respectively. The total number of possible configurations with errors (number of shaded cells in 4th column) is 7 for LNC_GF. The total number of possible configurations with errors (number of shaded cells in 6th column) is 5 for LNC_OR. Because the total number of possible configurations with errors for LNC_OR is less than the total number of possible configurations with errors for LNC_GF, the LNC_OR provides better PDR performance as compared to the LNC_GF as shown in Table 2.

Take, for example, at index = 3 in Table 2 to calculate the number of packets received with errors for LNC_GF and LNC_OR. $c_1 = c_4 = 0$, and $c_2 = c_3 = c_5 = c_6 = 1$ as shown in Table 2. The R node receives only x_1 in LNC_GF as shown by the following equations.

From Equations (2) and (3),

$$y_1 = (c_3 + c_1c_4)\ x_1 + c_2c_4\ x_2 = (c_3 + 0)\ x_1 + 0 = x_1$$

$$\begin{aligned} y_2 &= c_1\ c_5x_1 + (c_2c_5 + c_6)\ x_2 = 0 + (c_2c_5 + c_6)x_2 \\ &= 0 + (1+1)\ x_2 = 0 + (0)\ x_2 = 0. \end{aligned}$$

Thus, LNC_GF cannot retrieve the native packet x_2, because $y_2 = 0$ and $y_1 = x_1$. The number of receiving packets in errors is 1 as shown in Table 2 (index = 3).

On the other hand, when OR used and under the same conditions when c_1 and $c_4 = 0$ and c_2, c_3, c_5, and $c_6 = 1$, as shown in Table 2 (at index 3), in this case when OR used the coded packets received defined as

$$\begin{aligned} y_1 &= (c_3 + c_1c_4)\ x_1 + c_2c_4\ x_2 = (1+0)\ x_1 + 0 \\ &= (1)\ x_1 + 0 = x_1 \end{aligned}$$

$$\begin{aligned} y_2 &= c_1\ c_5x_1 + (c_2c_5 + c_6)\ x_2 = (0)\ x_1 + (1+1)\ x_2 \\ &= 0 + x_2 = x_2. \end{aligned}$$

LNC_OR with the same condition can retrieve the native packet x_1 from y_1, and the native packet x_2 from y_2. Thus, the number of received packets with errors is zero. Similarly, the analysis can be applied for other indexes in Table 2.

5.1.3 Case 3: Three coefficients are zero

Table 3 shows that the results obtained for the network topology with three coefficients are zero scenarios. There

Table 2. Number of error packets received for LNC_GF and LNC_OR in case two coefficients equal zero for the first network topology scenario

	Coding matrix coefficients						Packets received of LNC_GF			Packets received of LNC_OR		
Index	c_1	c_2	c_3	c_4	c_5	c_6	y_1	y_2	No. of errors	y_1	y_2	No. of errors
1	0	0	1	1	1	1	x_1	x_2	0	x_1	x_2	0
2	0	1	0	1	1	1	x_2	0	1	x_2	x_2	1
3	0	1	1	0	1	1	x_1	0	1	x_1	x_2	0
4	0	1	1	1	0	1	x_3	x_2	0	x_3	x_2	0
5	0	1	1	1	1	0	x_3	x_2	0	x_3	x_2	0
6	1	0	0	1	1	1	x_1	x_3	0	x_1	x_3	0
7	1	0	1	0	1	1	x_1	x_3	0	x_1	x_3	0
8	1	0	1	1	0	1	0	x_2	1	x_1	x_2	0
9	1	0	1	1	1	0	0	x_1	1	x_1	x_1	1
10	1	1	0	0	1	1	0	x_1	1	0	x_3	2
11	1	1	0	1	0	1	x_3	x_2	0	x_3	x_2	0
12	1	1	0	1	1	0	x_3	x_3	2	x_3	x_3	2
13	1	1	1	0	0	1	x_1	x_2	0	x_1	x_2	0
14	1	1	1	0	1	0	x_1	x_3	0	x_1	x_3	0
15	1	1	1	1	0	0	x_2	0	1	x_3	0	2
							Total number of errors		8			8
							Total number of possible configurations with errors		7			5

Table 3. Number of error packets received for LNC_GF and LNC_OR in case three coefficients equal zero for the first network topology scenario

	Coding matrix coefficients						Packets received of LNC_GF			Packets received of LNC_OR		
Index	c_1	c_2	c_3	c_4	c_5	c_6	y_1	y_2	No. of errors	y_1	y_2	No. of errors
1	0	0	0	1	1	1	0	x_2	1	0	x_2	1
2	0	0	1	0	1	1	x_1	x_2	0	x_1	x_2	0
3	0	0	1	1	0	1	x_1	x_2	0	x_1	x_2	0
4	0	0	1	1	1	0	x_1	0	1	x_1	0	1
5	0	1	0	0	1	1	0	0	2	0	x_2	1
6	0	1	0	1	0	1	x_2	x_2	1	x_2	x_2	1
7	0	1	0	1	1	0	x_2	x_2	1	x_2	x_2	1
8	0	1	1	0	0	1	x_1	x_2	0	x_1	x_2	0
9	0	1	1	0	1	0	x_1	x_2	0	x_1	x_2	0
10	0	1	1	1	0	0	x_3	0	2	x_3	0	2
11	1	0	0	0	1	1	0	x_3	2	0	x_3	2
12	1	0	0	1	0	1	x_1	x_2	0	x_1	x_2	0
13	1	0	0	1	1	0	x_1	x_1	1	x_1	x_1	1
14	1	0	1	0	0	1	x_1	x_2	0	x_1	x_2	0
15	1	0	1	0	1	0	x_1	x_1	1	x_1	x_1	1
16	1	0	1	1	0	0	0	0	2	x_1	0	1
17	1	1	0	0	0	1	0	x_2	1	0	x_2	1
18	1	1	0	0	1	0	0	x_3	2	0	x_3	2
19	1	1	0	1	0	0	x_3	0	2	x_3	0	2
20	1	1	1	0	0	0	x_1	0	1	x_1	0	1
							Total number of errors		20			18
							Total number of possible configurations with errors		14			14

are 20 possible configurations out of 3 coefficients zero for LNC_GF and LNC_OR. The total number of errors (summing all the values in the 4th column) is 20 for LNC_GF. The total number of errors (summing all the values in the 6th column) is 18 for LNC_OR. The total number of possible configurations with errors (total number of shaded cells in 4th column or 6th column) in both LNC_GF and LNC_OR is 14.

In the LNC_GF, there are 6 indexes (5, 10, 11, 16, 18, and 19) as shown in the 4th column of Table 3 where each has 2 error packets received. Whereas, in the LNC_OR, there are 4 indexes (10, 11, 18, and 19) as shown in the 6th column of Table 3 where each has 2 error packets received. Therefore, the probability of having 2 error packets received in LNC_GF is more than the probability of having 2 error packets received in LNC_OR.

This explains the reason for there are approximately 15% difference between the simulation results of PDR for LNC_OR (65%) and LNC_GF (50%) as shown in Figure 4 when simulation program runs around 1000 iterations. But if we take all the 20 cases of probability of coding matrix coefficients for LNC_OR and LNC_GF only one time, as shown in Table 3, this leads to make the theoretical value of PDR for LNC_OR and LNC_GF equal to 55% and 50%, respectively.

In real implementation and application, these 20 cases of probability of coding matrix coefficients shown in Table 3 happened randomly and more than one time and it is not necessary all the indexes apply when the simulation program run one iteration or a few numbers of iterations. Therefore, to obtain the average result (accurate result) and to be sure, all the index configurations take into consider a simulation program runs around 1000 iterations. In that case, the simulation result shows the PDR for LNC_OR in case three coefficients of coding matrix equal zero are about 65% and the PDR for LNC_GF is 50%.

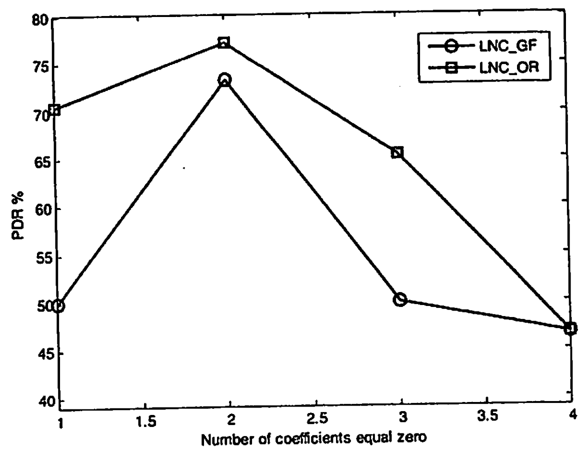


Figure 4. PDR versus number of coefficients equal zero for LNC_GF and LNC_OR when 1000 packets sent and simulation program run1000 repetition.

Table 4. Number of error packets received for LNC_GF and LNC_OR in case four coefficients equal zero for the first network topology scenario

	Coding matrix coefficients						Packets received of LNC_GF			Packets received of LNC_OR		
Index	c_1	c_2	c_3	c_4	c_5	c_6	y_1	y_2	No. of errors	y_1	y_2	No. of errors
1	0	0	0	0	1	1	0	x_2	1	0	x_2	1
2	0	0	0	1	0	1	0	x_2	1	0	x_2	1
3	0	0	0	1	1	0	0	0	2	0	0	2
4	0	0	1	0	0	1	x_1	x_2	0	x_1	x_2	0
5	0	0	1	0	1	0	x_1	0	1	x_1	0	1
6	0	0	1	1	0	0	x_1	0	1	x_1	0	1
7	0	1	0	0	0	1	0	x_2	1	0	x_2	1
8	0	1	0	0	1	0	0	x_2	1	0	x_2	1
9	0	1	0	1	0	0	x_2	0	1	x_2	0	1
10	0	1	1	0	0	0	x_1	0	1	x_1	0	1
11	1	0	0	0	0	1	0	x_2	1	0	x_2	1
12	1	0	0	0	1	0	0	x_1	1	0	x_1	1
13	1	0	0	1	0	0	x_1	0	1	x_1	0	1
14	1	0	1	0	0	0	x_1	0	1	x_1	0	1
15	1	1	0	0	0	0	0	0	2	0	0	2
							Total number of errors		16			16
							Total number of possible configurations with errors		14			14

Thus, as for overall, Figure 4 shows that LNC_OR provides the PDR gain over LNC_GF around 20% if analysing for one coefficient is equal zero, 4% if analysing for two coefficients are equal zero, and 15% if analysing for three coefficients are equal zero as compared for all cases to LNC_GF.

5.1.4 Case 4: Four coefficients are zero

The worst-case scenario happens when 4 coefficients are zero in the LNC. Table 4 illustrates the corresponding coding matrix coefficients configuration. The numbers of errors in the packet received are also shown in Table 4. The total number of received packets with errors (summing all the values of the 4th column) is 16 for both LNC_GF and LNC_OR. The total number of possible configurations with errors (number of shaded cells) is 14 for both LNCs.

5.2 PDR performance

Second, the performances of LNCs with the two properties of the coding matrix C coefficients are analysed based on the PDR (in percentage) for different cases of the coefficients coding matrix and evaluated with different number of native packets transmitted. In this work, it has been assumed that the nodes activities happen within one session period. Therefore, the performance throughput analysis for LNCs cannot be done. Instead, in this work, we choose to analyse the performance of LNCs based on a PDR.

In this paper, the PDR performance is investigated with three different network topologies scenarios. The first is one source router, one destination router with three network coding routers (i.e., six coefficients for coding matrix) as shown in Figure 2. The second is the two-source routers, one destination router with two network coding routers (i.e., four coefficients for coding matrix) as shown in Figure 6. The third is two-source routers, one destination router with five network coding routers (i.e., ten coefficients for coding matrix) as shown in Figure 8.

5.2.1 First network topology

Based on the network topology shown in Figure 2, the PDR performance of LNC_GF and LNC_OR is analysed with the assumption of having either 1, 2, 3, or 4 coefficients equal zero. In simulation, 1000 packets are sent for each iteration. In this work, to ensure all the index configurations possibilities for the coefficients of coding matrix take in consider, we use 1000 iterations in order to obtain accurate results. Simulation results show that the PDR performance of LNC_OR is better than LNC_GF.

Figure 4 shows the simulation results of PDR performances for LNC_GF and LNC_OR with 1000 packets transmitted for various coefficients coding matrix. The PDR gain is computed by taking the difference between LNC_OR's PDR and LNC_GF's PDR at one, two, and three coefficients equal zero. For example, in Figure 4, for the case of one coefficient equal zero, PDR is 70.6% for LNC_OR whereas it is 50.0% for LNC_GF.

As shown in Figure 4, for the case of two coefficients equal zero, the PDR is 77.2% for LNC_OR and 73.3% for LNC_GF. In the case of three coefficients equal to zero,

Table 5. PDR for LNC_GF and LNC_OR in case 1000 packets send and the simulation program run 1000 iteration for the first network topology scenario

	PDR % for one coefficient is zero	PDR % for two coefficients are zero	PDR % for three coefficients are zero	PDR % for four coefficients are zero	Average of PDR %
LNC_GF	50.031	73.301	50.015	46.679	55.0065
LNC_OR	70.577	77.234	65.392	46.679	64.9705

the PDR is 65.4% for LNC_OR, and 50.0% for LNC_GF. Thus, the difference is around 15%, as shown in Table 5.

Figure 5 illustrates the PDR performances versus the number of packets transmitted using LNC_GF and LNC_OR when the value of four coefficients matrix coding randomly change. Figure 5 clearly shows the PDR performance of LNC_OR is better than LNC_GF. The maximum PDR obtained in LNC_OR is 65% whereas the maximum PDR obtained in LNC_GF is 55% at various number of packets transmitted. The PDR gain here is 10%.

The performance of PDR depends on the network topology and the number of network coding routers (the number of coefficients in the coding matrix). In addition, the performance of PDR depends on the status of the coefficients coding matrix (i.e., how many coefficients equal zero).

5.2.2 Second network topology

The second network topology is shown in Figure 6 with two source routers S1 and S2, two relays (network coding routers) A and B, and one receiver router R1. In this

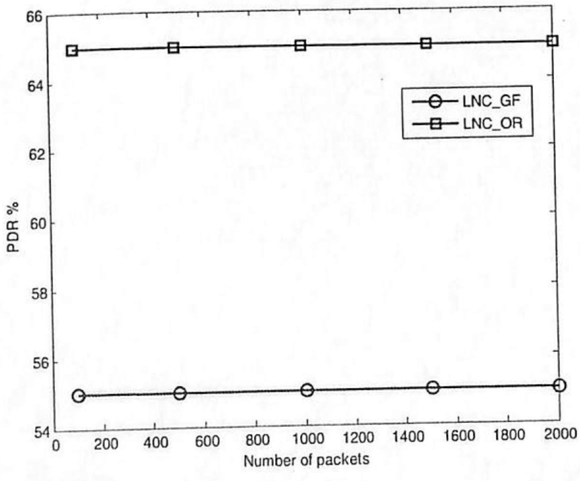


Figure 5. PDR versus number of packets sent for LNC_GF and LNC_OR when the number of coefficients equal zero is random.

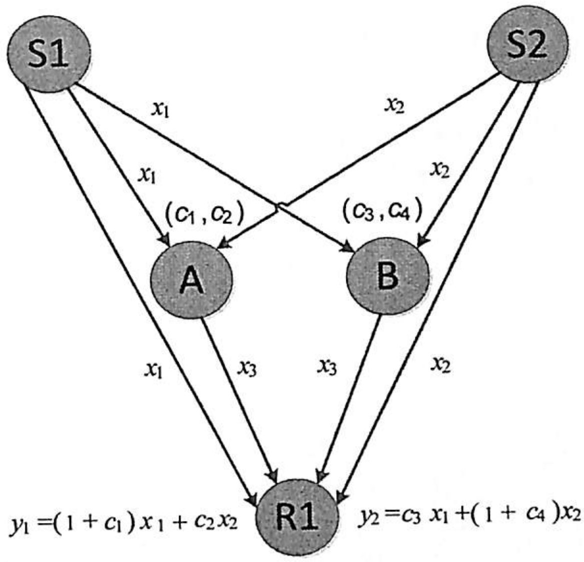


Figure 6. Second network topology for four coefficients of coding matrix.

scenario, S1 and S2 broadcast packets x_1 and x_2 to the relay routers (coding routers) A and B. R1 receives x_1 and x_2 from S1 and S2, respectively, and x_3 (coded packet) from the relay routers A and B. The receiver R1 receives y_1 and y_2, which are shown below

$$y_1 = (1 + c_1)\ x_1 + c_2 x_2 \tag{16}$$

$$y_2 = c_3\ x_1 + (1 + c_4) x_2. \tag{17}$$

Since the received packet is expressed as $Y = CX$, the linear equations of LNC, using the network topology as shown in Figure 6, is expressed as follows:

$$\begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = \begin{pmatrix} 1+c_1 & c_2 \\ c_3 & 1+c_4 \end{pmatrix} \begin{pmatrix} x_1 \\ x_2 \end{pmatrix}. \tag{18}$$

The coding matrix is given as

$$C = \begin{pmatrix} 1+c_1 & c_2 \\ c_3 & 1+c_4 \end{pmatrix}. \tag{19}$$

Table 6 shows the number of error packets received for LNC_GF and LNC_OR in case one coefficient is zero. The number of error packets received for LNC_GF is two less than LNC_OR. Table 7 shows the number of error packets received (2nd from bottom row) for LNC_GF is eight more than LNC_OR. However, the total number of possible configurations with errors (bottom row) shows that LNC_OR (one error) is lower than LNC_GF (six errors).

Table 6. Number of error packets received for LNC_GF and LNC_OR in case one coefficient equal zero for the second network topology scenario

Index	Coding matrix coefficients				Packets received of LNC_GF		No. of errors	Packets received of LNC_OR		No. of errors
	c_1	c_2	c_3	c_4	y_1	y_2		y_1	y_2	
1	0	1	1	1	x_3	x_1	0	x_3	x_3	2
2	1	0	1	1	0	x_1	1	x_1	x_3	0
3	1	1	0	1	x_2	0	1	x_3	x_2	0
4	1	1	1	0	x_2	x_3	0	x_3	x_3	2
					Total number of errors		2			4
					Total number of possible configurations with errors		2			2

Table 7. Number of error packets received for LNC_GF and LNC_OR in case two coefficient equal zero for the second network topology scenario

Index	Coding matrix coefficients				Packets received of LNC_GF		No. of errors	Packets received of LNC_OR		No. of errors
	c_1	c_2	c_3	c_4	y_1	y_2		y_1	y_2	
1	0	0	1	1	x_1	x_1	1	x_1	x_3	0
2	0	1	0	1	x_3	0	2	x_3	x_2	0
3	0	1	1	0	x_3	x_3	2	x_3	x_3	2
4	1	0	0	1	0	0	2	x_1	x_2	0
5	1	0	1	0	0	x_3	2	x_1	x_3	0
6	1	1	0	0	x_2	x_2	1	x_3	x_2	0
					Total number of errors		10			2
					Total number of possible configurations with errors		6			1

The simulation program sends 1000 packets for each random coding matrix coefficients that runs 1000 repetitions. Figure 7 shows simulation results for the PDR performance of LNC_GF and LNC_OR. The PDR of the proposed LNC_OR method is better than LNC_GF in case of two coefficients equal zero. However, the LNC_GF is better than LNC_OR in case one coefficient equal zero.

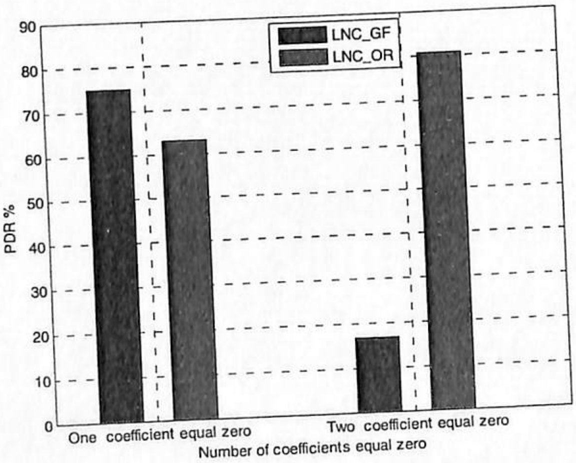


Figure 7. PDR versus number of coefficients equal zero for LNC_GF and LNC_OR when 1000 packets sent and simulation program run1000 repetition for network topology with two sources, two relays, and one receiver.

5.2.3 Third network topology

Figure 8 shows the third network topology implemented and analysed in this work. The topology consists of two source routers S1 and S2, five network-coding routers B, D, E, H, and I. The topology also contains routers A, C, F, and G as the intermediate routers and one receiver router R1. In this topology, S1 and S2 broadcast packets x_1 and x_2 to the routers A, B, and C. Router A forwards packet x_1 to coding routers D and H via links LAD and LAH, respectively. Router C forwards packet x_2 to coding router E and I via links LCE and LCI, respectively.

Network coding router B encodes packets x_1 and x_2, using the coefficients (c_1, c_2). Router B forwards the encoded packets to the network coding routers D and E. Thus, routers D and E have coefficients (c_3, c_4) and (c_5, c_6), respectively. Routers F and G forward the encoded packets that it received from routers D and E to the coding routers H and I via link LFH and link LGI. Thus, routers H and I have coefficients (c_7, c_8) and (c_9, c_{10}), respectively, which used to encode packets it received from routers A, F, G, and C. The receiver R1 receives y_1 and y_2 that are expressed as

$$y_1 = (c_1c_4c_8 + c_3c_8 + c_7)\ x_1 + c_2c_4c_8x_2 \tag{20}$$

$$y_2 = c_1\ c_5c_9x_1 + (c_2c_5c_9 + c_6c_9 + c_{10})x_2. \tag{21}$$

As mentioned, the received packets are expressed as $Y = CX$, therefore linear equations of LNC for network topology illustrated in Figure 8 is given by

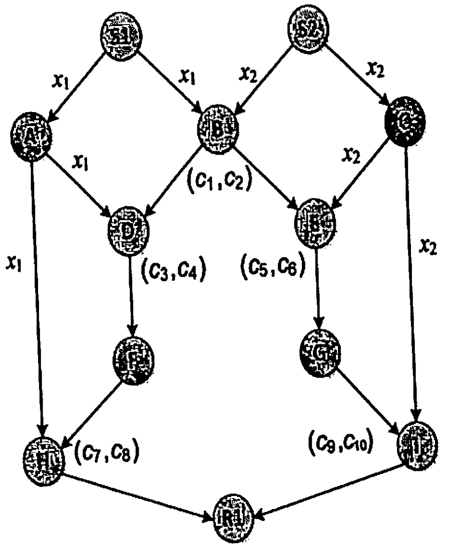
$$\begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = \begin{pmatrix} c_1c_4c_8 + c_3c_8 + c_7 & c_2c_4c_8 \\ c_1c_5c_9 & c_2c_5c_9 + c_6c_9 + c_{10} \end{pmatrix} \begin{pmatrix} x_1 \\ x_2 \end{pmatrix}.$$

The coding matrix with 10 coefficients is presented as

$$C = \begin{pmatrix} c_1c_4c_8 + c_3c_8 + c_7 & c_2c_4c_8 \\ c_1c_5c_9 & c_2c_5c_9 + c_6c_9 + c_{10} \end{pmatrix} \quad (22)$$

Similar analysis as the analysis for the first network topology with six coefficients of coding matrix shown in Figure 2 can be conducted. The network topology shown in Figure 8 consists of 10 coding matrix coefficients which is analysed to investigate the PDR performance of the system using LNC_GF and LNC_OR. Table 8 shows the possible configurations of the third network topology with one coefficient equal zero (Index 1–Index 10). The number of error packets received (2nd row from bottom) for LNC_GF (six errors) is less than LNC_OR (eight errors). However, the total number of possible configurations with errors (bottom row) shows that LNC_OR (four errors) matches with LNC_GF (four errors).

Table 9 shows the possible configurations of the third network topology with two coefficients equal zero (Index 1–Index 45). The number of error packets received (2nd row from bottom) for LNC_GF (32 errors) is more than LNC_OR (16 errors). The total number of possible configurations with errors (bottom row) shows that the LNC_OR (8 errors) is less than LNC_GF (26 errors). Thus, LNC_OR is better than LNC_GF.



$y_1 = (c_1c_4c_8 + c_3c_8 + c_7)x_1 + c_2c_4c_8x_2 \quad y_2 = c_1c_5c_9x_1 + (c_2c_5c_9 + c_6c_9 + c_{10})x_2$

Figure 8. Third network topology scenario for 10 coefficients of coding matrix.

The number of possible configurations with three, four, and five coefficients equal zero is 120, 210, and 252, respectively. Therefore, the analysis of those cases is omitted due to a limited number of pages.

The simulation program sends 1000 packets for each random coding matrix coefficients that runs 1000 repetitions. Figure 9 shows simulation results for the PDR performance of the system using LNC_GF and LNC_OR. The PDR of the proposed LNC_OR method is better than LNC_GF in case of two coefficients equal zero.

The PDR performances are further investigated on different numbers of coefficients coding matrix (from 1 to 9) equal zero as shown in Figure 9. The results

Table 8. Number of error packets received for LNC_GF and LNC_OR in case one coefficient equal zero for the third network topology scenario

	Coding matrix coefficients										Packets received of LNC_GF			Packets received of LNC_OR		
Index	c_1	c_2	c_3	c_4	c_5	c_6	c_7	c_8	c_9	c_{10}	y_1	y_2	No. of errors	y_1	y_2	No. of errors
1	0	1	1	1	1	1	1	1	1	1	x_2	x_2	1	x_3	x_2	0
2	1	0	1	1	1	1	1	1	1	1	x_1	x_1	1	x_1	x_3	0
3	1	1	0	1	1	1	1	1	1	1	x_2	x_3	0	x_3	x_3	2
4	1	1	1	0	1	1	1	1	1	1	0	x_3	2	x_1	x_3	0
5	1	1	1	1	0	1	1	1	1	1	x_3	0	2	x_3	x_2	0
6	1	1	1	1	1	0	1	1	1	1	x_3	x_1	0	x_3	x_3	2
7	1	1	1	1	1	1	0	1	1	1	x_2	x_3	0	x_3	x_3	2
8	1	1	1	1	1	1	1	0	1	1	x_1	x_3	0	x_1	x_3	0
9	1	1	1	1	1	1	1	1	0	1	x_3	x_2	0	x_3	x_2	0
10	1	1	1	1	1	1	1	1	1	0	x_3	x_1	0	x_3	x_3	2
											Total number of errors		6			8
											Total number of possible configurations with errors		4			4

Table 9. Number of error packets received for LNC_GF and LNC_OR in case two coefficient equal zero for the third network topology scenario

	Coding matrix coefficients										Packets received of LNC_GF			Packets received of LNC_OR		
Index	c_1	c_2	c_3	c_4	c_5	c_6	c_7	c_8	c_9	c_{10}	y_1	y_2	No. of errors	y_1	y_2	No. of errors
1	0	0	1	1	1	1	1	1	1	1	0	0	2	x_1	x_2	0
2	0	1	0	1	1	1	1	1	1	1	x_3	x_2	0	x_3	x_2	0
3	0	1	1	0	1	1	1	1	1	1	0	x_2	1	x_1	x_2	0
4	0	1	1	1	0	1	1	1	1	1	x_2	0	1	x_3	x_2	0
5	0	1	1	1	1	0	1	1	1	1	x_2	0	1	x_3	x_2	0
6	0	1	1	1	1	1	0	1	1	1	x_3	x_2	0	x_3	x_2	0
7	0	1	1	1	1	1	1	0	1	1	x_1	x_2	0	x_1	x_2	0
8	0	1	1	1	1	1	1	1	0	1	x_2	x_2	1	x_3	x_2	0
9	0	1	1	1	1	1	1	1	1	0	x_2	0	1	x_3	x_2	0
10	1	0	0	1	1	1	1	1	1	1	0	x_1	1	x_1	x_3	0
11	1	0	1	0	1	1	1	1	1	1	0	x_1	1	x_1	x_3	0
12	1	0	1	1	0	1	1	1	1	1	x_1	0	1	x_1	x_2	0
13	1	0	1	1	1	0	1	1	1	1	x_1	x_3	0	x_1	x_3	0
14	1	0	1	1	1	1	0	1	1	1	0	x_1	1	x_1	x_3	0
15	1	0	1	1	1	1	1	0	1	1	x_1	x_1	1	x_1	x_3	0
16	1	0	1	1	1	1	1	1	0	1	x_1	x_2	0	x_1	x_2	0
17	1	0	1	1	1	1	1	1	1	0	x_1	x_3	0	x_1	x_3	0
18	1	1	0	0	1	1	1	1	1	1	x_1	x_3	0	x_1	x_3	0
19	1	1	0	1	0	1	1	1	1	1	x_2	0	1	x_3	x_2	0
20	1	1	0	1	1	0	1	1	1	1	x_2	x_1	0	x_3	x_3	2
21	1	1	0	1	1	1	0	1	1	1	x_3	x_3	2	x_3	x_3	2
22	1	1	0	1	1	1	1	0	1	1	x_1	x_3	0	x_1	x_3	0
23	1	1	0	1	1	1	1	1	0	1	x_2	x_2	1	x_3	x_2	0
24	1	1	0	1	1	1	1	1	1	0	x_2	x_1	0	x_3	x_3	2
25	1	1	1	0	0	1	1	1	1	1	0	0	2	x_1	x_2	0
26	1	1	1	0	1	0	1	1	1	1	0	x_1	1	x_1	x_3	0
27	1	1	1	0	1	1	0	1	1	1	x_1	x_3	0	x_1	x_3	0
28	1	1	1	0	1	1	1	0	1	1	x_1	x_3	0	x_1	x_3	0
29	1	1	1	0	1	1	1	1	0	1	0	x_2	1	x_1	x_2	0
30	1	1	1	0	1	1	1	1	1	0	0	x_1	1	x_1	x_3	0
31	1	1	1	1	0	0	1	1	1	1	x_3	x_2	0	x_3	x_2	0
32	1	1	1	1	0	1	0	1	1	1	x_2	0	1	x_3	x_2	0
33	1	1	1	1	0	1	1	0	1	1	x_1	0	1	x_1	x_2	0
34	1	1	1	1	0	1	1	1	0	1	x_3	x_2	0	x_3	x_2	0
35	1	1	1	1	0	1	1	1	1	0	x_3	x_2	0	x_3	x_2	0
36	1	1	1	1	1	0	0	1	1	1	x_2	x_1	0	x_3	x_3	2
37	1	1	1	1	1	0	1	0	1	1	x_1	x_1	1	x_1	x_3	0
38	1	1	1	1	1	0	1	1	0	1	x_3	x_2	0	x_3	x_2	0
39	1	1	1	1	1	0	1	1	1	0	x_3	x_3	2	x_3	x_3	2
40	1	1	1	1	1	1	0	0	1	1	0	x_3	2	0	x_3	2
41	1	1	1	1	1	1	0	1	0	1	x_2	x_2	1	x_3	x_2	0
42	1	1	1	1	1	1	0	1	1	0	x_2	x_1	0	x_3	x_3	2
43	1	1	1	1	1	1	1	0	0	1	x_1	x_2	0	x_1	x_2	0
44	1	1	1	1	1	1	1	0	1	0	x_1	x_1	1	x_1	x_3	0
45	1	1	1	1	1	1	1	1	0	0	x_3	0	2	x_3	0	2
											Total number of errors		32			16
											Total number of possible configurations with errors		26			8

demonstrate that the performance of the LNC_OR proposed method is better than LNC_GF (traditional method) for the number of coefficient equal zero of (from 2 to 7). The highest value of PDR obtained using LNC_OR is 84% when the number of coefficient equal zero is three. However, it is 50% for LNC_GF in the same scenario. The PDR for one coefficient equal zero in case LNC_GF is more than by 2.5% for case LNC_OR. On the other hand, the PDR for LNC_OR and LNC_GF is the same in cases 8 and 9 coefficients equal zero.

6. CONCLUSION

The paper analyses the LNCs performances for the three different network topologies when using the binary (OR) property instead of the GF property in generating the coding matrix elements. The use of the GF property increases the number of received packets errors and reduces the probability of the packet recovery. On the other hand, the use of the binary OR property reduces the total number of received packets with errors and reduces the total number of possible configurations with errors in LNC.

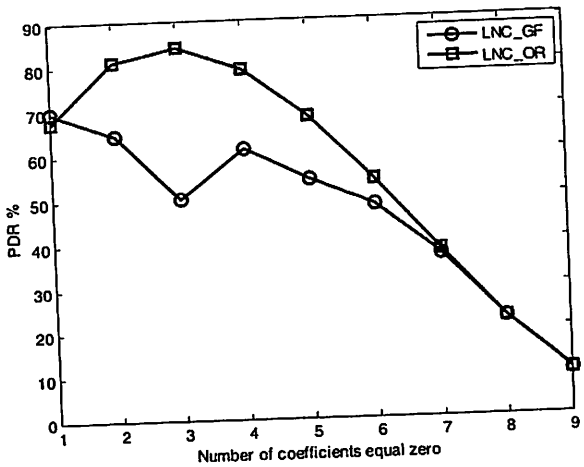


Figure 9. PDR versus number of coefficients equal zero for LNC_GF and LNC_OR when 1000 packets sent and simulation program run 1000 repetition for the third network topology scenario.

The PDR performance of LNC_OR and LNC_GF are analysed. Simulation results show for the first network scenario that LNC_OR's PDR gain is around 20% for one zero coefficient, 4% for two zero coefficient, and 15% for three zero coefficient as compared to the LNC_GF's PDR. The PDR gain which is around 10% in the case of the number of zero coefficient is random. Finally, as the conclusion, the LNC_OR proposed method provides effective and significant performances of PDR for a large number of coefficients.

DISCLOSURE STATEMENT

No potential conflict of interest was reported by the authors.

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A New PAPR Reduction Scheme: Wavelet Packet-based PTS with Embedded Side Information Data Scheme

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Reviewer: 1

Comments to the Author
1. The contribution from the paper is modest, so significant improvement of the quality is necessary. I did not see [illegible] comparison with existing techniques in the literature.

2. There are several unclear points in the development of the main results. The conditions imposed on the model/system looks strong, and remarks should be provided to show they are feasible and reasonable in practice.

3. Some details of the numerical examples are not fully specified. For instance, the selection of some parameters, and comparison of the results with expositing techniques.

4. The motivation of the practical use of the theoretic results obtained is weak.

5. English is poor. There are a number of grammar and composition mistakes, which make it difficult to read and follow the development in the paper in some stages. Also, some typos in the paper should be corrected as well.

Reviewer: 2

Comments to the Author
the idea is interesting but some problems are still to be solved:
1)the authors claim the advantage of WP system is to have excluded CP. however, with the proposed idea, somehow it is like CP, with 1/16 also considered in some cellular and 802.11 standards. The authors need to explain the overhead induced compared to other WP system without embedded SI.
2)more simulations are needed. In Fig.9, it is not a good excuse to leave out the BER curves just because it is not implemented in the references. With clear algorithm, it will be not difficult to obtain the curves, which will help a lot to confirm the proposed scheme.
3)again on simulation: Ref [6] gives a PAPR reduction scheme in WP modulation. so it is better to make the performance in Ref[6] as a benchmark. Otherwise, it is not a pure apple-to-apple comparison with present version.
4)some typo: in Page 8, one-sisteenth should be one-sixteenth; in Page 15, 'for the propose WP-PTS-ESID' should be 'for the proposed WP-PTS-ESID'

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A New PAPR Reduction Scheme: Wavelet Packet-based PTS with Embedded Side Information Data Scheme

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Abstract: Partial Transmit Sequence (PTS) is an effective scheme to reduce the high peak-to-average power ratio (PAPR) for the multicarrier modulation (MCM) signals transmission system. It produces the side information (SI) data as a result of the MCM signals optimization process. The generated SI data are required to be transmitted together with the original data over the channel for successful data recovery at the receiver. Effective method for the SI data transmission is still under research. Hence, we introduce a technique that embeds the SI data into the original data frame. In this work the Wavelet Packet-based Partial Transmit Sequence (WP-PTS) scheme has been selected as the MCM transmission method. The proposed scheme is known as the WP-PTS with embedded SI Data (WP-PTS-ESID). In this paper, a suitable reconstruction scheme for reconstructing the original data is also developed. Simulation result shows that the PAPR's performance of the proposed scheme improves up to 2.5dB at CCDF level of 10^{-4} as compared to the original WP-OFDM system without a PAPR reduction scheme with the number of selected disjoint subblock is 16.

1. Introduction

In recent years, research in Wavelet Packet-based Orthogonal Frequency Division Multiplexing (WP-OFDM) system has attracted many researchers. Implementing Wavelet Packet Transform (WPT) instead of Fast Fourier Transform (FFT) in OFDM system improves several issues:

1. Additional beneficial attributes to the signal property, known as inherent flexibility which provides better spectral efficiency where the orthogonal structure is found in both time-domain and frequency-domain of the signals [1, 2].
2. Remove the need for cyclic prefix in WP-OFDM system for intersymbol interference (ISI) and intercarrier interference (ICI) protections yet maintains BER performance [3, 4].

Certainly, the WP-OFDM transmitted signals also suffer from high peak-to-average power ratio (PAPR) due to abundant narrowband signals summed up in the time domain [5]. In general, the event of high PAPR causes nonlinear distortion of the signals and reduces the power efficiency of high power amplifier (HPA). Since to employ a transmitter of HPA with a wider linear region will cause a great cost as well as the overall system complexity is increased, hence, PAPR reduction is a good choice. Several excellent techniques have been proposed such as the Selected Mapping using Wavelet Packet Tree Pruning [6], Fast Adaptive Optimal Basis Search Algorithm [7], combination of iterative clipping and filtering method with Huffman Coding [8] and Genetic Algorithm [9, 10] to diminish PAPR.

Among of the recent well-known PAPR reduction techniques which is based on WP-OFDM systems, the Partial Transmit Sequence (PTS) technique is an attractive scheme since it can reduce PAPR with no distortion to the signal orthogonality. In [11], a PTS scheme in special multicarrier modulation system known as wavelet packet division multiplexing (WPDM) has been proposed using Daubechies-5 wavelet, *db*5. Another work in [12] proposes a method based on pruning the full-tree structure of wavelet packet modulation (WPM) with PTS scheme in order to reduce the number of nodes. Several works associate with low-density parity-check (LDPC) codeword have been found in [13, 14]. The authors develop the particular parity-check matrix for the concatenated LDPC-PTS code and the PTS side information or phase factor estimation are not needed before decoding [13]. Further improvement of previous work, the authors formulate an optimization problem to improve the joint decoding performance by optimizing the partition [14].

In the conventional PTS scheme which is based on discrete Fourier transform, input symbols are divided into V disjoint subblocks. Each subblocks is phase-rotated by particular phase factor during optimization process. The phase factor is the side information (SI) of the appointed frame and being sent along with main information. The receiver must have the uncorrupted SI in order to achieve successful data recovery [15, 16].

A very critical issue in common PTS schemes is the way they handle the side information (SI) in order to achieve successful data recovery. We do comprehend to the cream methods proposed in literature and classify them into two categories as the following:

1. Non-explicit SI approach. In [17], the work focuses on generating OFDM frames in time domain using a method called phase recursive cyclic (PRC) shifting. Using natural diversity of phase constellation, it detects and recovers the original signals. The work in [18] proposes that, for each subblock end of the OFDM frames, an additional pilot symbol is inserted deliberately. It can be decoded based on channel estimation at the receiver. In [19], the authors suggest to employ hard-decision maximum likelihood (ML) detector, a minimum mean square error (MMSE) estimator or a zero forcing (ZF) estimator in order to demap the quadrature amplitude modulation (QAM) symbols.

2. Modified-constellation approach. The work in [20] proposes the use of hexagonal constellation for PAPR mitigation. Since hexagonal constellation is the densest packing of spaced points in two dimensions, those two extra points can be exploited for SI elimination in the PTS scheme. The authors in [20] also prove that the uncoded 91-HEX modulation as proposed is comparable to the conventional 64-QAM modulation. In [21], the authors propose a reshaped quadrature amplitude modulation (R-QAM) constellation known as R-PTS scheme. At the receiver, the mean square error between the constellations of the received data and the R-QAM difference phase rotation factor are calculated. The work in [22] proposes a sub-optimal PAPR improvement using constellation extended scheme (CES) which is combined with the block-coded modulation (BCM) technique in the conventional PTS scheme. The extended constellation points of a 16-QAM modulation are arranged in symmetrical and asymmetrical structure where the points and error correction capabilities play important roles for the signals recovery.

However, in the first classified PTS method, i.e. the non-explicit SI approach, there is a drawback of an extra time is needed to detect or decode and as the complexity increased this cause a decreasing in effective transmission rate [23, 24]. Whilst, an unavoidable disadvantage to the second classified PTS method, i.e. the modified-constellation approach in which the variety type of constellation extensions are implemented hence requires higher transmit power to be transmitted for a mode of more dense constellation points [25, 26].

Motivated from the highlighted issue above, this paper addresses a new novel simple technique. In this paper, the SI data which are obtained through Phase Factor Optimization (PFO) processes, are embedded to the respective space in the frame as symbols and concatenated with the information symbols to form a complete frame.This frame is then transformed into the OFDM signals via WPT modulation. The entire proposed scheme is known as Wavelet Packet-based PTS with Embedded SI Data (WP-PTS-ESID). To the best of our knowledge, there has been no work published related to this idea.

Simulation results show the best performance of proposed WP-PTS-ESID scheme is 2.5dB PAPR improvement from the original WP-OFDM curve at CCDF level of 10^{-4} when the number of disjoint subblock chosen is 16. This achievement incurs only minor degradation to bit error rate (BER) performance when we compare the proposed scheme with other reknown works.

The rest of the paper is organized as follows: Section 2 describes the WP-OFDM system and briefs the conventional PTS scheme. The proposed WP-PTS with embedded SI index scheme and its reconstruction scheme are explained in Section 3. Simulation results are shown in Section 4. Finally, Section 5 concludes the paper.

2. Background

This section presents PAPR computation and describes the fundamental structure of the conventional PTS (C-PTS) scheme and its features.

2.1. Peak-to-Average Power Ratio

Consider an OFDM systems with an N subcarriers, an inverse discrete Fourier transform (IFFT) is applied to the complex-valued phase-shift keying (PSK) or quadrature amplitude modulation (QAM) input OFDM block $\mathbf{X} = [X_0, X_1, \ldots, X_{N-1}]^T$ of length N to generate the OFDM signal. The discrete time-domain OFDM signal is

$$x_n = \frac{1}{\sqrt{n}} \sum_{k=0}^{N-1} X_k e^{\frac{-j2\pi nk}{N}}, n = 0, 1, \cdots, N-1 \tag{1}$$

which can be also rewritten in matrix form as $\mathbf{x} = [x_0, x_1, \ldots, x_{N-1}]^T = \mathbf{F}^{-1}\mathbf{X}$, where $\mathbf{F}^{-1}$ is the inverse DFT matrix. In general, the PAPR of OFDM signal is x defined as

$$\text{PAPR} = \frac{\max |x(t)|^2}{E|x(t)|^2} \tag{2}$$

where $E[\cdot]$ denotes expectation.

2.2. Conventional PTS Scheme

In conventional PTS scheme, an input data block of length N is partitioned into several disjoint subblocks. The IFFT for each one of these subblocks is computed separately and then weighted by a phase factor. The phase factors are selected in such a way as to minimize the PAPR of the combined signal of all the subblocks [27, 28, 29]. Fig. 1 shows a block diagram of the OFDM

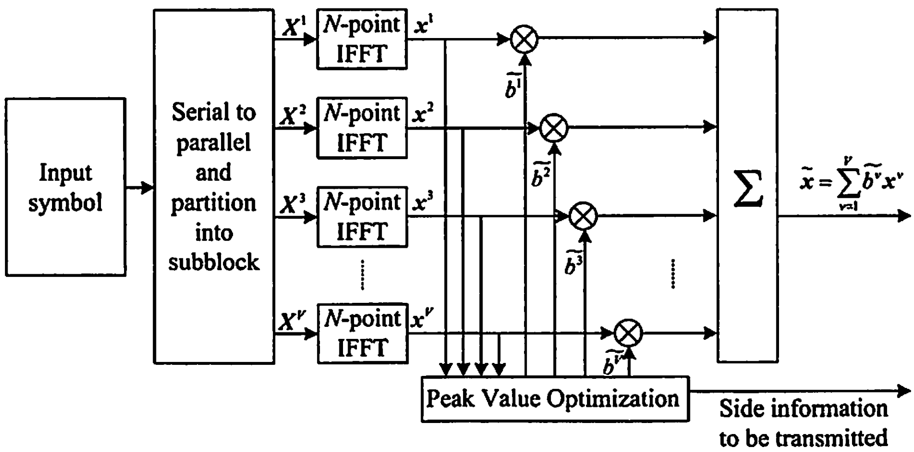

Fig. 1. Conventional PTS scheme.

transmitter with PTS scheme. Let an input data block, $\mathbf{X} = [X_0, X_1, \ldots, X_{N-1}]$ be partitioned into V disjoint subblocks, $\mathbf{X} = [\mathbf{X}^1, \mathbf{X}^2, \ldots, \mathbf{X}^\mathbf{V}]$, $1 \leq v \leq V$ such that any two of these subblocks are orthogonal and X is the combination of all the V subblocks as

$$\mathbf{X} = \sum_{v=1}^{V} \mathbf{X}^v \tag{3}$$

Then the IDFT for each subblock, $x^v, 1 \leq v \leq V$, is computed and weighted by a phase factor $b^v = e^{j\phi_v}$, where $\phi_v = [0, 2\pi)$, $1 \leq v \leq V$. The objective now is to select the set of phase factors, b^v's that minimizes the PAPR of the combined time domain signal $\mathbf{x}$, where $\mathbf{x}$ is defined as

$$\mathbf{x} = \sum_{v=1}^{V} b^v \cdot \mathbf{x}^v \tag{4}$$

where

$$\mathbf{x}^v = \text{IFFT}\left\{\mathbf{X}^v\right\} \tag{5}$$

During the stage of phase factor optimization (PFO), the search of the best phase factor, $\tilde{b}^v$ (i.e. tilde of b) is usually limited to a finite number of elements to reduce search complexity [5]. Assume that the set of allowed phase factors is defined as $b^v = e^{\frac{j2\pi k}{W}}$, where $k = 0, 1, \cdots, W-1$, and W is the number of allowable phase factors. The first phase factor b^1 usually set to 1 without any loss of performance, therefore,there are $V-1$ phase factors are to be found by an exhaustive search. Hence, W^{V-1} sets of phase factors are searched to find the optimum one. The reduction in PAPR attainable depends on V and W. It can be observed that the larger is the number of sub-blocks V, the greater is the reduction in PAPR but on the other hand, the search complexity is also increasing exponentially with V. In addition, V subblocks are needed to implement the PTS scheme, requiring $\lfloor \log_2 W^{V-1} \rfloor$ bits of side information to be transmitted [15]. The optimized transmitted signal with lowest possible PAPR can be written as

Table 1 Mathematical symbols used

Symbol	Description
b	Phase factor
$\tilde{b}$	Optimized phase factor
$\tilde{b}'$	Received phase factor
$\tilde{B}$	Grouped optimized phase factor
$\dot{b}$	Integer factor
$\dot{b}'$	Received integer factor
v	Index number of disjoint subblocks per frame
V	Maximum number of disjoint subblocks
w	Number of phase factor

$$\tilde{\mathbf{x}} = \sum_{v=1}^{V} \tilde{b}^v \cdot \mathbf{x}^v \tag{6}$$

3. Proposed WP-PTS-ESID Scheme

This section presents the proposed scheme known as Wavelet Packet-based Partial Transmit Sequence with Embedded SI Data (WP-PTS-ESID) Scheme. This section consists of three main parts: First, the general overview of the proposed scheme followed by the construction of data frame and finally the algorithms involved. In order to facilitate reading, the mathematical symbols used in this paper are listed in Table 1.

3.1. Overview of WP-PTS-ESID Scheme

The block diagram of the proposed WP-PTS-ESID scheme is shown in Fig. 2. An input sequence consists of the combination of two sub-sequences i.e. the dummy symbols and input symbols. This input sequence is called as the data frame. In Fig. 3, the structure of the data frame is shown. The SI data represents dummy symbols block with length of R while the original data represents input symbols with the length of P. Thus, for a single data frame, it has a length of N. Initially, each dummy symbol contains the value of 1 during the pre-optimization stage. This value may change after the optimization process. The output from the optimization process is called the SI data.

Now, let the sequence of P input symbol be $\mathbf{X_{data}}$ and the sequence of R symbols be $\mathbf{X_{SI}}$ are viewed as single data frame $\mathbf{X_{tx}}$, where

$$\mathbf{X_{tx}}[n] = \begin{cases} X_{SI}[n] & \text{for} \quad 0 \leq n < R \\ X_{data}[n] & \text{for} \quad R \leq n < N \end{cases} \tag{7}$$

where $R = N - P$. Then, the serial data frame $\mathbf{X_{tx}}$ are converted into parallel form and evenly divided into V disjoint subblocks. Let the disjoint subblock of $\mathbf{X_{tx}}$ along with the complex element

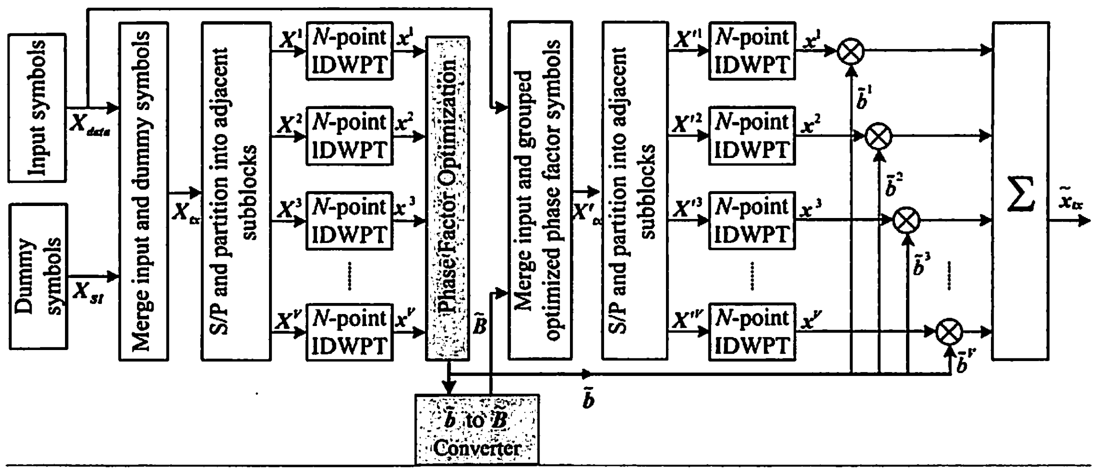

Fig. 2. Proposed WP-PTS-ESID scheme.

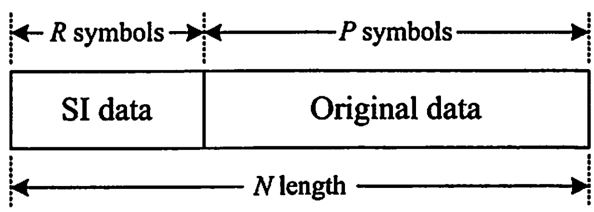

Fig. 3. Data Frame Structure.

b^v be represented as X, which is

$$\mathrm{X} = \sum_{v=1}^{V} b^v \cdot \mathrm{X_{tx}}^v \tag{8}$$

Note that, the complex element b^v is used to obtain the optimized signal prior transmission. Utilizing signal X, we produce the time-domain signal x obtained from IDWPT as

$$\begin{aligned} \mathrm{x} &= \mathrm{IDWPT}\{\mathrm{X}\} \\ &= \mathrm{IDWPT}\left\{\sum_{v=1}^{V} b^v \cdot \mathrm{X_{tx}}^v\right\} \end{aligned} \tag{9}$$

Exploiting the signal transform linear property Eq. (9) becomes

$$\mathrm{x} = \sum_{v=1}^{V} b^v \cdot \mathrm{IDWPT}\{\, \mathrm{X_{tx}}^v\} \tag{10}$$

Until this point, the term x is not shown in Fig. 2 but it is employed to generate the corresponding optimized phase factors known as SI data. The key that plays role as a signal optimizer is the parameter b^v. Hence, the $\{b^v \cdot \mathrm{IDWPT}\{\mathrm{X_{tx}}^v\}\}$ term is fed into Phase Factor Optimization block as shown in Fig. 2. This process searches the appropriate phase factors for subblock 1 to V that

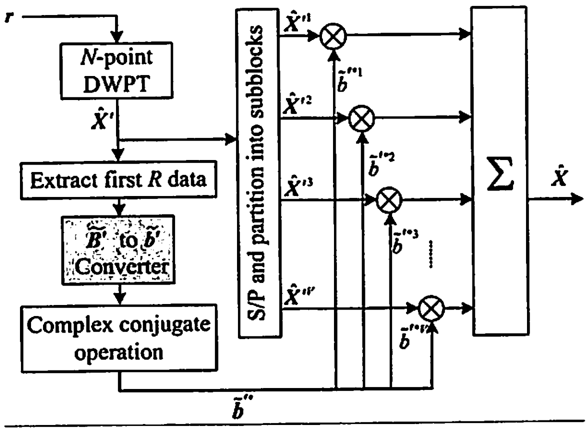

Fig. 4. Proposed reconstruction of WP-PTS-ESID scheme.

produce the lowest PAPR of the transmitted signal. To compromise with the fixed length R of the SI codeword, the length of V sequence of phase factors must be set to the length R via the process in "$\tilde{b}$ to $\tilde{B}$ Converter". These two algorithmic blocks are explained in Section 3.3.

The output from the "$\tilde{b}$ to $\tilde{B}$ Converter" block is $\tilde{B}$ or the SI data and Eq. (7) can be rewritten as

$$\mathbf{X}'_{\mathbf{tx}}[n] = \begin{cases} X_{SI}[n] = \tilde{B}[n] & \text{for} \quad 0 \leq n < R \\ X_{data}[n] & \text{for} \quad R \leq n < N \end{cases} \tag{11}$$

where the values of X_{SI} are optimized and may be grouped, and $R = N - P$. Finally, Eq. (10) can be rewritten as

$$\tilde{\mathbf{x}}_{\mathbf{tx}} = \sum_{v=1}^{V} \tilde{b}^v \cdot \mathrm{IDWPT}\{\mathbf{X_{tx}}^v\} \tag{12}$$

Fig. 4 shows the proposed reconstruction scheme of the original signal. The signal $\mathbf{r}$ is the received data and the original data $\hat{\mathbf{X}}$ are obtained after the DWPT operation. Then, the first R data is extracted from $\hat{\mathbf{X}}'$ and labelled as

$$\tilde{X}'_{SI}[n] = \tilde{B}'_n \qquad \text{for} \qquad 0 \leq n < R \tag{13}$$

Since $\tilde{B}'$ can be a single or group of several $\tilde{b}'$s, the block of "$\tilde{B}'$ to $\tilde{b}'$ Converter" is utilized to generate the $\tilde{b}'$ sequence by taking the complex conjugate operation. This block is further elaborated in Section 3.3. The complex conjugate of $\tilde{b}'$ elements can be described as

$$\tilde{\mathbf{b}}'^* = [\tilde{b}'^{*1}, \tilde{b}'^{*2}, \ldots, \tilde{b}'^{*V}] \tag{14}$$

Finally, $\hat{\mathbf{X}}'$ data sequence are also divided into evenly V disjoint adjacent subblocks and are multiplied with $\tilde{\mathbf{b}}'^*$. The recovered information $\hat{\mathbf{X}}$ is expressed as

$$\hat{\mathbf{X}} = \sum_{v=1}^{V} \tilde{b}'^{*v} \cdot \hat{\mathbf{X}}'^v \tag{15}$$

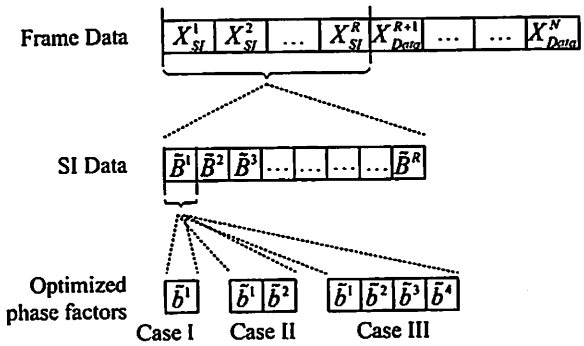

Fig. 5. Frame division and the element of SI data according to the cases.

3.2. Data Structure

As shown in Fig. 3 previously, the data frame structure of the input is made up by concatenating R symbols (of SI data) and P symbols (of original data). We employ the length of codeword R as one-sisteenth of N codeword length for simple division of data frame.

Another important parameter is the total number of disjoint adjacent subblock V. The number of phase factors consumed also must be the same as the number of V. In this paper, we decide to allocate any number of the multiple 2^j of the optimized phase factor elements into fixed length of codeword R, where j $\forall$ positive integer number. Hence, we define three cases as follows:

1. Case I – the total number of disjoint subblock, V is less than or equal to the length of SI data codeword, R i.e. $V \leq R$
2. Case II – the total number of disjoint subblock, V is double of SI data codeword length R i.e. $V = 2R$
3. Case III – the total number of disjoint subblock, V is four times the length of SI data codeword length R i.e. $V = 4R$

Fig. 5 illustrates the frame division and SI data element according to their cases. In Case I, each phase factor is represented by a single SI data. However, in Case II and III, the multiple phase factors are combined. Table 2 shows the relationship of parameters N, R, V and their various cases.

The another important parameter in this work is the number of phase factor that can be selected, w. In this work, $w = 2$ or 4 are chosen for simple analysis. In Table 3, the possible value for phase factor to be chosen during optimization process with respect to w is tabulated.

3.3. The Proposed Algorithms

There are two algorithmic blocks in the proposed scheme; Phase Factor Optimization and "$\tilde{b}$ to $\tilde{B}$ Converter". While at the receiver, "$\tilde{B}'$ to $\tilde{b}'$ Converter" is also developed.

Algorithm #1: Phase Factor Optimization

The aim of this algorithm is to find the most appropriate phase factor, $\tilde{b}$ for each disjoint subblock to produce the lowest PAPR value. The optimized phase factor sequence is denoted as $\underline{\text{b}}$. These parameters of V and w must be defined before the operation begins.

Table 2 Relationships between N, R, V and cases

N	R	V	Case type
64	4	1	I
		2	I
		4	I
		8	II
		16	III
		32	n/a
		64	n/a
128	8	1	I
		2	I
		4	I
		8	I
		16	II
		32	III
		64	n/a
256	16	1	I
		2	I
		4	I
		8	I
		16	I
		32	II
		64	III

Table 3 Possible value for phase factor, b

No. of phase factor, w	Phase factor, b
2	$1, -1$
4	$1, j, -1, -j$

Refering to Algorithm 1, the term v is used as a counter for parameter V where v is the current disjoint subblock. The term m is used as a counter for parameter w where m is the integer that associates the possible value of phase factor shown in Table 3. If $w = 4$ is chosen, then the counter of m has the component of $m = \{1, 2, 3, 4\}$, hence the corresponding phase factor b is $b =$

Algorithm 1 Phase Factor Optimization

Require: V, w and $\mathbf{X_{tx}}$
Ensure: $\tilde{\mathbf{b}}^{\mathbf{v}} = [\tilde{b}^1, \tilde{b}^2, \tilde{b}^3, \ldots, \tilde{b}^V]$

$v \leftarrow 1$
$m \leftarrow 1$
$\mathbf{x_{tx}} = \sum_{v=1}^{V} b^v \cdot \mathrm{IDWPT}\{\, \mathbf{X_{tx}}^v \}$
$PAPRm \leftarrow PAPR(\mathbf{x_{tx}})$ {Calculate initial PAPR}
if $v = 1$ **then**
  $PAPRmin \leftarrow PAPRm$
  $v \leftarrow v + 1$
end if
for $v = 2$ **to** V **do**
  for $m = 1$ **to** w **do**
    $b_{before} \leftarrow b^v$ {b^v is copied from b sequence element of order v}
    $b^v \leftarrow b^m$
    $\mathbf{x_{tx,new}} = \sum_{v=1}^{V} b^v \cdot \mathrm{IDWPT}\{\, \mathbf{X_{tx}}^v \}$
    $PAPRm \leftarrow PAPR(\mathbf{x_{tx,new}})$
    if $PAPRm < PAPRmin$ **then**
      $PAPRmin \leftarrow PAPRm$
    else
      $b^v \leftarrow b_{before}$
    end if
  end for
end for
for $v = 1$ **to** V **do**
  $\tilde{b}^v \leftarrow b^v$
end for

$\{1, j, -1, -j\}$. Besides, in this algorithm the term b_{before}, *PAPRm* and *PAPRmin* are considered as the memory elements.

The function to compute PAPR for the current data frame is indicated as $PAPR\{\cdot\}$. Finally, this algorithm produces the optimized phase factors sequence, denoted as $\tilde{\mathbf{b}}$.

Algorithm #2: $\tilde{b}$ to $\tilde{B}$ Converter
This block receives the optimized phase factor elements, $\tilde{\mathbf{b}}$ in multiple of 2^j integers and converts them into R elements of SI data. For this purpose, the parameters N, V and case type must be identified by referring to Table 2.

For example, during the optimization process when $w = 4$, the element in $\tilde{\mathbf{b}}$ sequence consists of $\{1, j, -1, -j\}$. We have to manipulate these elements into integer factor (i.e. symbol with upper dot) as represented in Table 4.

Then, to generate the side information (SI) data, $\tilde{\mathbf{B}}$ sequence, we introduce SI integer, M_B. The relationships between integer factor (i.e. $\dot{b}$) and SI integer (i.e. M_B) is according to the case type as follows:

Table 4 Current phase factor to integer factor conversion

Optimized phase factor, $\tilde{b}$	Integer factor, $\dot{b}$
1	0
-1	1
j	2
$-j$	3

Case 1: $V \leq R$

$$M_B = \dot{b} \tag{16}$$

Case 2: $V = 2R$

$$M_B = \dot{b}^1 \times w^1 + \dot{b}^2 \times w^0 \tag{17}$$

Case 3: $V = 4R$

$$M_B = \dot{b}^1 \times w^3 + \dot{b}^2 \times w^2 + \dot{b}^3 \times w^1 + \dot{b}^4 \times w^0 \tag{18}$$

where $\dot{b}^k$ is the integer factor, (k is the order within the current group of data frame) while w^n is the corresponding number of allowable phase factors to the power of n.

In order to calculate the SI data, $\tilde{B}$, two parameters are required: SI integer, i.e. M_B and the maximum value of SI integer, $\max(M_B)$. The SI data, $\tilde{B}$ can be calculated as follows:

$$\tilde{B} = \cos\left(\frac{M_B}{\max(M_B)+1} \times 2\pi\right) + j\sin\left(\frac{M_B}{\max(M_B)+1} \times 2\pi\right) \tag{19}$$

The value of $\max(M_B)$ can be determined according to the case type. Through out the result view, we employ the number of phase factor, $w = 2$. Table 5 shows the important parameters and the value of $\max(M_B)$. The calculation of $\max(M_B)$ is based on Eq. (16)-(18) when maximum value of base 2 is considered.

Table 5 Integer factor with maximum value.

Case	Integer factor codeword	Max. value of base w ($w = 2$)	$\max(M_B)$
I	$\dot{b}$	1	1
II	$\dot{b}^1\dot{b}^2$	11	3
III	$\dot{b}^1\dot{b}^2\dot{b}^3\dot{b}^4$	1111	15

Algorithm #3: $\tilde{B}'$ to $\tilde{b}'$ Converter

There are R componenet of the SI data extracted from received data frame $\hat{\mathbf{X}}'$ (refer to Fig. 4). This

Algorithm 2 To obtain the exact $M_B{}'$

Require: $M_{B1}{}'$, $M_{B2}{}'$ and $\max(M_B{}')$
Ensure: $M_B{}'$

$sign = M_{B2}{}'/|M_{B2}{}'|$
if $(sign = 0$ **or** $sign \geq 0)$ **then**
  $M_B{}' \leftarrow M_{B1}{}'$
else if $(sign \leq 0)$ **then**
  $M_B{}' \leftarrow \max(M_B{}') + 1 - M_{B1}{}'$
end if

block reproduces V phase factors for data reconstruction. Thus, each $\tilde{B}'$ symbol can be rewritten to follow Eq. (19) as

$$\begin{aligned}\tilde{B}' = \cos&\left(\frac{M_B{}'}{\max(M_B{}')+1} \times 2\pi\right) + \\ j\sin&\left(\frac{M_B{}'}{\max(M_B{}')+1} \times 2\pi\right)\end{aligned} \tag{20}$$

where the prime notation means that the parameters are at the receiver part i.e. $M_B{}'$ is equivalent to M_B (as in the transmitter). Hence, for each element of $\tilde{B}'$ symbol, we get two different representation for the term $M_B{}'$ (which is labelled as $M_{B1}{}'$ and $M_{B2}{}'$) as follows:

$$M_{B1}{}' = \cos^{-1}\left(\tilde{B}'_{real}\right) \cdot \frac{\max(M_B{}')+1}{2\pi} \tag{21a}$$

$$M_{B2}{}' = \sin^{-1}\left(\tilde{B}'_{imag}\right) \cdot \frac{\max(M_B{}')+1}{2\pi} \tag{21b}$$

where $\tilde{B}'_{real}$ is the real part of $\tilde{B}'$ and $\tilde{B}'_{imag}$ is its imaginary part. Using Algorithm 2, we can obtain $M_B{}'$ perfectly. Then, the received integer factors $\dot{b}'$ can be found by manipulating $M_B{}'$ according to the following:

Case 1: $V \leq R$

$$\dot{b}' = M_B{}' \tag{22}$$

Case 2: $V = 2R$

$$\dot{b}'^1 = \lfloor M_B{}'/w^1 \rfloor \tag{23a}$$

$$\dot{b}'^2 = M_B{}' \bmod w^1 \tag{23b}$$

Case 3: $V = 4R$

$$\dot{b}'^1 = \lfloor M_B{}'/w^3 \rfloor \tag{24a}$$

$$\dot{b}'^2 = \left\lfloor \frac{M_B{}' - \dot{b}'^1 \times w^3}{w^2} \right\rfloor \tag{24b}$$

$$\dot{b}'^3 = \left\lfloor \frac{M_B{}' - \dot{b}'^1 \times w^3 - \dot{b}'^2 \times w^2}{w^1} \right\rfloor \tag{24c}$$

$$\dot{b}'^4 = (M_B{}' - \dot{b}'^1 \times w^3 - \dot{b}'^2 \times w^2) \bmod w^1 \tag{24d}$$

Table 6 Received integer factor to received phase factor conversion

Received integer factor, $\dot{b}'$	Received phase factor, $\tilde{b}'$
0	1
1	-1
2	j
3	$-j$

where w^n is the corresponding number of the allowable phase factors to the power of n, $\lfloor \cdot \rfloor$ is the floor bracket and a mod b returns the remainder from the division of a by b.

Finally, based on Table 6, we change each of the received integer factor to the received phase factor elements for the recovering the information on the next operation of complex conjugate.

4. Results and Discussion

In this paper, we study the performance of the proposed WP-PTS-ESID scheme and compares with other two re-known schemes categorized as non-explicit SI (i.e. work in [17]) and modified-constellation (i.e. work in [20]) which have been discussed in Section 1. Since there has been no work presented on SI management for wavelet packet-based PTS scheme then a typical one-to-one comparison cannot be done. Hence, in order to have a sense of performance comparison measure for the propose WP-PTS-ESID scheme, we compare our work with the conventional Fourier-based PTS scheme. In this comparison, we mainly focus on the trade-off between PAPR reduction against BER performance.

The complementary cumulative distribution function (CCDF) of the PAPR is utilized to evaluate the performance. The CCDF of the PAPR is defined as

$$\mathrm{CCDF}(\mathrm{PAPR}(x[n])) = \Pr(\mathrm{PAPR}(x[n]) > \mathrm{PAPR}_0$$

where PAPR_0 is a certain threshold value that is usually given in decibels relative to the root mean square (RMS) value.

Fig. 6 shows the CCDFs of the proposed WP-PTS-ESID scheme for the set of the first approach of non-explicit SI which employs 16-QAM modulation with subcarriers $N = 256$ with subblock $V = 4, 8$ respectively. The plot of the original uncoded WP-OFDM system which is without PAPR reduction is also shown. We also consider a comparison with the scheme in [17] with has the same main design parameters. This figure shows when $V = 8$ the proposed scheme is able to reduce the PAPR signal almost 2dB from its uncoded WP-OFDM curve at CCDF of 10^{-4}.

The BER performance of the proposed WP-PTS-ESID scheme over AWGN channel for sub-block $V = 4, 8$ respectively with 16-QAM modulation are presented in Fig. 7. For comparison purpose, we include the BER curve from scheme in [17]. From this figure, a good BER performance can be achieved through the proposed WP-PTS-ESID scheme when compared as the curve plotted from [17]. It is explicitly seen that the BER of the proposed scheme is almost the same with the WP-OFDM curve which is no PAPR reduction involved and around 4dB degradation is experienced by the scheme in [17] at the similar BER level of 10^{-4}. We postulate that the scheme in [17] works under sub-optimal process deteriorate the BER behaviour indirectly.

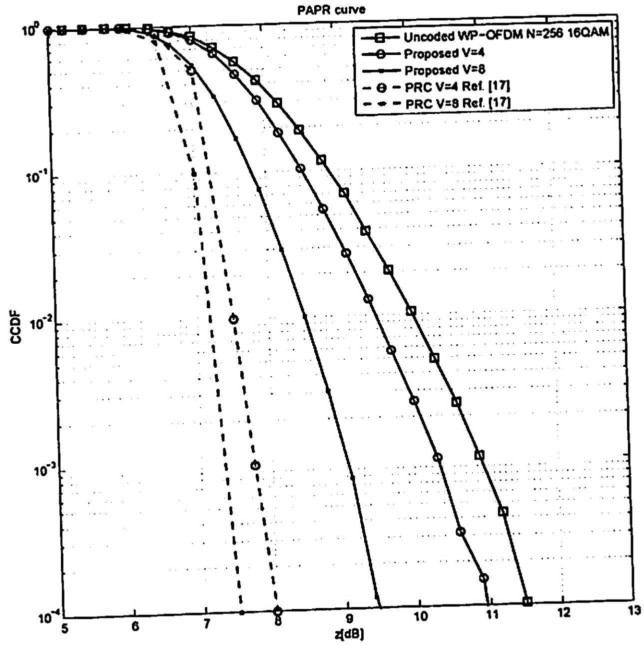

Fig. 6. CCDF of the proposed WP-PTS-ESID scheme compared to the PRC Scheme as in [17], for $V = 4$ *and* 8.

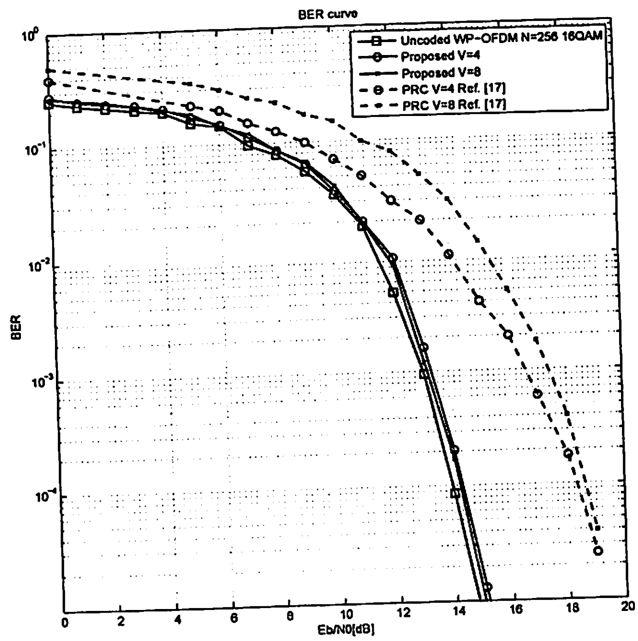

Fig. 7. BER performance of the proposed WP-PTS-ESID scheme, compared to the PRC Scheme as in [17], for $V = 4$ *and* 8.

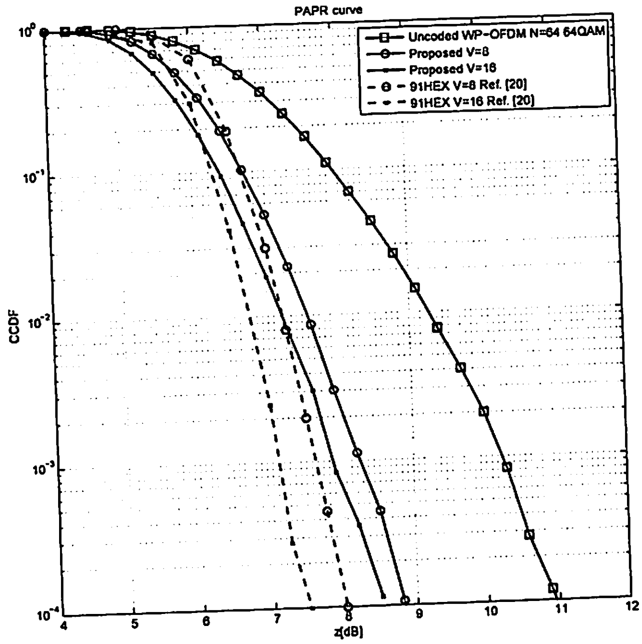

Fig. 8. CCDF of the proposed WP-PTS-ESID scheme compared to scheme represented as 91-HEX as in [20], $V = 8$ *and* 16.

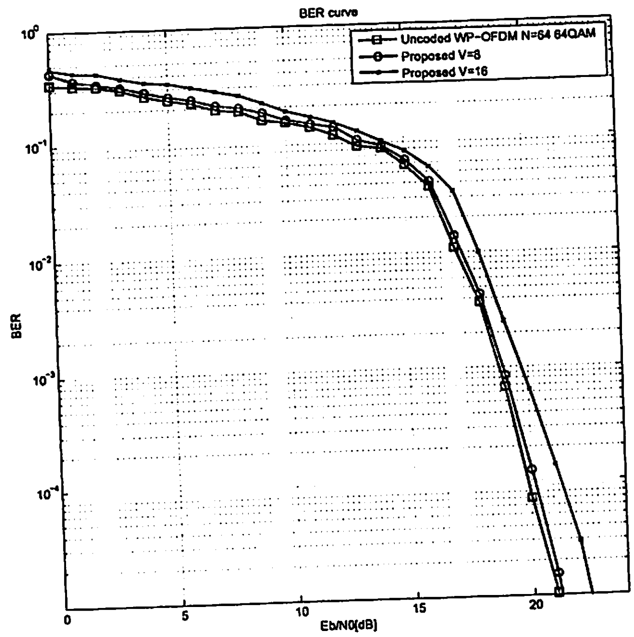

Fig. 9. BER performance of the proposed WP-PTS-ESID scheme for $V = 8$ *and* 16.

In this paper, we also analyze another set of parameters for second approach of modified-constellation i.e. 64-QAM modulation technique with 64 subcarriers and compared with the CCDF of the work in [20]. Fig. 8 shows the CCDF of the proposed scheme with subblock $V = 8, 16$ and the PAPR curve from the re-known work done in [20]. We also plot the CCDF of the uncoded WP-OFDM system for comparison purpose. It can be clearly seen that the proposed scheme offers improvement in PAPR almost 2.5dB at CCDF of 10^{-4} for the chosen subblock $V = 8$ as compared to the uncoded uncoded WP-OFDM curve at CCDF. The PAPR reduction for the proposed WP-PTS-ESID scheme and the work in [20] are almost 1dB difference at CCDF of 10^{-4}. Fig. 9 show the BER performance of the proposed scheme with the uncoded curve with similar modulation technique and number of subcarriers for comparison. It is confirmed that our method gives not much degradation on signals quality at the receiver. Since in the work [20] did not present the BER performance thus, in this paper no comparison done in term of BER with that work.

We like to inform that all the comparison results is based on the number of phase factor, $w = 2$ since the reknown works implement the similar value.

5. Conclusion

In this paper, we proposed a new PAPR reduction technique which known as Wavelet Packet-based PTS with Embedded SI Data (WP-PTS-ESID) scheme. This scheme embeds the side information data into its data frame. Simulation results shows that the proposed WP-PTS-ESID scheme can achieve PAPR reduction as that of the reknown works. Commonly, the BER increase of the proposed scheme is insignificant as compared to the uncoded original signals.

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Three description lattice vector quantization for efficient data transmission

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Lattice vector quantization is an ascendant technique that suits very well for multiple description coding (MDC) system. This paper introduces the lattice ℤ4 to be utilized for labeling function in the three description MDC system that consists of three encoders and seven decoders. Projection of a tesseract in three-space of lattice ℤ4 yields four outputs and the data are transmitted via three channels where one of the outputs is defined as time. Simulation results show that the three description quantization system is efficient that provides low distortion and good PSNR reconstruction quality.

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Three Description Lattice Vector Quantization for Efficient Data Transmission

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***Abstract*—Lattice vector quantization is an ascendant technique that suits very well for multiple description coding (MDC) system. This paper introduces the lattice $\mathbb{Z}_4$ to be utilized for labeling function in the three description MDC system that consists of three encoders and seven decoders. Projection of a tesseract in three-space of lattice $\mathbb{Z}_4$ yields four outputs and the data are transmitted via three channels where one of the outputs is defined as time. Simulation results show that the three description quantization system is efficient that provides low distortion and good PSNR reconstruction quality.**

Keywords—Multiple description; lattice vector quantization; labeling function

I. Introduction

Prior to the introduction of multiple description coding (MDC) by Ozarow in [1-3], there were two ways in dealing with packet transmission lost i.e. using the repeat transmission request and reconstruct the received data using the available packets. In request for a retransmission, it is feasible to have a lossless transmission data, while for the later method requires some redundancy data to be added in the packets [4].

Multiple description lattice vector quantization (SVS-MDLVQ) was first introduced by Servetto, Vaishampayan and Sloane in [5]. In that work, the authors produced two descriptions by exploiting the lattice codebook for transmission through two different channels. In [6], the authors discussed detail analysis of SVS-MDLVQ's extension. The concerns on asymmetric MDLVQ have been studied in [7], [8] and [9]. These works improved the labeling function in [5] which the authors considered only the balanced and symmetric lattice quantizer. The work in [8] introduced a source coding scheme to traverse the balanced multiple description quantizers and concurrently introduced an entirely hierarchical successive refinement quantizers technique in the system.

The design of symmetric entropy constrained MDLVQ is introduced in [10]. This design is essentially a greedy type and the algorithm optimizes the index assignment. The asymptotic analysis of the multiple description vector quantization (MDVQ) with lattice codebook for sources with smooth probability density functions (pdfs) is discussed in [11]. It also disclosed that the uniform central quantizer cells are not optimal in finite dimension.

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In recent years, majority of researches who work in MDC are concerned with two description only as presented in [12-16]. However, most of the practical applications require more than two packets of transmission for achieving better reconstruction data quality since the packets transmitted over Internet protocol (IP) networks are limited in size. In fact, a three description coding system can be more efficient since it provides lower distortion and promote better PSNR reconstruction quality as shown in [17-20].

There have been several works coming that introduce more than two description system as presented in [17-23]. In [17], the authors extended the SVS-MDLVQ work [5], yet the solutions for issue regarding index assignment mapping were unconvincing. The index assignment for three description system was introduced in [18] and in [19] the authors modified the index assignment algorithm by using the concept of two dimension hexagonal lattice A_2. However, in that work the authors considered only two-dimensional lattice.

In [20] the authors introduced a scalar quantization technique for three description system (MDSQ). In [21] the work combined the MDSQ with wavelet based image coding technique. In [22], the authors introduced a three description hexagonal A_2 lattice vector quantization using coinciding similar sublattice (MDCLVQ) which abolishes the traditional labeling function. The work used the coinciding sublattice vector quantizer which was inspired by the work in [23]. Hexagonal lattice sustains more distortion if compare to lattice $\mathbb{Z}$ as proven in work [24]. The vectors are more inclined to appear in the area of horizontal cross hatching, as granted by square lattice $\mathbb{Z}$ rather than diagonal cross hatching which furnished by the hexagonal lattice.

In this paper, a three description multiple description coding with $\mathbb{Z}_4$ lattice vector quantization (3DLVQ-$\mathbb{Z}_4$) is proposed. The scheme consists of three encoders and seven decoders (including a central decoder). In this system, K-channels transmit K-description into corresponding encoders with (2^K-1) decoders. The assumption is that all these encoders are capable to transmit information through the channels reliably. The encoders produce three equivalent rate, R_1 , R_2 and R_3 bits/source sample (bpss). The three description are transmitted separately through the channels accordingly. The correspondence description is allowed to have different distortions. In this paper, a new labeling algorithm using $\mathbb{Z}_4$ lattice for three description multiple description coding (3DLVQ- $\mathbb{Z}_4$) is proposed. The main contribution of this paper is the use of lattice $\mathbb{Z}_4$ for the labeling function. In this method the lattice $\mathbb{Z}_4$ offers more lattice points as neighbours that lead the central decoder to achieve better reconstruction quality.

Simulation results show that for central decoder the PSNR of the proposed 3DLVQ-$\mathbb{Z}_4$ accomplishes 42.63 dB and 32.13 dB for average side decoders, both at bit rate of 1.0 bpp. Simultaneously, the proposed 3DLVQ-$\mathbb{Z}_4$ expands the central reconstruction quality from 7.9 % to 9.2 % over the 3DLVQ-$\mathbb{Z}_2$ at varying bit rate.

The rest of the paper is presented as follows. The proposed technique with labeling function for the lattice $\mathbb{Z}_4$ at $N=9$ is presented in Section 2. Section 3 presents thorough experimental results and analysis of the proposed 3DLVQ-$\mathbb{Z}_4$ at $N=9$. In Section 4 concludes the paper.

II. Proposed Technique

The data source, x are processed in the encoders which consists of two main parts i.e. the lattice vector quantizer and the labeling function known as index assignment. The lattice vector quantizers map the data source x to the nearest codeword in the lattice codebook to minimize the distortion. The labeling function produces the codeword indexes then maps them into the three labels (description), each to be transmitted over separate channel, respectively.

The decoder will notice which description is accrued and then decodes accordingly. In this scheme, the distortion caused by using all three description is defined as central distortion. Side distortion refers to the distortion resulted by using only one or two description. The reconstructed output refers to the combination of the decoded data.

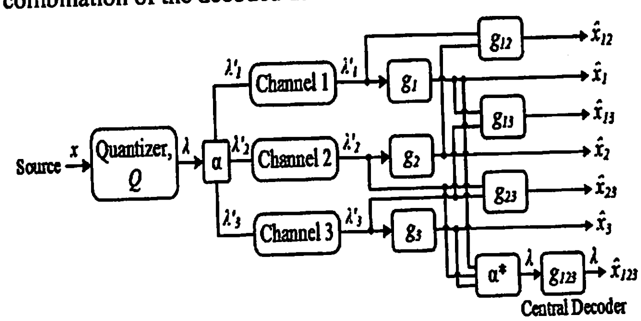

Fig. 1. The proposed MDLVQ scheme with three description.

A simplified block diagram of the proposed three description of lattice vector quantization (3DLVQ) system is demonstrated as in Fig. 1. The quantization of a vector x to the nearest vector λ in a lattice is denoted by $\Lambda \subset \mathbb{R}^4$. The mapping of the quantizer is expressed as $\lambda = Q(x)$. The code vector λ is sent via three channels, subject to rate constraint imposed by the individual channel. Assumptions are made that the sublattice of Λ contains the codebooks of channel 1, channel 2 and channel 3 respectively. These codebooks are denoted as Λ'_1, Λ'_2 and Λ'_3, respectively.

N_i is defined as the reuse index of sublattice Λ'_i and the index $[\Lambda : \Lambda'_i]$ is indicated by N_i, $i=1,2,3$. Suppose each of Λ'_i is geometrically similar to Λ, which means that when a similarity (a rotation, scale changing or reflection) is applied, Λ'_i can be acquired from Λ. However, in this paper, reflection is forbidden in order to simplify the analysis. Remark that the points in the lattice Λ are denoted as λ and the points in the sublattice Λ' are denoted as λ'.

The labeling function α maps $\lambda \in \Lambda$ to $(\lambda'_1, \lambda'_2, \lambda'_3) \in \Lambda' \times \Lambda' \times \Lambda'$ where Λ' be a sublattice of Λ with index $N = 9$. Given that the data transmitted to channel 1 is used to identify a code vector $\lambda'_1 \in \Lambda'_1$, the data transmitted to channel 2 is used to identify a code vector $\lambda'_2 \in \Lambda'_2$ and the data transmitted to channel 3 is used to identify a code vector $\lambda'_3 \in \Lambda'_3$. The component functions of α are denoted as $\alpha_1(\lambda) = \lambda'_1$, $\alpha_2(\lambda) = \lambda'_2$ and $\alpha_3(\lambda) = \lambda'_3$.

The α mapping is required to restitute λ respectively when all of the three channels are working. This can be achieved by setting the ordered pair $(\lambda'_1, \lambda'_2, \lambda'_3)$ used only once in the labeling scheme. In each of the side decoders, if there is only one channel available, for instance a channel 1, the corresponding received the sublattice point is selected to decode λ'_1. Similarly this can happen to channel 2 or channel 3 as well. Therefore, in this case the quality of reconstruction will be very low yet acceptable. If there are two of the three channels available, i.e., if channel 1 and channel 2 are available, the received data is used to decode (λ'_1, λ'_2). Similarly there could be the combination of channel 1 and channel 3 (λ'_1, λ'_3) or channel 2 and channel 3 (λ'_2, λ'_3).

This proposed scheme maps the given input x to the quadruple $(\lambda, \lambda'_1, \lambda'_2, \lambda'_3)$ ordered pair mapping. The three channels distortion $\bar{d}_{123}$ is given by $\|x-\lambda\|^2$, the side distortions $\bar{d}_i$ by $\|x-\lambda_i\|^2$, where $i = \{1,2,3,12,13,23\}$, as shown in Fig. 1. Suppose that the inner product of 4-dimensional vectors $x = (x_1, x_2, x_3, x_4)$ and $y = (y_1, y_2, y_3, y_4)$ are expressed as:

$$\langle x, y \rangle = \frac{1}{4} \sum_{i=1}^{4} x_i y_i \tag{1}$$

The Euclidean norm is defined as $\|x\| = \langle x, y \rangle^{\frac{1}{2}}$. Notice that the inner product and the Euclidean norm are dimensional-normalized. Hence, the relevant average distortion are given by

$\bar{d}_{123}$, $\bar{d}_1$, $\bar{d}_2$, $\bar{d}_3$, $\bar{d}_{12}$, $\bar{d}_{13}$ and $\bar{d}_{23}$. The goal of this paper is to design the labeling function α in order to minimize $\bar{d}_{123}$ subject to $\bar{d}_i \le D_i$, for given rates (R_1, R_2, R_3) and distortions D_i, where $i = \{1,2,3,12,13,23\}$.

The average three channels distortion is expressed as [8]:

$$\bar{d}_{123} = \sum_{\lambda\in\Lambda} \int_{V_\Lambda(\lambda)} \|x-\lambda\|^2 dx \tag{2}$$

In order to label the lattice $\mathbb{Z}_4$ in term of Λ using the sublattice Λ' of index $[\Lambda:\Lambda'] = N$, some requirements are set to abridge the encumbrances of lattice construction. Suppose that the generator matrix, G generates lattice Λ, so as a matrix G' generates similar sublattice Λ', this requires the generator matrix of $c\tilde{G}G$, where c is a scalar and $\tilde{G}$ is a unitary matrix. The smallest group of $\Gamma = \{I_L, -I_L\}$ is used to normalize the matrix G, where I_L is the L-dimensional an identity matrix.

The Γ normalizes the matrix G and denoted as $\tilde{G}$, where G and $\tilde{G}$ are given as:

$$G = \begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{pmatrix} \tag{3}$$

$$\tilde{G} = \begin{pmatrix} a & -b & -c & -d \\ b & a & -d & c \\ c & d & a & -b \\ d & -c & b & a \end{pmatrix} \tag{4}$$

Let $G' = \tilde{G}G$. The generator matrices are $G = I_4$ and the group Γ are given as:

$$\Gamma = \left\{ \begin{matrix} \pm I_4, \pm \begin{pmatrix} 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 \\ 0 & 0 & -1 & 0 \end{pmatrix}, \\ \pm \begin{pmatrix} 0 & 0 & -1 & 0 \\ 0 & 0 & 0 & -1 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{pmatrix}, \pm \begin{pmatrix} 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \\ 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{pmatrix} \end{matrix} \right\} \tag{5}$$

In this proposed scheme, $\Gamma = \{I_4, -I_4\}$ since it is a 4-dimensional MDLVQ system. Say that lattice $\mathbb{Z}_4$ has a geometrically-similar and clean sublattice of index N if and only if N is odd. In this work, the index N is set to nine. Therefore, there are nine points to be produced from the three labels.

Let express N in form of $N = a^2 + b^2 + c^2 + d^2$ and this is true at any integer, $(a,b,c,d) \subset \mathbb{Z}$. The set of admissible index values for lattice $\mathbb{Z}_4$ of the clean similar sublattice Λ' is given by integer sequence A016754 [25]: $1, 9, 25, 49, 81, 121, 169, 225, 289, 361, \cdots$

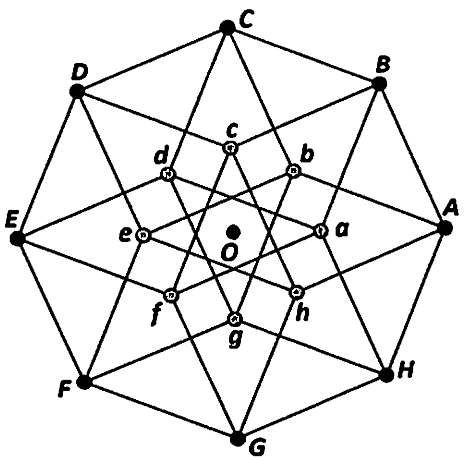

Fig. 2. Lattice $\mathbb{Z}_4$ with the Voronoi region and the discrete Voronoi region in four dimensional view.

The lattice $\mathbb{Z}_4$ is on a tesseract based, also called the hypercube in R^4 [26]. The term tesseract is denoted as four lines from each vertex to other vertices. A projection of the tesseract of lattice $\mathbb{Z}_4$ into three-space produces nine lattice points ($N = 9$) as illustrated in Fig. 2. The point of origin is labelled as O is basically a sphere. Lower case letters indicate the lattice points and the upper case letters indicate the sublattice points. The discrete Voronoi set or sublattice with N elements point is denoted as $\lambda' \in \Lambda'$. The Voronoi region, $V(\lambda)$ and the discrete Voronoi region, $V_0(\lambda')$ are defined as:

$$V(\lambda) = \{x : \|x-\lambda\| \le \|x-\tilde{\lambda}\|, \forall \tilde{\lambda} \in \Lambda\} \tag{6}$$

$$V_0(\lambda') = \{\lambda \in \Lambda : \|\lambda-\lambda'\| \le \|\lambda-\lambda''\|, \forall \lambda'' \in \Lambda'\} \tag{7}$$

Given that $|V_0(0)| = 9$. The discrete Voronoi set for the lattice $\mathbb{Z}_4$ as in Fig. 2 is shown as:

$$V_0(0)=\{O,a,b,c,d,e,f,g,h\} \tag{8}$$

The projection of the triple tuple points $(\lambda_1',\lambda_2',\lambda_3')$ into the tesseract in Fig. 2 produces four outputs, where one of them is defined as time. The rest are transmitted through the three description respectively.

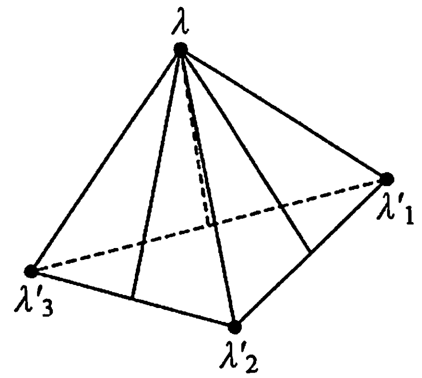

Fig. 3. Lattice point and its label $(\lambda_1',\lambda_2',\lambda_3')$ for lattice $\mathbb{Z}_4$.

The new labeling function maps the sublattice point to the three nearest lattice points which satisfy the triple ordered pair $(\lambda_1',\lambda_2',\lambda_3')$ condition. Mapping process of a lattice point λ to a triple ordered pair $(\lambda_1',\lambda_2',\lambda_3')$ is shown in Fig. 3. The distance of the edges should be as short as possible for the three tuple points $(\lambda_1',\lambda_2',\lambda_3')$. Moreover, the centroid of the three tuple points $(\lambda_1',\lambda_2',\lambda_3')$ should be as near as possible to the lattice point, λ.

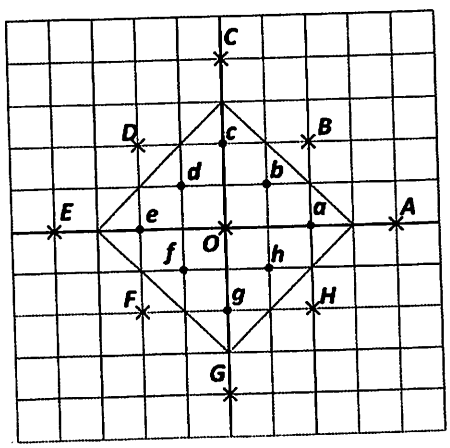

Fig. 4. Lattice point and its label $(\lambda_1',\lambda_2',\lambda_3')$ for lattice $\mathbb{Z}_4$ in Cartesian sight.

Fig. 4 portrays the Voronoi region, $V(0)$ and the discrete Voronoi region, $V_0(0)$ for lattice $\mathbb{Z}_4$ in Cartesian sight. This simplification is made to abbreviate the complexity of the labeling process. Lattice points are labeled by lower case letters and the sublattice points are labeled by the upper case letters. Point O is the origin for both sublattice and lattice $\mathbb{Z}_4$. The lattice point is having a rotation of 45° for one orbit, equally rotated by $\frac{\pi}{4}$ radians about the origin. By applying directed edges to label the points in the discrete Voronoi set ($V_0(0)$) acquires nine labels:

$$\varepsilon_u(0)=\begin{Bmatrix}\{O,A\},\{O,B\},\{O,C\},\{O,D\}, \\ \{O,E\},\{O,F\},\{O,G\},\{O,H\}\end{Bmatrix} \tag{9}$$

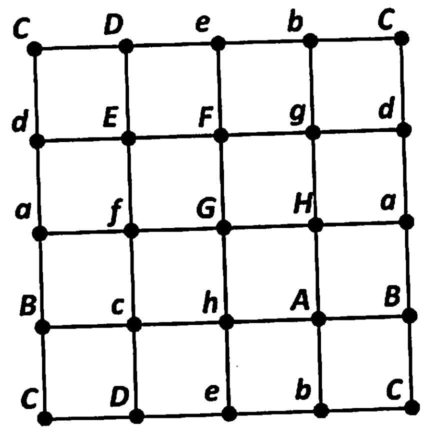

Fig. 5. Isomorphic adjacency of vertices in a four dimensional cube for the lattice $\mathbb{Z}_4$.

Fig. 5 illustrates the isomorphic adjacency of the vertices in a four dimensional cube, also known as a tesseract as displayed in Fig. 2. The figure shows a 4×4 array with opposite edges have been identified. The same 16 subsets or points can be arranged in a 4×4 array, when the array's opposite edges are joined together, the same adjacencies as those of the above four dimensional cube of the lattice $\mathbb{Z}_4$ can be achieved.

III. Results Analysis And Discussions

The proposed labeling algorithm using lattice $\mathbb{Z}_4$ for 3DLVQ scheme is developed using MATLAB R2014a (8.3.0.532) on Intel core i7-2630QM, 2.00 GHz, 8.00 GB RAM in the Windows 8.1 Pro environment. Training image of grey-scale Lena of size 512×512 is used for analysis. Ideal MDC network is considered in this experiment for its description to have either intact or completely lost.

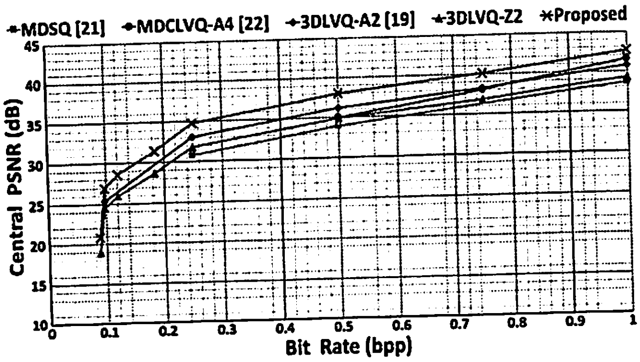

Fig. 6. Comparisons of central PSNR (dB) at varying bit rate (bpp).

Performance comparison of the central decoder reconstruction quality between the proposed scheme with three renowned schemes [21], [22], [19] and 3DLVQ- $\mathbb{Z}_2$, are as shown in Fig. 6. Obviously, multiple description scalar quantization gives the worst reconstruction quality, which yields only 28.88 dB at bit rate lower than 1.0 bpp. For the proposed scheme, central decoder PSNR attains as 42.63 dB at bit rate of 1.0 bpp which outperforms the other three schemes. The central PSNR of 3DLVQ- A_2 [19] and 3DLVQ- $\mathbb{Z}_2$ reach 40.75 dB and 39.26 dB, respectively, both at bit rate of 1.0 bpp. As higher dimension lattice competent to allocate more neighbors for each lattice point, the reconstruction quality of central decoder is elevated. The central decoder PSNR of the proposed scheme well outperforms the other four schemes.

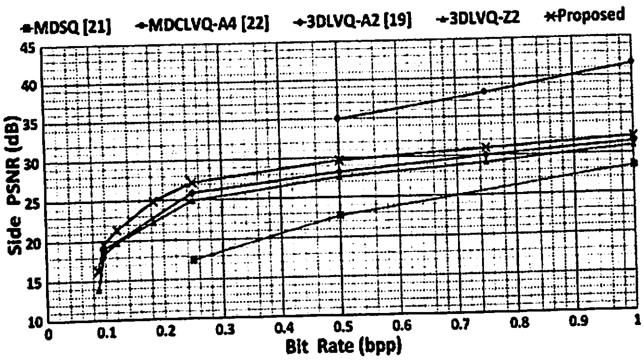

Fig. 7. Comparisons of average side PSNR (dB) at varying bit rate (bpp).

Fig. 7 illustrates the average side reconstruction quality for the proposed 3DLVQ- $\mathbb{Z}_4$ at $N=9$, as compared to the MDSQ [21], MDCLVQ- A_4 [22], the 3DLVQ- A_2 [19] and 3DLVQ- $\mathbb{Z}_2$ scheme at varying bit rate. Notice that the proposed 3DLVQ- $\mathbb{Z}_4$ is having 6 side decoder outputs. Hence, the side reconstruction quality is obtained as the average of the side decoder outputs. The side PSNR for the proposed scheme achieves up to 32.13 dB at bit rate of 1.0 bpp as 3DLVQ- A_2 [19] yields 31.46 dB at the same bit rate. The 3DLVQ- $\mathbb{Z}_2$ and MDCLVQ- A_4 [22] side PSNR achieve 30.91 dB and 41.46 dB at bit rate of 1.0 bpp, respectively. The MDSQ [21] performs the worst which produces around 28.42 dB at bit rate of 1.0 bpp. Nonetheless, MDCLVQ- A_4 [22] manages to perform better than proposed 3DLVQ- $\mathbb{Z}_4$.

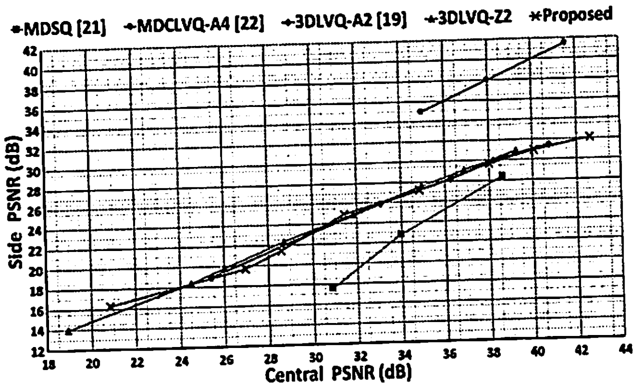

Fig. 8. Performance comparison of Lena for three different schemes with the proposed 3DLVQ- $\mathbb{Z}_4$.

Comparison of the proposed 3DLVQ- $\mathbb{Z}_4$ at $N=9$ with the referenced methods, MDSQ [21], MDCLVQ- A_4 [22], 3DLVQ- A_2 [19] and 3DLVQ- $\mathbb{Z}_2$ in terms of side PSNR versus central PSNR is plotted in Fig. 8. Both of the four lattice based vector quantization schemes are using the three description coding system and the performances are having the same trend. The performance of the proposed 3DLVQ- $\mathbb{Z}_4$ at $N=9$ surpasses three other referenced works. However, MDCLVQ- A_4 [22] accomplishes better than proposed 3DLVQ- $\mathbb{Z}_4$.

The results shown in Fig.7 and Fig. 8 indicate that the MDCLVQ- A_4 [22] which utilizes the coinciding similar sublattice of hexagonal lattice can generates two different descriptions. This allows the scheme to make good approximation for the original lattice point. This means that the descriptions are capable to produce representations that are very near to the source data.

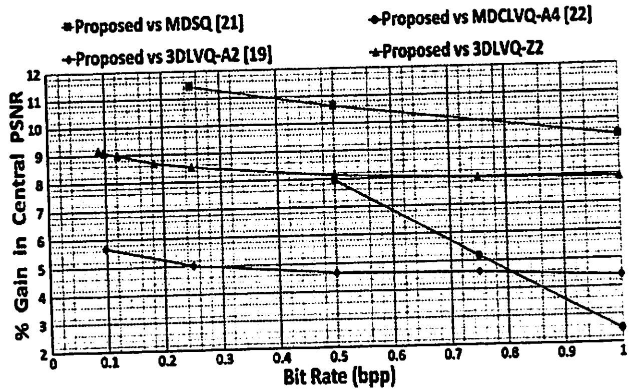

Fig. 9. Percentage gain in central PSNR (dB) for proposed 3DLVQ- $\mathbb{Z}_4$.

Fig. 9 shows the proposed 3DLVQ-$\mathbb{Z}_4$ outpaces about 4.4 % to 5.6 % over the 3DLVQ-A_2 [19]. The proposed 3DLVQ-$\mathbb{Z}_4$ at $N=9$ expands the central reconstruction quality from 7.9 % to 9.2 % over the 3DLVQ-$\mathbb{Z}_2$. Improvement of 9.41 % to 11.43 % is achieved as compared the proposed 3DLVQ-$\mathbb{Z}_4$ over MDSQ [21]. Nevertheless, proposed 3DLVQ-$\mathbb{Z}_4$ is in average 2.49 % to 7.95 % higher than the central reconstruction quality of the MDCLVQ-A_4 [22].

IV. Conclusion

This paper presented a new labeling function of lattice $\mathbb{Z}_4$ at $N=9$ with triple ordered pair $(\lambda_1', \lambda_2', \lambda_3')$ for the three description MDLVQ-$\mathbb{Z}_4$ system. The proposed 3DLVQ-$\mathbb{Z}_4$ scheme outperforms the renowned MDC schemes. The central reconstruction quality is promoted to 42.63 dB and the average side reconstruction quality achieves 32.13 dB, both at bit rate of 1.0 bpp for the proposed 3DLVQ-$\mathbb{Z}_4$ scheme. By applying higher dimension lattice, offers each lattice to have more neighbors, therefore the decoded data via central decoder and side decoder are better quality.

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