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Channel Coding Analysis for High-Speed Telecommunication System

Khin Saw*, Lei Lei Yin Win, Hla Myo Tun, Thanda Win, Mya Mya Aye, Khin Kyu Kyu Win, Devasis Pradhan

***Abstract*—The paper mainly focuses on the channel coding design for high-speed telecommunication systems. The challenging issues in this study are based on (1) the growing demand for high data speed and an increase in subscribers, and (2) high-speed telecommunication networks allow users to avoid them due to better speed and more bandwidth. The objectives of this study are (1) to obtain a higher data rate, higher spectral efficiency, higher throughput, higher bandwidth, and higher energy efficiency at lower latency and (2) to detect/correct errors caused when information is transmitted through noisy channels. Therefore, high-speed telecommunication channel coding techniques will play a major role in achieving fast communication with minimum errors. The linear block and turbo codes are fundamental to analyzing the channel coding scheme for specific purposes. Theoretical concepts with numerical simulation are used to conduct the analyses. The simulation results on BER analyses confirm that the performance criteria could be met with real-world applications.**

***Index Terms*—Channel Coding, High-Speed Telecommunication System, Efficient Coding Techniques, Numerical Analysis, Wireless Communication Channel**

I. Introduction

THE communication system is a means of conveying information from one user to another user. The digital communication system is one in which the data is transmitted digitally. A digital communication system schematic diagram is shown in Fig.1. Source coding is used to remove redundancy from source information for efficient transmission. The transmitted signal power and channel bandwidth are the key parameters in the digital communication system design. The signal energy per bit (Eb) to noise power spectral density (N0) ratio is determined using these parameters. This ratio uniquely assesses the bit error probability, often called bit error rate (BER). In practice, for a fixed Eb/N0, acceptable BER is possible with channel coding. This can be achieved by adding additional digits to the transmitted information stream. These extra digits do not have any new information, but they make it possible for the receiver to detect and correct errors, thereby reducing the overall probability of error [1-6].

The aspiration of channel coding is absolutely different: to protect the message signal in opposition to channel noise. It could be insisted on necessitating for taking channel noise into account to the point of building their survival as its unambiguous property. If the result of that noise is a symbol error probability unable to coexist with the specified restitution quality, the initial message by such coding that it increases transmission security in the presence of noise was proposed to transform. The theory does not even eliminate the extreme case where individual superiority is the total absence of errors. The objectives of source coding and channel coding thus become incompatible. They are even incongruous since source coding increases the vulnerability to errors while improving concision [7-16].

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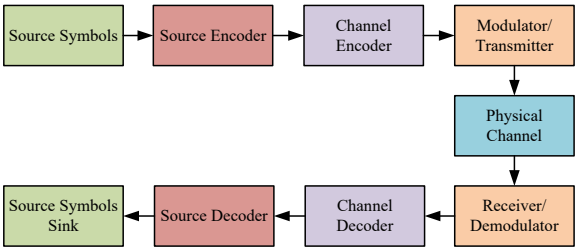


Fig. 1. Digital communication system with coding.

The paper is organized as follows. Section II presents the coded data stream. Section III mentions the linear block codes. Section IV expresses the turbo code. Section V focuses on the implementation and results with discussions. Section VI concludes the study.

II. Coded Data Stream

The data stream is broken into blocks of k bits, and each k-bit block is encoded into a block of n bits with n > k bits as illustrated in Fig.2. The n-bit block of the channel block encoder is called the code word. The code word is formed by adding (n-k) parity check bits derived from the k message bits.

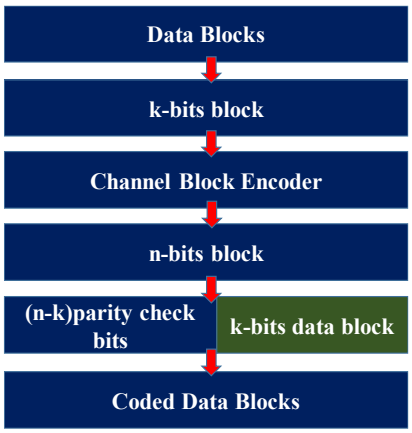


Fig. 2. Flow diagram of code data stream.

III. Linear Block Codes

The two essential properties of linear block codes are as follows.

Property 1: The linear combination of any set of code

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words is a code word.

Property 2: The minimum distance of a linear block code equals the minimum weight of any nonzero word in the code. The two well-known bounds on the minimum distance are

1) Singleton Bound
The minimum distance of a (n, k) linear block code is bounded by

$$d_{min} \tag{1}$$

2) Hamming Bound
An (n, k) block code can correct up to tec errors per code word, provided that n and k satisfy the Hamming bound

$$2^{n-k} \geq \sum_{i=0}^{t_{ec}} \binom{n}{i} \tag{2}$$

The relation is the upper bound on the d_{min} and is known as the Hamming bound.
where

$$\binom{n}{i} = \frac{n!}{(n-1)!i!};\ t_{ec} = (d_{min}()) \tag{3}$$

A. Generator and Parity Check Matrices

Let $\{g_0; g_1; \ldots; g_{k-1}\}$ be a basis of code words for the (n, k) linear block code C and $m = (m_0; m_1; \ldots; m_{k-1})$ the message to be encoded.

It follows that the code word $c = (c_0; c_1; \ldots; c_{n-1})$ for the message is uniquely represented by the following linear combination of $g_0; g_1; \ldots; g_{k-1}$

$$c = m_0 g_0 + \ldots + m_{k-1} g_{k-1} \tag{4}$$

for every code word $c \in C$.

Since every linear combination of the essential elements must also be a code word, there is a one-to-one mapping between the set of k-bit blocks $(a_0; a_1; \ldots; a_{x-1})$ over GF (2) and the code words in C.

A matrix G is constructed by taking the vectors in the basis as its rows.

$$G = \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{k-1} \end{bmatrix} = \geq \begin{bmatrix} g_{0,0} & g_{0,1} & \cdots & g_{0,n-1} \\ g_{1,0} & g_{1,1} & \cdots & g_{1,n-1} \\ \vdots & \vdots & \ddots & \vdots \\ g_{k-1,0} & g_{k-1,1} & \cdots & g_{k-1,n-1} \end{bmatrix} \tag{5}$$

This matrix is a generator matrix for the code C. It can be used to directly encode k-bit blocks in the following manner.

$$mG = (m_0, m_1, \ldots, m_{k-1}) \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{k-1} \end{bmatrix} = m_0 g_0 + m_1 g_1 + \ldots + m_{k-1} g_{k-1} = c \tag{6}$$

The dual space of a linear block code C is the dual code of C, and a basis $\{h_0; h_1; \ldots; h_{n-k-1}\}$ can be found for the dual code of C, and the following parity check matrix can be constructed.

$$H = \begin{bmatrix} h_0 \\ h_1 \\ \vdots \\ h_{n-k-1} \end{bmatrix} = \geq \begin{bmatrix} h_{0,0} & h_{0,1} & \cdots & h_{0,n-1} \\ h_{1,0} & h_{1,1} & \cdots & h_{1,n-1} \\ \vdots & \vdots & \ddots & \vdots \\ h_{n-k-1,0} & h_{n-k-1,1} & \cdots & h_{n-k-1,n-1} \end{bmatrix} \tag{7}$$

In a systematic linear block code, the last k bits of the code word are the message bits, that is,

$$c_i = m_{i-(n-k)}, i = n-k, \ldots, n \tag{8}$$

The last n - k bits in the code word are check bits generated from the k message bits according to

$$\begin{aligned} c_0 &= p_{0,0} m_0 + p_{1,0} m_1 + \cdots + p_{k-1,0} m_{k-1} \\ c_0 &= p_{0,1} m_0 + p_{1,1} m_1 + \cdots + p_{k-1,1} m_{k-1} \\ &\vdots \\ c_{n-k-1} &= p_{0,n-k-1} m_0 + \\ & p_{1,n-k-1} m_1 + \ldots + p_{k-1,n-k-1} m_{k-1} \end{aligned} \tag{9}$$

The above equations can be written in matrix form as

$$[c_0, c_1, \ldots, c_n] = [m_0, m_1, \ldots, m_{k-1}] \times \begin{bmatrix} p_{0,0} & p_{0,1} & \cdots & p_{0,n-k-1} & 1000 & \cdots & \cdots & 0 \\ p_{1,0} & p_{1,1} & \cdots & p_{1,n-k-1} & 0100 & \cdots & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ p_{k-1,0} & p_{k-1,1} & \cdots & p_{k-1,n-k-1} & 0000 & \cdots & \cdots & 1 \end{bmatrix} \tag{10}$$

or

$$c = mG \tag{11}$$

where G is the matrix on the right-hand side of the Above Equation.

The k × n matrix G is called the generator matrix of the code, and it has the form

$$G = [P \ \vdots \ I_k]_{k \times n} \tag{12}$$

The matrix I_k is the identity matrix of order k and p is an arbitrary k by n - k matrix.

When p is specified, it defines the (n,k) block code completely.

The parity check matrix H corresponding to the above generator matrix G can be obtained as

$$H = \begin{bmatrix} 1 & 0 & 0 & 0 \cdots & 0 & p_{0,0} & p_{1,0} & \cdots & p_{k-1,0} \\ 0 & 1 & 0 & 0 \cdots & 0 & p_{0,1} & p_{1,1} & \cdots & p_{k-1,1} \\ \vdots & \vdots & \vdots & \vdots \cdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & 0 & 0 \cdots & 1 & p_{0,n-k-1} & p_{1,n-k-1} & \cdots & p_{k-1,n-k-1} \end{bmatrix}$$
$$H = [I_{n-k} \ \vdots \ P^T] \tag{13}$$

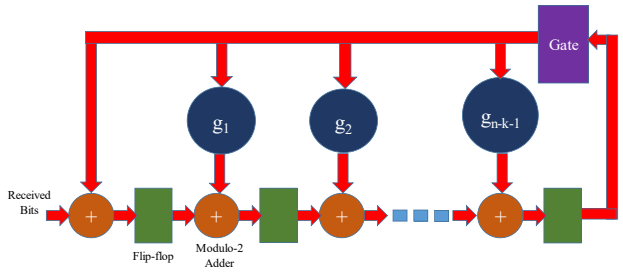


Fig. 3. Block diagram of syndrome calculator.

B. Syndrome Calculator

The syndrome calculator is similar to the encoder. The only difference is that the received bits are fed from the left into the (n – k) stages of the feedback shift register. At the end of the last received bit shifting, the shift register contains the desired syndrome s. If the syndrome is zero, there are no transmission errors in the received word or else the received code word includes a transmission error.

Knowing the value of syndrome, we can determine the corresponding error pattern and make the appropriate correction.

C. *BER Performance of RS Codes for BPSK Modulation in AWGN and Rayleigh Fading Channels*

The redundancy introduced by the RS code increases the channel symbol transmission rate, reducing the received E_b/N_0. For a code with rate R, this is for BPSK in the AWGN channel and Rayleigh fading channel.

$$P = Q\left(\sqrt{2R\frac{E_b}{N_0}}\right)$$

$$P = \frac{1}{2}\left(1 - \sqrt{\frac{R_{\bar{\gamma}}}{1+R_{\bar{\gamma}}}}\right)$$

$$\text{where } R = k/n. \qquad (14)$$

IV. Turbo Code Design

The parallel concatenation of two convolutional codes forms a class of Turbo codes. The turbo encoder structure is formed by parallel concatenating two identical Recursive Systematic Convolutional (RSC) component codes. An interleaver separates these two RSC encoders. The interleaved arranges the data in a predetermined manner. The output of the Turbo encoder is constructed using the systematic output bits of the first RSC encoder and the parity bits of both the RSC encoders. The second RSC encoder outputs systematic bits, which is the permuted version of the first RSC encoder and hence is not considered in the final output.

A. *Non-recursive and Recursive Systematic Convolutional Encoders*

The following rate-1/2 convolutional encoder is an example of a systematic convolutional encoder.

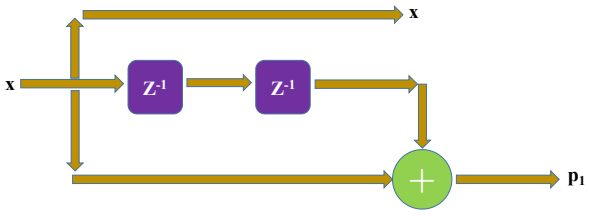


Fig. 4. Non-recursive systematic convolutional (SC) encoder.

B. *Recursive Systematic Convolutional (RSC) Encoder*

The conventional convolutional encoder with rate 1/2 and constraint length 3 is shown in Fig.5.

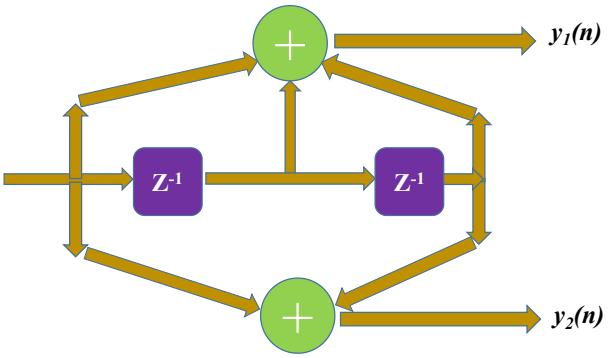


Fig. 5. Non-recursive nonsystematic convolutional (NSC) encoder.

The generator sequences of the above non-recursive nonsystematic encoder are g_1 =[111] and g_2 =[101].

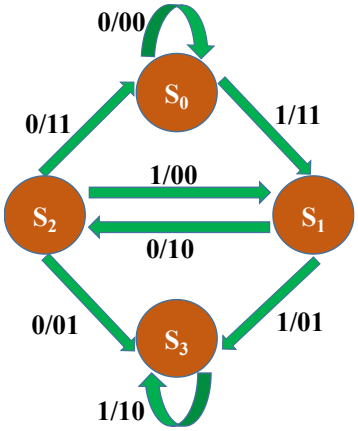


Fig. 6. NSC encoder state diagram.

C. *Equivalent RSC Encoder of Non-recursive Nonsystematic Encoders*

The equivalent RSC encoder of the non-recursive, nonsystematic encoder is shown in Fig.7.

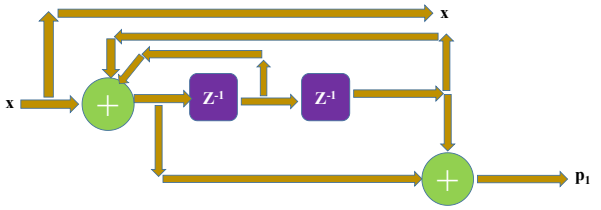


Fig. 7. Recursive systematic convolutional (RSC) encoder.

D. *RSC Encoder State Diagram*

Fig. 8 shows the state diagram of the RSC encoder. It is obtained by feeding back the contents of the memory elements to the input with the generator function.

$$G = \left[1 \quad \frac{g_2}{g_1}\right] = \left[1 \quad \frac{1+D^2}{1+D+D^2}\right] \qquad (15)$$

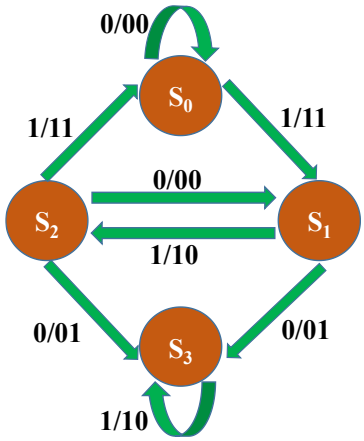


Fig. 8. State diagram of the RSC encoder.

E. Turbo Encoder

A turbo encoder structure consists of two identical RSC encoders in parallel concatenation as shown in Fig.9. It is a rate 1/3 encoder.

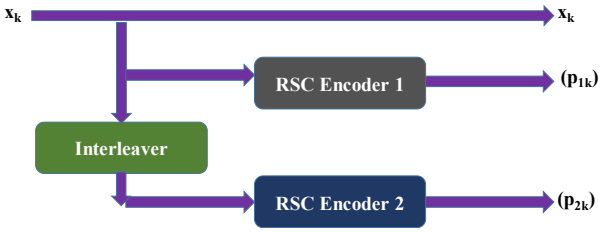


Fig. 9. Turbo encoder.

F. Turbo Coding Illustration

The input data sequence x ={1 1 0 0} assuming the RSC encoder 1 trellis is terminated and the interleaver is {5, 2, 4, 0, 1, 3}.

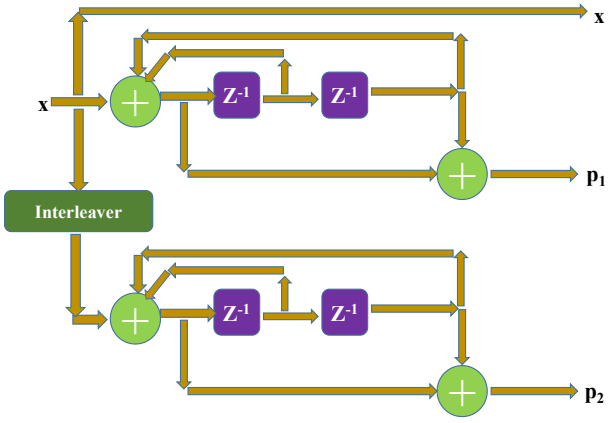


Fig. 10. Illustration of turbo coding.

The two RSC encoders work synergistically in parallel. The RSC encoder 1 takes the data bits x and produces low-weight parity bits (p_{1k}) from them. The RSC encoder 2 gets the data bits x scrambled by an interleaver and computes highweight parity bits (p_{2k}) from the scrambled input bits. Thus, moderate-weight turbo code is generated with combined low-weight code from encoder 1 and high-weight code from encoder 2. Finally, the original input sequence x and the two parity bits strings are transmitted over the channel.

TABLE I
TRANSITIONS FOR TURBO ENCODER

x(n)	State at n	State at n + 1	p1(n)
0	S0(0,0)	S0	0
1		S1	1
0	S1(1,0)	S3	1
1		S2	0
0	S2(0,1)	S1	0
1		S0	1
0	S3(1,1)	S2	1
1		S3	0

G. Implementation

This section shows the step-by-step implementation of the coding process, and Table II shows the encoder results from the implementation. Fig.11 shows the Trellis diagram for the encoder. Fig.12 illustrates the Trellis path corresponding to the input sequence. Fig.13 demonstrates the Pseudo-random interleaver. Fig.14 mentions the Trellis path corresponding to the interleaved input sequence.

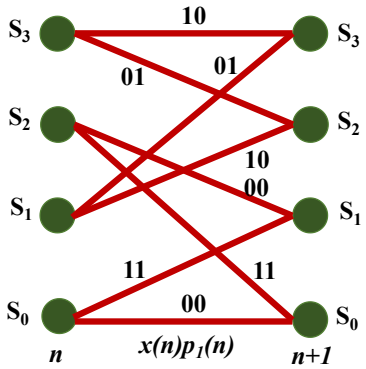


Fig. 11. Trellis diagram for encoder.

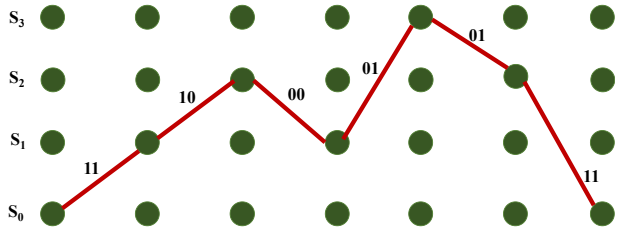


Fig. 12. Trellis path corresponding to the input sequence.

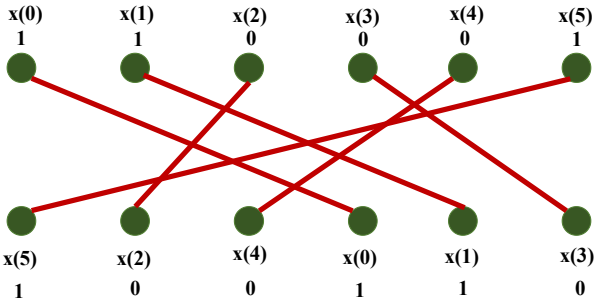


Fig. 13. Pseudo-random interleaver.

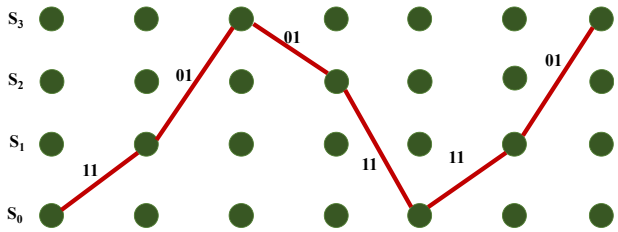


Fig. 14. Trellis path corresponding to interleaved input sequence.

The path started from S_0 to S_0 goes six segments of 11, 10, 00, 01, 01, 11, respectively in Fig.12. That trajectory is the input sequence for channel coding techniques in reality S_0 to S_0 via S_1, and S_2. The path started from S_0 to S_3 goes six segments of 11, 01, 01, 11, 11, 01, respectively in Fig.14. That trajectory is the optimal solution of respective interleaved input sequence for channel coding techniques from S_0 to S_3 in reality.

TABLE II
OUTPUT OF ENCODER RESULTS

x(n)	p1(n)	p2(n)		x(n)	p1(n)	p2(n)
1	1	1	➔	1	1	1
1	0	1	➔	1	-1	1
0	0	1	➔	-1	-1	1
0	1	1	➔	-1	1	1
0	1	1	➔	-1	1	1
1	1	1	➔	1	1	1

H. Mathematical Model for BER Analysis

$$L_R = ln\left(\frac{P[m_k=1|\hat{y}]}{P[m_k=0|\hat{y}]}\right) \quad (16)$$

The total LLR after any iteration is calculated as

$$L_{ul}(\text{total}) = L_{ul}(\text{channel}) + L^{e(2\to1)}_{u(\text{Deint})(l)} + L^{e(1\to2)}_{ul} \quad (17)$$

where L_{ul}(channel)is the channel LLR.

The extrinsic information sent from SISO Decoder 1 to 2 is $Lul\ e(1\to2)$. The deinterleaved position of ul is indicated by $Deint(l)$.

V. RESULTS AND DISCUSSIONS

Fig. 15 shows the decoding error probability for RS codes using coherent BPSK over an AWGN channel. The probability approach decreases as the E_b/N_0 and code dimension decrease in this analysis. This can be attributed to the highly imperfect nature of RS codes.

Decoding error probability for (127,63) RS codes using coherent BPSK over an AWGN channel and Rayleigh fading channel is shown in Fig.16. From Fig.16, it is seen that the coded AWGN and Rayleigh fading channels exhibit much better BER performance than the uncoded AWGN and Rayleigh fading channels.

Fig. 17 shows the decoding error probability for (127,63) RS codes using non-coherent BPSK. It shows that the coded AWGN and Rayleigh fading channels exhibit much better BER performance than the uncoded AWGN and Rayleigh fading channels. However, the performance is not better than that of BPSK modulation.

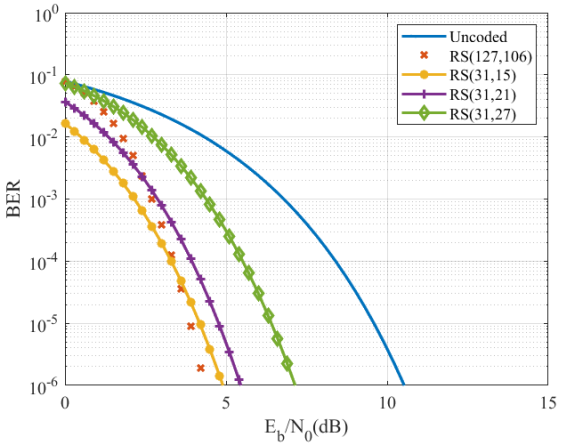


Fig. 15. BER vs E_b/N_0 with AWGN channel RS(127,106).

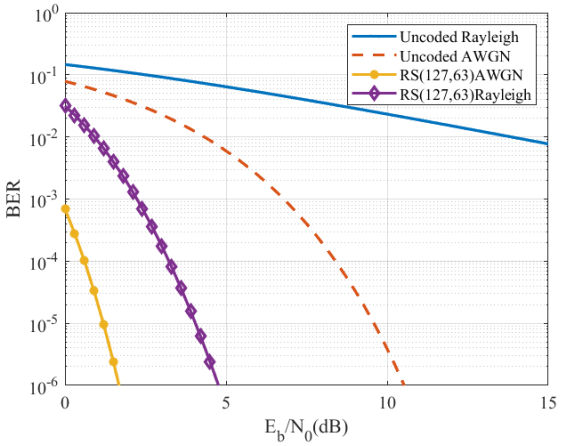


Fig. 16. BER vs Eb/N0 with AWGN channel RS(127,63).

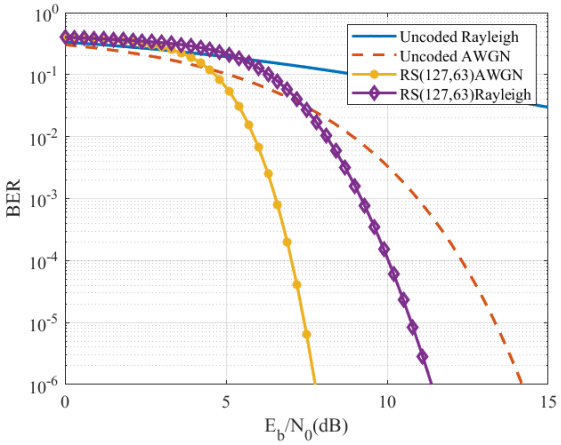


Fig .17. BER vs Eb/N0 with AWGN channel RS(127,63) [non-coherent condition].

In Maximum Ratio Combining (MRC), the receiver weights the incoming signals on antennas by the conjugates of the complex fading random variables—the pair-wise bit error probability with MRC in a Rayleigh

fading channel for the BPSK case.

The upper bound on the BER performance of turbo codes in the Rayleigh fading channel with MRC diversity can be calculated. Based on numerical analysis, the plot of Eb/N0 versus upper bound BER Performance of the turbo encoder with an interleaved length of 100 was implemented. Turbo coding approaches could affect channel modelling in high-speed telecommunication. Fig.18 demonstrates the upper bound BER performance of turbo encoder with interleaver length of 100.

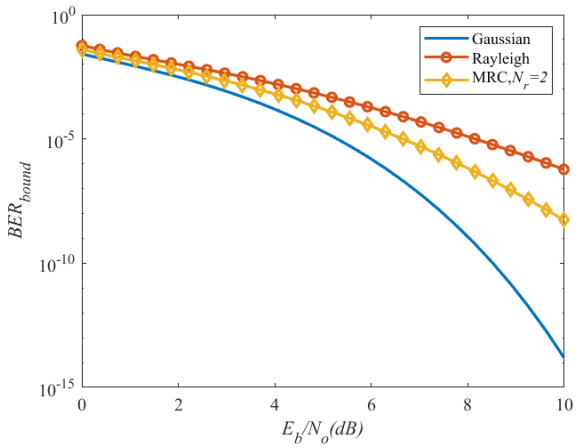


Fig. 18. Upper bound BER performance of turbo encoder with interleaver length of 100.

The coding method in this study is very suitable for high performance telecommunication channel model like 5G and beyond. The coding methods specifically could contribute to the unique challenges of 5G systems (e.g., massive MIMO, ultra-low latency) because of the performance achievements with several analyses in this study. Table III of statistic table gives the performance comparison between LDPC (Low-Density Parity Check) codes or Polar codes and this study.

VII. Conclusion

5G channel coding techniques will play a major role in achieving fast communication with minimum errors. Coding methods and modulation techniques are essential in 5G. The best modulation techniques, QPSK and QAM, were analyzed. The BER results obtained using QPSK for communication over AWGN, E_b/N_0 show that the Polar code achieves superior BER performance compared to the LDPC code. MIMO is used in 5G because MIMO is a vital solution for wireless communication. 5G channel coding techniques will play a major role in achieving fast communication with minimum errors. Coding methods and modulation techniques are essential in 5G. The best modulation techniques, QPSK and QAM, were analyzed. The BER results obtained using QPSK for communication over AWGN, E_b/N_0 show that the Polar code achieves superior BER performance compared to the LDPC code. MIMO is used in 5G because MIMO is a vital solution for wireless communication. According to the constraint of hardware implementation, the numerical analyses on the performance measurement could be extended to future works for next research works. But the results say that the coding schemes would evolve with the development of 6G networks or new wireless technologies after modifying the limitations of the study in this work.

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TABLE III
Statistic Table

Name of Channel Coding Techniques	Strengths	Weaknesses	Applications
Linear Block Code	Convolutional codes, a specific type of linear block code, offer low latency and flexibility.	Performance can be limited compared to LDPC and Turbo codes in some scenarios.	General error correction, especially where low latency is crucial.
Turbo Code	Good performance at moderate code rates, particularly with iterative decoding.	Higher decoding complexity and latency compared to LDPC and polar codes.	Older generations of mobile communication systems, some high-reliability scenarios.
Polar Code	Low encoding and decoding complexity, reach channel capacity, and simple implementation.	Performance can be slightly lower than LDPC codes for longer block lengths.	5G NR control channels, scenarios with short block lengths.
LDPC	Near-Shannon-limit performance, good error floor, and decoding complexity that decreases with increasing code rate.	Higher encoding complexity compared to polar codes.	5G NR data channels, high-throughput systems.

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