Diversity-Enhancing Schemes with Asymmetric Diversity Modulation in Wireless Fading Relay Channels

Jaemin Han

Department of Electrical and Electronic Engineering

The Graduate School

Yonsei University

Diversity-Enhancing Schemes with Asymmetric Diversity Modulation in Wireless Fading Relay Channels

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of Jaemin Han is approved.

(Thesis Supervisor: Daesik Hong)

(Hongyeop Song : Thesis Committee Member #1)

(Seunglyun Kim : Thesis Committee Member #2)

The Graduate School

Yonsei University

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한 재 민 드림

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Abstract

Diversity-Enhancing Schemes with Asymmetric Diversity Modulation in Wireless Fading Relay Channels

In this thesis, we propose an asymmetric diversity modulation (ADM) scheme for a single-source relay system that utilizes the relay's higher transmission ability as a form of diversity. To achieve this, the proposed method transmits multiple source bits over a high-order modulating relay as a way to provide additional time diversity. The spatial and time diversity then undergo 'bit'-based combining at the destination. Using the proposed 'bit'-based channel combining method, we derive the theoretical bit error rate (BER) for such a system. Moreover, we investigate the fact that the proposed scheme shows a performance trade-off between bit power and time diversity resulting from the reduced bit power caused by a high-order modulating relay.

In addition to this asymmetric diversity modulation (ADM) scheme, we further propose a way to enhance diversity order without wasting additional resources by XORing at the relay. An XOR-Relay offers an increased number of diversity paths, enabling the destination to construct a cycle-free decoding structure, and resulting in an enhanced diversity order. Combined with the proposed 'bit'-based channel combining method, we also derive its theoretical bit error rate (BER) and show that the proposed way approaches the increased diversity order with a few iterations at the destination.

Keywords : Wireless fading relay channel, diversity, modulation, bit error probability, iterative MAP decoding.

Chapter 1

Introduction

SIGNALS transmitted via wireless environments suffer from severe attenuation and random fluctuation in amplitude known as fading. Recently, there has been an upsurge of interest in the use of relaying in cellular-based infrastructure as a practical solution to mitigate fading and improve reliability [1]. The major advantage of relaying is that independent transmission paths to the destination can be provided with low deployment costs. Recent proposals have shown that this spatial diversity can be realized in a distributed fashion [2]-[3].

In addition to the potential of increased diversity, relaying can offer improved link qualities thanks to decreased propagation loss. This means that higher transmission rates can be supported over relays using a spectrally efficient modulation scheme. However, the transmission rate used by a relaying node cannot exceed the rate generated at the source, even though the relay link quality may support higher rates. For a single-source network, however, this very fact can be exploited to obtain additional performance enhancement.

Based on this concept, we propose an asymmetric diversity modulation (ADM) scheme that utilizes the relay's higher transmission ability as a form of diversity. The proposed scheme is considered in a three-node fading network, called a wireless fading relay channel, consisting of a source, a relay, and a destination, as shown in Fig. 2.1. It is known that second-order diversity can be achieved through two spatially independent links, i.e., source-destination and relay-destination [5]. In addition to this spatial diversity, the proposed scheme additionally provides time diversity through the relay by transmitting multiple repeated source bits using high-order modulation. With this approach, each source bit will experience increased fading on the relay-destination link depending on the order of modulation employed at the relay. Since high-order modulation reduces bit power, the proposed method shows a bit-level performance trade-off between bit power and time diversity in the relay-destination link. Using the proposed 'bit'-based channel combining method, we derive its theoretical bit error rate (BER) performance and study this trade-off relationship. Moreover, in the relay-destination link, we compare the average BER performance with the existing signal space diversity (SSD) techniques [4].

In addition to this asymmetric diversity modulation (ADM) scheme, we further propose a simple encoding method that XOR the consecutive bits at the relay. XORing allows the relay to offer each bit the increased diversity paths without wasting additional resources, even though it is XORed with other bits. And the XORed redundancies of the relay-destination link can be utilized to construct an iterative maximum a posteriori (MAP) decoding structure at the destination with the transmitted systematics from the source-destination link. Although the diversity performance is rather limited compared to the high-complex Forward Error Correcting (FEC) codes' redundancies such as an Low Density Parity Check (LDPC) code, the proposed XOR-Relay scheme efficiently increases time diversity with a few iterations at the destination thanks to the cycle-free decoding structure and we demonstrate this via simulation results. Furthermore, using the proposed 'bit'-based channel combining method, we derive theoretical bit error rate (BER) of the proposed scheme in combination with the asymmetric diversity modulation (ADM) scheme and conform its increased diversity order.

Chapter 2

Proposed Asymmetric Diversity Modulation (ADM) Scheme

IN this chapter, the proposed asymmetric diversity modulation (ADM) scheme and transmission signal model are described in the single-source relaying topology. In order to reflect the practical channel environment, the proposed scheme is considered in the block fading channel model with the proper assumptions. Section 2.1 introduces the asymmetric diversity modulation (ADM) scheme and the general demodulation process with the Maximum Likelihood (ML) criterion is explained. Section 2.2 discusses the problem in deriving the theoretical bit error rate (BER) when using the conventional hypothesis test of bit log-likelihood ratio (LLR) from the Maximum Likelihood (ML) criterion.

2.1 The system and transmission model of the proposed ADM scheme

Let us define source, relay, and destination as $\mathcal{S}$, $\mathcal{R}$ and $\mathcal{D}$, respectively. Consider a time-divided transmission of $\mathcal{S}$ and $\mathcal{R}$ to $\mathcal{D}$ as shown in Fig. 2.1. If the block of binary source bits constitutes one packet for transmission, $\mathcal{S}$ broadcasts its packet to $\mathcal{R}$ and $\mathcal{D}$ in the first time slot, and $\mathcal{R}$ decodes and forwards (DF) to $\mathcal{D}$ in the second time slot.

This single-source relaying topology can achieve second-order spatial diversity. However, the unreliability of the $\mathcal{S} \rightarrow \mathcal{R}$ link is a critical bottleneck which could degrade diversity performance down to the first order [5]. Hence, it is necessary to assume a perfect $\mathcal{S} \rightarrow \mathcal{R}$ link for the relay to transmit without errors. We also assume that the $\mathcal{S} \rightarrow \mathcal{D}$ and $\mathcal{R} \rightarrow \mathcal{D}$ links are independent Rayleigh fading channels with mutually independent white Gaussian noise [6].

In the proposed ADM scheme, $\mathcal{R}$ duplicates the received packet to the order of modulation employed and rearranges them. For the case shown in Fig. 2.1 using a 4^{th}-order modulating relay, the 4 bits A, B, C, and D are repeated and rearranged to make up the relay packet while being relocated to a different bit position at a different symbol time. Thus, each bit is transmitted as part of a set of 4^{th}-order modulation symbols ($x_{\mathcal{R}_1}$, $x_{\mathcal{R}_2}$, $x_{\mathcal{R}_3}$, and $x_{\mathcal{R}_4}$) in the relay packet. In order to provide additional time diversity for a specific bit, every symbol conveying that bit must go through the uncorrelated channel fading. That is, in Fig. 2.1, $x_{\mathcal{R}_1}$, $x_{\mathcal{R}_2}$, $x_{\mathcal{R}_3}$, and

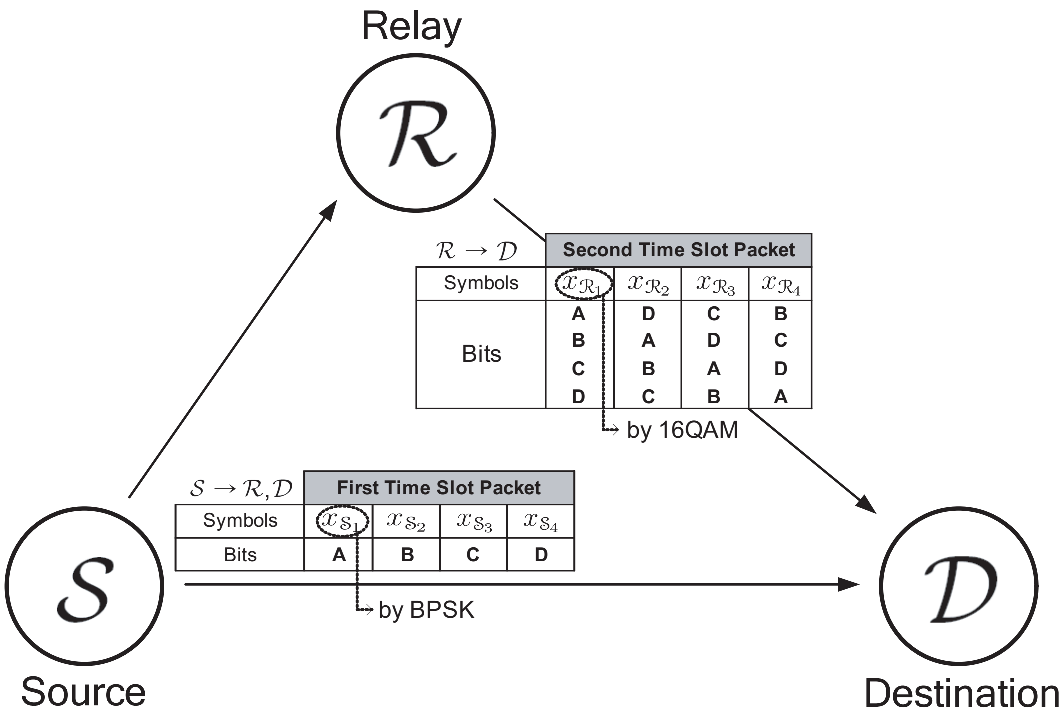

Figure 2.1: Proposed ADM scheme in a wireless fading channel with one relay. The source is transmitting its packet by BPSK. In order to achieve additional 4^{th}-order time diversity in the $\mathcal{R}\to\mathcal{D}$ link, the relay repeats and rearranges the packet and transmits it using a 4^{th}-order modulation, such as 16QAM.

$x_{\mathcal{R}_4}$ need to be faded independently in order for bits A, B, C, and D to have 4 additional time diversities.

In practical environments, however, the coherence time of typical wireless channels is on the order of tens to hundreds of symbols, so that the channel varies slowly relative to the symbol rate [7]. Thus, we assume that each packet is composed of several hundreds of symbols so as to give several independent fading gains. Therefore, in the proposed ADM scheme, if the order of modulation employed at the relay is less than or equal to the number of independent fading gains, $\mathcal{R}$ can successfully lead its bits to have additional time diversity by arranging the same bits sufficiently far apart in time.

With these assumptions, symbols in the received packets at the destination from the time-divided transmissions of the source packet and the relay packet can be written as

$$y_{\mathcal{S}_i} = h_{\mathcal{S}\mathcal{D}_i}\, x_{\mathcal{S}_i} + W \quad \text{and} \quad y_{\mathcal{R}_j} = h_{\mathcal{R}\mathcal{D}_j}\, x_{\mathcal{R}_j} + W, \tag{2.1}$$

where W is the destination complex white Gaussian noise (WGN) with variance N, and index i and j denote the i^{th} and j^{th} symbol position in each packet. Thus, $h_{\mathcal{S}\mathcal{D}_i}$, $h_{\mathcal{R}\mathcal{D}_j}$ are Rayleigh fading variables from the $\mathcal{S}\to\mathcal{D}$ and $\mathcal{R}\to\mathcal{D}$ links, where the i^{th} symbol $x_{\mathcal{S}_i}$ of the source packet and j^{th} symbol $x_{\mathcal{R}_j}$ of the relay packet have gone through, respectively. We fix each Rayleigh fading power to be equal to $2\sigma^2$. However, we set the power constraints for each transmitting node, $\mathbb{E}[x_{\mathcal{S}}] = P_{\mathcal{S}}$ and $\mathbb{E}[x_{\mathcal{R}}] = P_{\mathcal{R}}$, so that we can reflect the relay's smaller propagation loss by controlling their transmission powers.

Since the destination combines diversity performance based on 'bit', each bit in the received packets must be expressed as a soft value to be combined before decision. By applying Maximum Likelihood (ML) criterion with coherent demodulation, we can obtain a measurement for a received bit, i.e., a bit log-likelihood ratio (LLR). Thus, in the proposed scheme, a bit decision variable is expressed as combining the bit LLRs corresponding to the same bit. In the general situation where the source is using $K_{\mathcal{S}}^{\text{th}}$-order modulation and the relay is using $K_{\mathcal{R}}^{\text{th}}$-order modulation, total combined LLR for a specific bit A at the destination from the received symbols of (2.1) is

$$Ł(\mathrm{A}) = \log \frac{\prod_{k=1}^{K_{\mathcal{S}}} P(y_{\mathcal{S}_{i(k)}}|b_k=0) \prod_{k=1}^{K_{\mathcal{R}}} P(y_{\mathcal{R}_{j(k)}}|b_k=0)}{\prod_{k=1}^{K_{\mathcal{S}}} P(y_{\mathcal{S}_{i(k)}}|b_k=1) \prod_{k=1}^{K_{\mathcal{R}}} P(y_{\mathcal{R}_{j(k)}}|b_k=1)} = \sum_{k=1}^{K_{\mathcal{S}}} Ł_{x_{\mathcal{S}_{i(k)}}}(b_k) + \sum_{k=1}^{K_{\mathcal{R}}} Ł_{x_{\mathcal{R}_{j(k)}}}(b_k) \,, \tag{2.2}$$

where $Ł_{x_{\mathcal{S}_i}}(b_k) = \log \frac{\sum_{Q \in \{x_{\mathcal{S}_i}:\, b_k=0\}} \exp(-\frac{(y_{\mathcal{S}_i} - h_{\mathcal{SD}_i} Q)^2}{N})}{\sum_{Q \in \{x_{\mathcal{S}_i}:\, b_k=1\}} \exp(-\frac{(y_{\mathcal{S}_i} - h_{\mathcal{SD}_i} Q)^2}{N})}$ and $Ł_{x_{\mathcal{R}_j}}(b_k) = \log \frac{\sum_{Q \in \{x_{\mathcal{R}_j}:\, b_k=0\}} \exp(-\frac{(y_{\mathcal{R}_j} - h_{\mathcal{RD}_j} Q)^2}{N})}{\sum_{Q \in \{x_{\mathcal{R}_j}:\, b_k=1\}} \exp(-\frac{(y_{\mathcal{R}_j} - h_{\mathcal{RD}_j} Q)^2}{N})}$ are bit LLRs [8] each from the source and the relay. $i(k)$ and $j(k)$ are indices indicating symbol position in each packet where the k^{th} bit, b_k, of the symbols $x_{\mathcal{S}_{i(k)}}$ and $x_{\mathcal{R}_{j(k)}}$ is bit A, respectively.

Let us assume that binary phase shift keying (BPSK) is used for source transmission ($K_{\mathcal{S}}=1$), so that we can focus on the trade-off between bit power and diversity for the high-order modulating relay, as shown in Fig. 2.1. The destination then combines the LLR of bit A from the symbol $x_{\mathcal{S}_1}$ in the source packet with the corresponding LLRs of bit A from the symbols $x_{\mathcal{R}_1}$, $x_{\mathcal{R}_2}$, $x_{\mathcal{R}_3}$, and $x_{\mathcal{R}_4}$ in the relay packet. Finally, the decision of whether to demodulate the bit A to zero or one is made as $\L(\mathrm{A}) \underset{1}{\overset{0}{\gtrless}} 0$. However, the non-linear and complicated forms of the bit LLR (2.2) make it difficult to derive the exact BER and to show its trade-off relationship.

2.2 The problem of the conventional hypothesis test in deriving the theoretical BER

In order to derive the exact BER performance of the proposed scheme with the bit decision variable, we first need to determine the decision boundaries of the received symbols for a specific bit to be zero or one. Due to the symmetry of the scheme, without loss of generality we focus on a bit A in calculating the BER. Then, in the case of $K_{\mathcal{S}} = 1$ and $K_{\mathcal{R}} = 4$, the LLR of bit A is expressed as

$$\begin{aligned} \L(\mathrm{A}) &= \L_{x_{\mathcal{S}_1}}(\mathrm{A}) + \L_{x_{\mathcal{R}_1}}(\mathrm{A}) + \L_{x_{\mathcal{R}_2}}(\mathrm{A}) + \L_{x_{\mathcal{R}_3}}(\mathrm{A}) + \L_{x_{\mathcal{R}_4}}(\mathrm{A}) \\ &= \log \frac{\exp\left(-\frac{(y_{\mathcal{S}_1} + h_{\mathcal{SD}_1})^2}{N}\right)}{\exp\left(-\frac{(y_{\mathcal{S}_1} - h_{\mathcal{SD}_1})^2}{N}\right)} + \sum_{k=1}^{4} \log \frac{\sum\limits_{Q \in \{x_{\mathcal{R}_k} :\, b_k = 0\}} \exp\left(-\frac{(y_{\mathcal{R}_k} - h_{\mathcal{RD}_k} Q)^2}{N}\right)}{\sum\limits_{Q \in \{x_{\mathcal{R}_k} :\, b_k = 1\}} \exp\left(-\frac{(y_{\mathcal{R}_k} - h_{\mathcal{RD}_k} Q)^2}{N}\right)}\,, \end{aligned} \tag{2.3}$$

where b_k represents the k^{th} bit of a transmitted symbol. Thus, Q indicates the elements of equally likely constellation points of a symbol corresponding to when the k^{th} bit is 0 or 1, respectively.

$\L(\mathrm{A})$ is now a function of received symbols $(y_{\mathcal{S}_1}, y_{\mathcal{R}_1}, y_{\mathcal{R}_2}, y_{\mathcal{R}_3}, y_{\mathcal{R}_4})$ given the channel fadings $(h_{\mathcal{SD}_1}, h_{\mathcal{RD}_1}, h_{\mathcal{RD}_2}, h_{\mathcal{RD}_3}, h_{\mathcal{RD}_4})$. The hypothesis test of (2.3), $\L(\mathrm{A}) \underset{H_1}{\overset{H_0}{\gtrless}} 0$, determines the decision boundaries of the received symbols $(y_{\mathcal{S}_1}, y_{\mathcal{R}_1}, y_{\mathcal{R}_2}, y_{\mathcal{R}_3}, y_{\mathcal{R}_4})$ for bit A to be demodulated as zero or one. The joint areas of the received symbols which always lead (2.3) to be H_1 are,

$$JA_{H_1}\ (\text{Joint Areas for } H_1) = \{\ y_{\mathcal{S}_1}, y_{\mathcal{R}_1}, y_{\mathcal{R}_2}, y_{\mathcal{R}_3}, y_{\mathcal{R}_4}\ \mid\ \text{Ł}(\text{A}) < 0\ \}\ , \tag{2.4}$$

Then, we can compute the average BER using the above decision boundaries and the probability of incorrect hypothesis H_1 when the zero bit was transmitted. That is,

$$P_{BER,\,ADM} = \mathbb{E}_{h_{\mathcal{SD}_1},\,h_{\mathcal{RD}_1},\,h_{\mathcal{RD}_2},\,h_{\mathcal{RD}_3},\,h_{\mathcal{RD}_4}}\Big[\ P(H_1\,|\,\text{A}\!=\!0)\ \Big]\ , \tag{2.5}$$

where $P(H_1\,|\,\text{A}\!=\!0) = \int\!\cdots\!\int_{JA_{H_1}} P(y_{\mathcal{S}_1}|\text{A}\!=\!0)\prod_{k=1}^{4} P(y_{\mathcal{R}_k}|\text{A}\!=\!0)\, dy_{\mathcal{S}_1} dy_{\mathcal{R}_1} dy_{\mathcal{R}_2}\ dy_{\mathcal{R}_3} dy_{\mathcal{R}_4}$. It is absolutely too complicated to solve the exact BER of the proposed scheme from the bit decision variable of the ML criterion.

Therefore, in order to derive the exact BER in the situation of asymmetric diversity modulation (bits are transmitted through different modulation symbols by the source and the relay), we need another approach. This proposed approach is called the 'bit-based channel combining method', which is explained in great detail in the next chapter.

Chapter 3

Performance Analysis and Trade-off of the proposed ADM scheme

In this chapter, the proposed asymmetric diversity modulation (ADM) scheme is analyzed through the proposed 'bit'-based channel combining method. Since it is nearly impossible to derive the theoretical bit error rate (BER) when directly computing the decision variable of (2.2), we propose another simpler approach to derive the theoretical bit error rate (BER). Section 3.1 describes the proposed 'bit'-based channel combining method. In Section 3.2, the bit-based channel fading effects are derived depending on the order of modulation employed at each diversity paths. Using the derived channel fading effects in Section 3.2, we give the achieved diversity order of the upper-bounded average bit error rate (BER) in Section 3.3. Section 3.4 presents the closed form bit error rate (BER) of the proposed asymmetric diversity modulation (ADM) scheme. Furthermore, we analyze the performance trade-off between bit power and time diversity with the proposed 'bit'-based channel probability density function (pdf) in Section 3.5. Finally, Section 3.6 shows the simulation results and Section 3.7 compares the proposed scheme with the existing signal space diversity (SSD) techniques.

We begin by establishing the notation to be used throughout. We define the transmission power from the source and relay to the destination noise ratio as $\mathtt{SNR}_{\mathcal{S}} = \frac{P_{\mathcal{S}}}{N}$ and $\mathtt{SNR}_{\mathcal{R}} = \frac{P_{\mathcal{R}}}{N}$, respectively. We also denote G$=\frac{\mathtt{SNR}_{\mathcal{R}}}{\mathtt{SNR}_{\mathcal{S}}}$, in order to represent the power gain from the smaller propagation loss in the relay. The pdf of chi-square random variable z with n degree of freedoms is given by

$$p_{\mathrm{z}}(z) = \frac{z^{\frac{n}{2}-1}\exp(-\frac{z}{2\sigma^2})}{(2\sigma^2)^{\frac{n}{2}}\Gamma(\frac{n}{2})} \ , \tag{3.1}$$

where σ^2 is a component variance. For convenience, we will let z $= \chi_n^2(\sigma^2)$. The random variable of Rayleigh fading with its expectation $2\sigma^2$ can then be represented as $\chi_2^2(\sigma^2)$.

3.1 Proposed Bit-based Channel Combining Method

In the proposed scheme, it is a 'bit' that exploits channel diversity by combining spatial and time diversity in the form of a bit LLR. This means that the same bit experiences several independent channel fadings through different modulation symbols in each diversity path. Considering the fact that it is a modulation symbol (conveying bits) that undergoes the fading channel, and not the bit itself, if we can calculate the effect of channel fading directly on a given bit, this will make it easier to perform a 'bit'-based analysis to derive the average BER.

First, let us define a notation for the concept described above. If a bit A is part of a symbol through a channel fading h, $\left(|h|^2\right)_{\mathrm{A}}$ denotes a random variable of the effects of channel fading directly on bit A. With this notation, we can easily combine the effects of channel fadings directly on a specific bit delivered by different modulation symbols. Since the source is using BPSK and the relay is using $K_{\mathcal{R}}^{\mathrm{th}}$-order modulation, a random variable of the combined channel fading effects on a received bit at the destination, $\mathrm{z}_{\mathcal{D}}$, can be defined as (for a specific bit A):

$$\mathrm{z}_{\mathcal{D}} = \mathrm{z}_{\mathcal{SD}} + \mathrm{z}_{\mathcal{RD}} = \left(|h_{\mathcal{SD}_i}|^2\right)_{\mathrm{A}} + \sum_{k=1}^{K_{\mathcal{R}}} \left(|\, h_{\mathcal{RD}_{j(k)}}|^2\right)_{\mathrm{A}}, \tag{3.2}$$

where $h_{\mathcal{SD}_i}$, $h_{\mathcal{RD}_{j(k)}}$ are channel fadings of the i^{th} and $j(k)^{\mathrm{th}}$ symbol position on the $\mathcal{S}{\rightarrow}\mathcal{D}$ and $\mathcal{R}{\rightarrow}\mathcal{D}$ links where the corresponding bit A has been delivered. For the case shown in Fig. 2.1, $\mathrm{z}_{\mathcal{D}} = \left(|h_{\mathcal{SD}_1}|^2\right)_{\mathrm{A}} + \sum_{j=1}^{4} \left(|h_{\mathcal{RD}_j}|^2\right)_{\mathrm{A}}$. Since all bits experience the same fading scenario, there is no difference in $\mathrm{z}_{\mathcal{D}}$ for any specific bit.

In general, BER analysis over fading has a form that averages the channel fading on the conditional (fading-dependent) bit error probability (BEP). For example, the average BER for BPSK over fading with coherent demodulation is [9]

$$P_{BER,\,BPSK} = \int_0^{\infty} Q\Big(\sqrt{2\,z\,SNR}\,\Big)p_{\mathrm{z}}(z)\,dz\,, \tag{3.3}$$

where z indicates the random variable of the channel fading with a pdf, $p_{\mathrm{z}}(z)$.

Since the proposed ADM scheme employs a different modulation order on each diversity path, the conditional BEPs from the source and relay are different. Performing bit-based channel combining for the BER requires a bit-level BEP that can serve as the basis for the combining. We regard the BEP of BPSK as a good candidate, because one BPSK symbol means exactly one bit. In other words, $Q\big(\sqrt{2zSNR}\,\big)$ is the basis for combining the bit-based channel fading effects that the corresponding bit has undergone. If we take $\mathtt{SNR}_{\mathcal{S}}$ as our performance measure, the average BER over the combined bit-based channel fading effects can be solved in a manner similar to equation (3.3) as

$$P_{BER,\,ADM} = \int_0^{\infty} Q\Big(\sqrt{2\,z_{\mathcal{D}}\,\mathtt{SNR}_{\mathcal{S}}}\,\Big)p_{\,\mathrm{z}_{\mathcal{D}}}(z_{\mathcal{D}})\,dz_{\mathcal{D}}\,, \tag{3.4}$$

with the random variable $\mathrm{z}_{\mathcal{D}}$ of the combined bit-based channel fading effects in (3.2).

3.2 Derivation of the Bit-based Channel Fading Effects

Since we assume that the source transmission is done by BPSK, the channel fading effect from the $\mathcal{S}\!\rightarrow\!\mathcal{D}$ link to bit A ($\mathrm{z}_{\mathcal{S}\mathcal{D}}$ in (3.2)) is the same as a random variable of the channel fading, i.e., $\Big(|h_{\mathcal{S}\mathcal{D}_1}|^2\Big)_{\mathrm{A}} = \chi_2^2(\sigma^2)$.

For a relay employing M-ary quadrature amplitude modulation (M-QAM) (M$=2^{K_{\mathcal{R}}}$), the bit-based channel fading effects can be derived from the conditional BEP of M-QAM. In general, the k^{th} bit BEP of any real or imaginary axis using gray-coded M-QAM over WGN is expressed as $\frac{2^k}{\sqrt{M}}Q\Big(\sqrt{\frac{3SNR}{M-1}}\Big)$ for high SNRs [10]. Over fading,

$$P_{BER,\ M\text{-}QAM}(k) = \int_0^\infty \frac{2^k}{\sqrt{M}} Q\left(\sqrt{\frac{3z\mathtt{SNR}_{\mathcal{R}}}{M-1}}\right) p_{\mathrm{z}}(z)\, dz\,, \tag{3.5}$$

where z indicates a random variable of the channel fading with a pdf, $p_{\mathrm{z}}(z)$.

By changing variables, (3.5) can be expressed equivalently as

$$P_{BER,\ M\text{-}QAM}(k) = \frac{2^k}{\sqrt{M}} \int_0^\infty Q(\sqrt{2z'\mathtt{SNR}_{\mathbb{S}}}) \left\{\frac{2(M{-}1)}{3\mathrm{G}} p_{\mathrm{z}}\Big(\frac{2(M{-}1)}{3\mathrm{G}} z'\Big)\right\} dz'. \tag{3.6}$$

Hence, the effect of the channel fading on the k^{th} bit of any axis in an M-QAM symbol is now solved as shown above into the changed pdf $\left\{\frac{2(M-1)}{3\mathrm{G}} p_{\mathrm{z}}\Big(\frac{2(M-1)}{3\mathrm{G}} z\Big)\right\}$ from its original pdf $p_{\mathrm{z}}(z)$ of the channel fading. Any other linear modulation scheme can be employed at the relay, and the pdf of the bit-based channel fading effects can also be derived using the above approach.

In the case of Rayleigh fading, that is when z $= \chi_2^2(\sigma^2)$, the changed pdf is

$$\left\{\frac{2(M{-}1)}{3\mathrm{G}} p_{\mathrm{z}}\Big(\frac{2(M{-}1)}{3\mathrm{G}} z\Big)\right\} = \frac{\exp\Big(-z/\frac{3\sigma^2\mathrm{G}}{(M-1)}\Big)}{\frac{3\sigma^2\mathrm{G}}{(M-1)}}\quad , \tag{3.7}$$

and (3.7) is the same as a pdf of two degrees of freedom chi-square random variable with its component variance scaled from σ^2 to $\frac{3\sigma^2\mathrm{G}}{2(M-1)}$.

From (3.7), we can see that transformation to the bit-level basis by changing variables translates the reduced bit power of a higher-order modulation into the reduced component variance of the random variable of the bit-based channel fading effects. Furthermore, we can see in (3.7) that the power of the bit-based channel fading effects decreases exponentially as the QAM order increases. This is because the minimum distance between the constellation points decreases exponentially as the order of modulation increases, since the required number of points is proportional to 2 to the power of the modulation order.

Combining accordingly, the random variable of the total bit-based channel fading effects from the $\mathcal{R}\rightarrow\mathcal{D}$ link in the case of Fig. 2.1 is

$$\mathrm{z}_{\mathcal{RD}} = \textstyle\sum_{k=1}^{K_{\mathcal{R}}} \left(|h_{\mathcal{RD}_{j(k)}}|^2\right)_{\mathrm{A}} = \sum_{k=1}^{K_{\mathcal{R}}} \chi_2^2\left(\frac{3\sigma^2 \mathrm{G}}{2(M-1)}\right) = \chi_{2K_{\mathcal{R}}}^2\left(\frac{3\sigma^2 \mathrm{G}}{2(M-1)}\right) , \tag{3.8}$$

whereby its expectation is $\mathbb{E}\Big[\chi_{2K_{\mathcal{R}}}^2\big(\frac{3\sigma^2 \mathrm{G}}{2(M-1)}\big)\Big] = \mathrm{G}\frac{3K_{\mathcal{R}}\sigma^2}{2^{K_{\mathcal{R}}}-1}$.

3.3 Achievable Diversity Order

Utilizing the result of equations (3.2) and (3.8) in (3.4), the average BER of the proposed ADM scheme is

$$P_{\mathit{BER,\,ADM}} = \int_0^\infty \int_0^\infty Q\Big(\sqrt{2(z_{\mathcal{SD}} + z_{\mathcal{RD}})\mathtt{SNR}_{\mathcal{S}}}\,\Big) p_{\mathrm{z}_{\mathcal{SD}}}(z_{\mathcal{SD}}) p_{\mathrm{z}_{\mathcal{RD}}}(z_{\mathcal{RD}})\; dz_{\mathcal{SD}} dz_{\mathcal{RD}} \;, \tag{3.9}$$

where $\mathrm{z}_{\mathcal{SD}} = \chi_2^2(\sigma^2)$ and $\mathrm{z}_{\mathcal{RD}} = \chi_{2K_{\mathcal{R}}}^2\left(\frac{3\sigma^2\mathrm{G}}{2(M-1)}\right)$.

We use the upper bound $Q(x) \leq e^{-x^2/2}$, for $x > 0$ [7], in (3.9) to get

$$P_{\mathit{BER,\,ADM}} \leq \left(\frac{1}{1 + 2\sigma^2 \mathtt{SNR}_{\mathcal{S}}}\right)\left(\frac{1}{1 + \frac{3\sigma^2 G}{M-1}\mathtt{SNR}_{\mathcal{S}}}\right)^{K_{\mathcal{R}}} , \tag{3.10}$$

At high $\mathtt{SNR}_{\mathcal{S}}$, the above bound on the average BER of the proposed ADM scheme becomes

$$P_{BER,\,ADM} \leq \left(\frac{1}{\frac{6\sigma^4 G}{M-1}}\right)\mathtt{SNR}_{\mathcal{S}}{}^{-(K_{\mathcal{R}}+1)} = \left(\frac{2^{K_{\mathcal{R}}-1}}{6\sigma^4 G}\right)\mathtt{SNR}_{\mathcal{S}}{}^{-(K_{\mathcal{R}}+1)} . \qquad (3.11)$$

Thus, the average BER achieves $(K_{\mathcal{R}}+1)^{\text{th}}$-order diversity performance, one from the source and $K_{\mathcal{R}}$ from the relay as we intended.

3.4 Closed Form Analysis

In order to compute the average BER of the proposed scheme (3.9), we use $Q(x) = \frac{1}{\pi}\int_0^{\infty} \exp\left(-\frac{x^2}{2\sin^2\phi}\right)d\phi,\ x\geq 0$ from [11] and the moment-generating function (MGF) $\int_0^{\infty} \exp(-az)p_{\mathrm{z}}(z)dz = \left(\frac{1}{1+a2\sigma^2}\right)^{\frac{n}{2}}$ of a chi-square random variable $\mathrm{z} = \chi_n^2(\sigma^2)$. Then,

$$P_{BER,\,ADM} = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \left(\frac{\sin^2\phi}{\sin^2\phi + 2\sigma^2\mathtt{SNR}_{\mathcal{S}}}\right)\left(\frac{\sin^2\phi}{\sin^2\phi + \frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}}\right)^{K_{\mathcal{R}}} d\phi\ . \qquad (3.12)$$

The integral of equation (3.12) can be rendered in a closed form by [12, eqs. (5A.58), (5A.59) and (5A.60)] as

$$\begin{aligned} P_{BER,\,ADM} = \ & c\left[1 - \sqrt{\frac{2\sigma^2\mathtt{SNR}_{\mathcal{S}}}{1+2\sigma^2\mathtt{SNR}_{\mathcal{S}}}} - \frac{3\mathrm{G}}{2(M-1)}\sum_{k=0}^{K_{\mathcal{R}}-1}\left(1-\frac{3\mathrm{G}}{2(M-1)}\right)^k\right. \\ & \left.\left(1 - \sqrt{\frac{\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}}{1+\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}}}\left\{1+\sum_{n=1}^{k}\frac{(2n-1)!!}{n!2^n\left(1+\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}\right)^n}\right\}\right)\right], \end{aligned} \qquad (3.13)$$

where ! denotes the factorial and !! the double factorial notation denoting the product of only odd integers from 1 to $2n-1$. The coefficient c in (3.13) is calculated as $\frac{\left(\frac{2}{K_{\mathcal{R}}}\sum_{k=1}^{K_{\mathcal{R}}/2}\frac{2^k}{\sqrt{M}}\right)^2}{2\left(1-\frac{3\mathrm{G}}{2(M-1)}\right)^{K_{\mathcal{R}}}}$.

3.5 Trade-off between bit power and time diversity

We can see that the trade-off between bit power and additional time diversity in the $\mathcal{R}\rightarrow\mathcal{D}$ link is effectively resolved into the random variable of the bit-based channel fading effects (3.8), in the form of its increased degree of freedom and reduced component variance for high-order QAM. More specifically, Fig. 3.1 shows (3.8) for the case of 4QAM, 16QAM and 64QAM when G equals 1. The Rayleigh fading power $2\sigma^2$ is set to 1. As its expectation indicates, we can see that the expectation of $z_{\mathcal{RD}}$ falls as the order of modulation rises, although higher-order modulation presents a greater diversity-combining effect because of its sharper pdf. This is due to the fact that bit power decreases exponentially but diversity combining increases linearly as the order of modulation increases in the proposed method.

Since a bit-based constellation of relay faded by $z_{\mathcal{RD}}$ has its BER performance determined based on the destination WGN, a low value of $z_{\mathcal{RD}}$ requires low noise power to be error-free. As the order of modulation rises, the higher probability that $z_{\mathcal{RD}}$ will be low, as shown in Fig. 3.1, causes greater performance loss over the same noise level. Therefore, a higher signal power is required for the higher-order modulated relay packet to exhibit an increased time diversity performance over the destination WGN.

This characteristic also appears in the upper-bounded average BER of (3.11). A higher modulation order at the relay yields faster BER decay but decreases the coding gain $\left(\frac{2^{K_{\mathcal{R}}}-1}{6\sigma^4 G}\right)$ of (3.11) beyond the diversity performance. This explains the performance loss over the WGN for an increased time diversity. Consequently, this

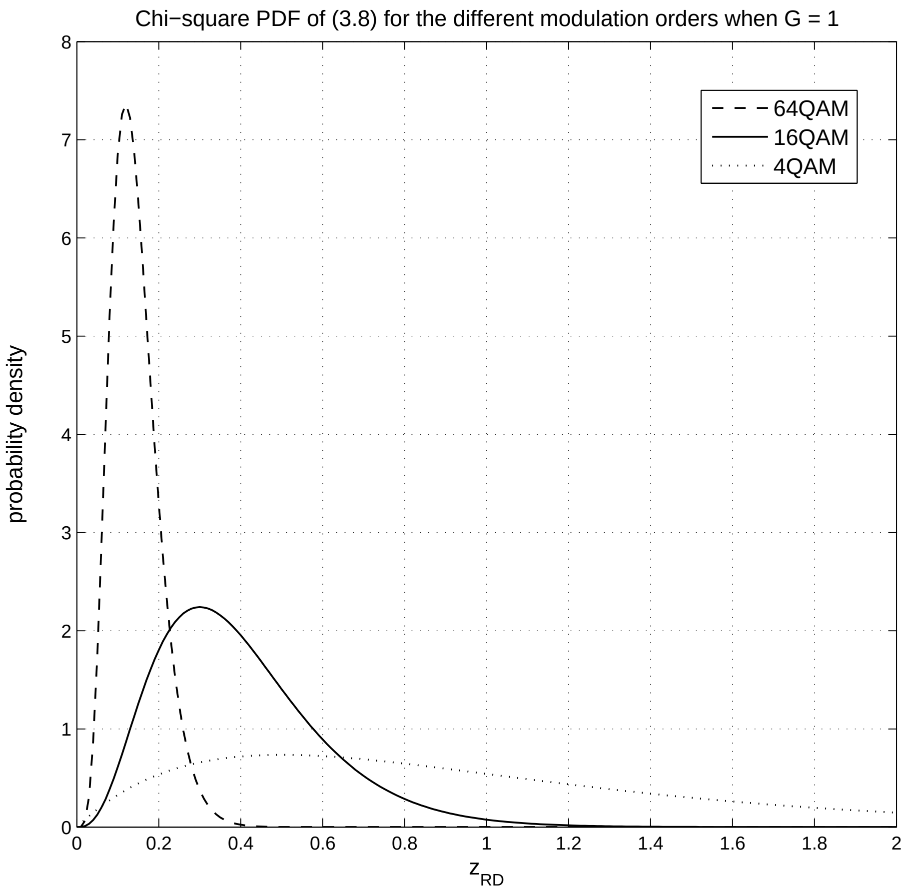

Figure 3.1: Chi-square pdf of (3.8) for the cases of 4QAM, 16QAM and 64QAM when G = 1. The figure clearly shows that the expected power of $z_{\mathcal{RD}}$ falls as the order of modulation rises, even though higher-order modulation has a greater diversity-combining effect because of its sharper pdf.

trade-off appears in the form of the outperforming modulation orders over certain ranges of $\mathtt{SNR}_{\mathcal{S}}$, as we will discuss in more detail in the next section.

Furthermore, we can also see in the expectation of (3.8) that G clearly provides a power gain no matter what order of modulation is employed at the relay. As in (3.11), G increases the coding gain. Thus, in the general situation where the relay provides less propagation loss than the source to the destination, G allows the relay to use higher-order modulation for higher time diversity performance by compensating the reduced bit power. Therefore, in a single source relaying topology, the proposed ADM scheme, which obtains additional time diversity from the $\mathcal{R}\rightarrow\mathcal{D}$ link, becomes practical.

3.6 Simulation Results

Fig. 3.2 depicts the average BER of the proposed ADM scheme and its theoretical performance under (3.13) when $\mathrm{G}=1$. A packet size of 240 symbols (conveying 240 bits per packet) was used in this simulation. The simulated block channel fading model has the same number of independent fadings as the order of QAM modulation employed at the relay. The Rayleigh fading power $2\sigma^2$ is set to 1 for the sake of convenience. A scheme where the relay uses BPSK (BPSK-ADM) without extracting additional time diversity in the $\mathcal{R}\rightarrow\mathcal{D}$ link is also simulated for comparison. Fig. 3.2 confirms that the numerical analysis of (3.13) closely matches the simulation results for $\mathtt{SNR}_{\mathcal{S}}$ above 5dB, since we derive (3.13) from the high-SNR approximated BEP of M-QAM.

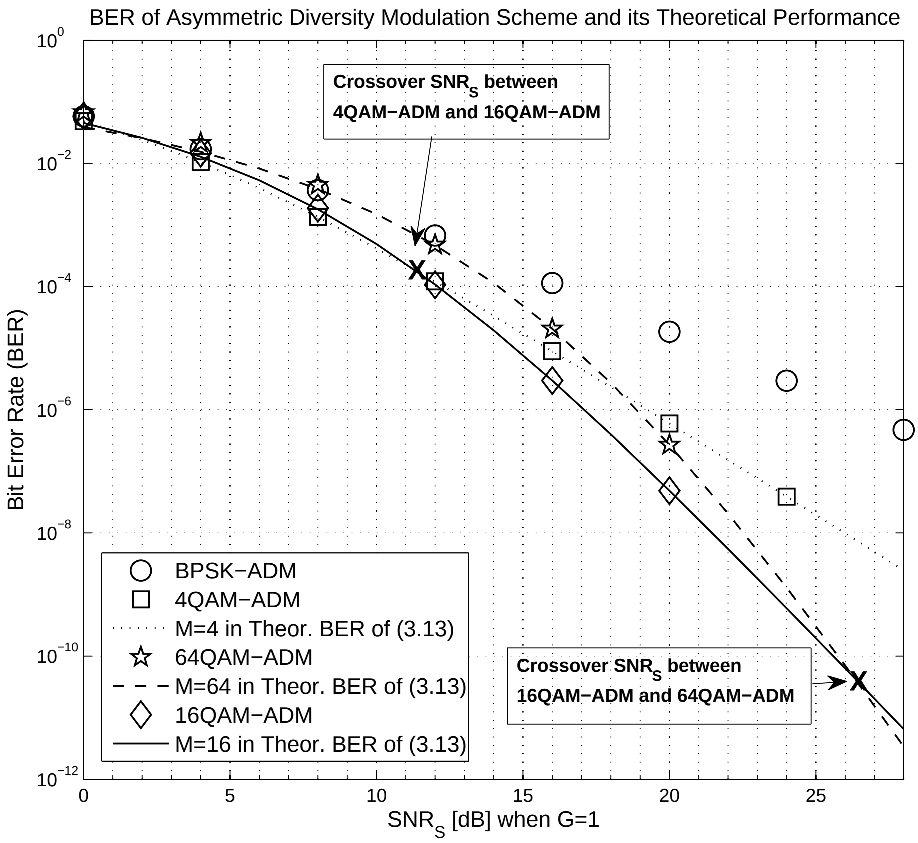

Figure 3.2: BER for the proposed ADM scheme and its theoretical performance in a wireless fading channel with one relay. A scheme showing the relay using BPSK (BPSK-ADM) is also presented for comparison. The theoretical performance tracks the corresponding simulation results quite well. Furthermore, due to its bit-based trade-off, a higher $\mathtt{SNR}_\mathtt{S}$ is required for the higher-order modulating relay to exhibit its additional time diversity performance.

Furthermore, there are clearly outperforming modulation orders for certain ranges of $\texttt{SNR}_{\mathcal{S}}$. Roughly speaking, 4QAM-ADM outperforms the others below 11dB of $\texttt{SNR}_{\mathcal{S}}$, 16QAM-ADM is best from 11dB to 26dB and so on for 64QAM-ADM. This is due to the fact that increased time diversity performance by higher-order modulating relays requires higher $\texttt{SNR}_{\mathcal{S}}$ values, as explained in Section 3.5. BPSK-ADM shows the worst performance over all ranges of $\texttt{SNR}_{\mathcal{S}}$ due to the limited degree of freedom in the time diversity aspect.

Table 3.1 shows the approximated crossover $\texttt{SNR}_{\mathcal{S}}$ between the different relay modulation orders for several values of G. As G increases, the $\texttt{SNR}_{\mathcal{S}}$ crossover points drop. As shown in the expectation of (3.8) and in the coding gain of (3.11), the power gain achieved by the decreased propagation loss of the relay does not depend on the order of modulation at the relay. Thus, G moves the entire BER curve left by the same amount. This lowers the crossover $\texttt{SNR}_{\mathcal{S}}$ due to the different diversity performance on the slope of each BER curve.

So, as the link quality between the relay and destination rises, the benefit of increasing additional time diversity using a higher-order modulating relay becomes realistic for a target $\texttt{SNR}_{\mathcal{S}}$. Therefore, for a single-source network, the proposed ADM scheme is an efficient diversity-enhancing method which converts the good $\mathcal{R}\rightarrow\mathcal{D}$ link quality into additional time diversity.

Table 3.1: Crossover $\texttt{SNR}_{\mathcal{S}}$ for several values of G considering the power gain ratio of the $\mathcal{R}\rightarrow\mathcal{D}$ link propagation loss versus the $\mathcal{S}\rightarrow\mathcal{D}$ link propagation loss. As G increases, the crossover $\texttt{SNR}_{\mathcal{S}}$ falls due to the fact that the power gain from the relay shifts the entire BER curve left by the same amount of $\texttt{SNR}_{\mathcal{S}}$, resulting in the crossover points moving leftward due to the different diversity slopes in the BER.

Case	Crossover $\texttt{SNR}_{\mathcal{S}}$ 4QAM-ADM vs 16QAM-ADM	Crossover $\texttt{SNR}_{\mathcal{S}}$ 16QAM-ADM vs 64QAM-ADM
G = 0.5	14.4 dB	29.3 dB
G = 1	11.3 dB	26.3 dB
G = 5	4.2 dB	19.3 dB
G = 15	-0.4 dB	14.5 dB

3.7 The BER Performance Comparison to the Signal Space Diversity (SSD) Techniques

In this section, we compare the BER performance of the proposed ADM scheme with the existing signal space diversity (SSD) schemes in the considered block channel fading model.

3.7.1 The principle of Signal Space Diversity (SSD) Techniques

The principle of SSD is to give multi-dimensional constellation points (multi-dimensional codewords) as many as distinct components as possible between codewords by rotation, thereby providing greater protection against the effects of channel fadings [4]. The SSD scheme is realized as follows: All inputs to the rotation matrix are bitstreams represented as 1 or -1 in the in-phase dimension of the input symbol. For an L-dimensional SSD scheme, each L input symbol in the packet is rotated by the $\mathrm{L} \times \mathrm{L}$ rotation matrix, which generates an L-dimensional codeword. This codeword goes through channel fading and white Gaussian noise (WGN). In order for each generated codeword to have time diversity, each L component of the L-dimensional codeword must undergo the uncorrelated channel fadings. For decoding, we search for the nearest codeword using the Maximum Likelihood (ML) criterion.

3.7.2 Comparison Environment

Let's say that we are transmitting 16 bits through a packet consisting of 16 symbols. If the channel model were to provide independent channel use for each symbol time in the packets transmitted, 16-dimensional codewords could be formed with the SSD schemes due to the uncorrelated channel gains in each component of a codeword. However, since we consider a more practical channel model with coherence time in order to keep the comparison fair, we compared the same time-diversity-achieving schemes. Therefore, in the proposed ADM scheme, we only consider the $\mathcal{R}\rightarrow\mathcal{D}$ link where the relay leads each bit to have additional time diversity performance.

3.7.3 2-dimensional Case

Fig. 3.3 describes the 2-dimensional schemes that we compare in terms of the average BER performance. The first scheme is the proposed 4QAM-Relay-ADM, the others are 2-dimensional SSD schemes that can be applied in the block fading channel model with 2 independent channel gains.

In the 2-dimensional SSD schemes, two consecutive bits are rotated to create a codeword (dotted-line box in Fig. 3.3). For example, bit A and bit B are rotated to create a codeword with the components $\mathrm{X_A}$ and $\mathrm{X_B}$. Then, each component is rearranged to different coherence times (gray box in Fig. 3.3). Other bits go through the same procedure.

Clearly, the average BER performance is determined by the rotation matrix in the SSD schemes. Let us define the rotation matrices by denoting R and we distin-

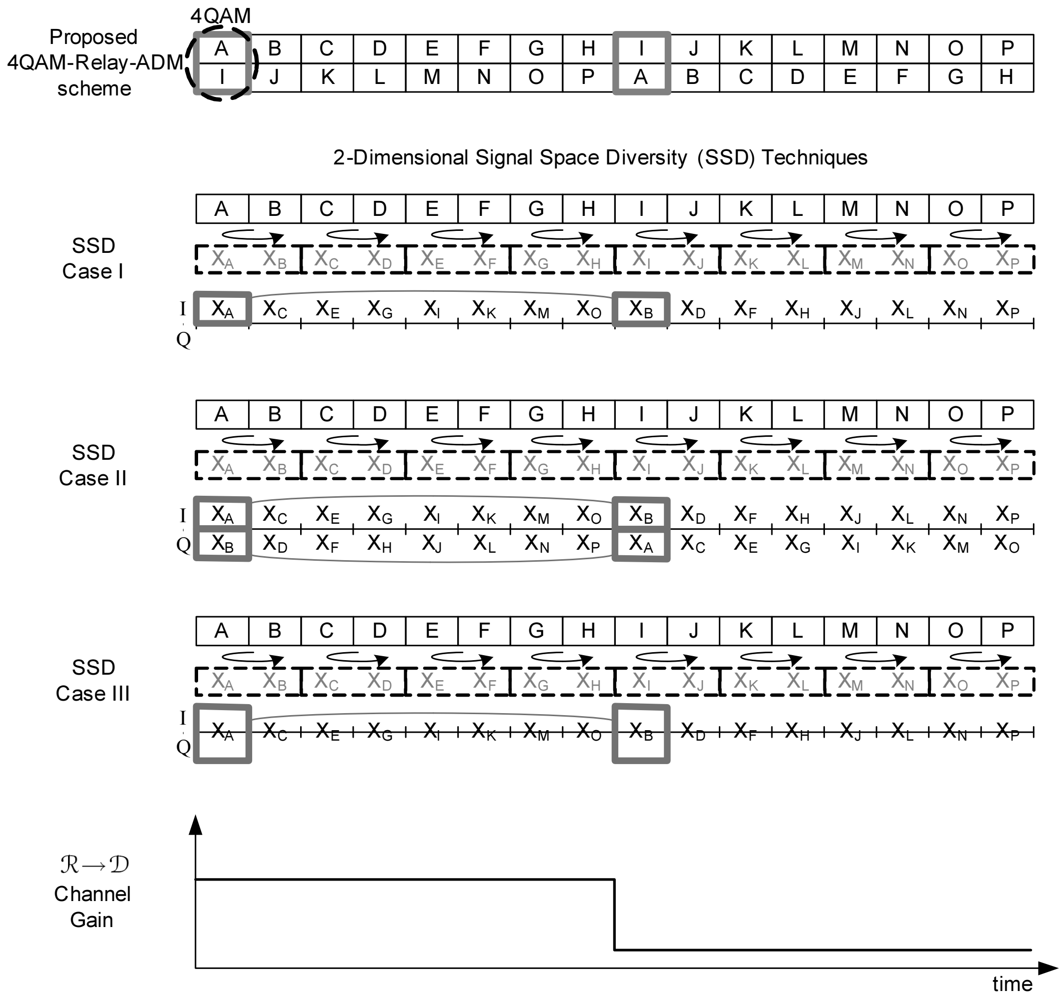

Figure 3.3: Proposed ADM schemes with 4QAM in the $\mathcal{R} \rightarrow \mathcal{D}$ link and the 2-dimensional SSD schemes for the average BER performance comparison. These schemes are considered in the block fading channel model with 2 independent channel gains.

guish them using two superscripts (the first superscript indicates its dimension and the second superscript its case indicator). For SSD Cases I and II, we use a simple rotation code which rotates two real input symbols into two real dimensions. This simplest rotation matrix is $\mathrm{R}_{2,0}=\left[\begin{smallmatrix}\cos\theta & -\sin\theta\\ \sin\theta & \cos\theta\end{smallmatrix}\right]$ with the optimum $\theta=\frac{1}{2}\tan^{-1}2$ [7]. The difference between the two cases is the I (in-phase) and Q (quadrature) channel used in the symbol space. Since the two channels are orthogonal, SSD Case I only exploits real space while SSD Case II fully utilizes the complex space by repeating the codeword, which is denoted as the complex rotation matrix of $\mathrm{R}_{2,1}=\left[\begin{smallmatrix}1 & j\\ j & 1\end{smallmatrix}\right]\mathrm{R}_{2,0}$. For SSD Case III, we use the 2-dimensioanl rotation matrix from the general L-dimensional matrix, achieving not only full diversity but also maximizing the minimum coding gain [13]. That general L-dimensional matrix is

$$\mathrm{R}_{\mathrm{L},2}=\frac{1}{\sqrt{\mathrm{L}}}\begin{bmatrix}1 & \alpha_0 & \cdots & \alpha_0^{\mathrm{L}-1}\\ 1 & \alpha_1 & \cdots & \alpha_1^{\mathrm{L}-1}\\ \vdots & \vdots & \ddots & \vdots\\ 1 & \alpha_{\mathrm{L}-1} & \cdots & \alpha_{\mathrm{L}-1}^{\mathrm{L}-1}\end{bmatrix} \tag{2.5.1}$$

where $\mathrm{L}=2^k$ for $k\in\mathbb{N}$ and $\alpha_i=\exp\big(j2\pi(i+1/4)/\mathrm{L}\big)$ for $i=0,1,\ldots,\mathrm{L}-1$. $\mathbb{N}$ stands for the positive integers. Then, the 2-dimensional rotation matrix for SSD Case III is $\mathrm{R}_{2,2}$, which expands the real input symbols into 2-dimensional complex space. Note that these rotation matrices are all unitary, so they preserve the energy and the Euclidean distance between the L-dimensional codewords.

The average BER performance comparison with the 2-dimensional SSD schemes is shown in Fig. 3.4. As you can see, the proposed 4QAM-Relay-ADM achieves the

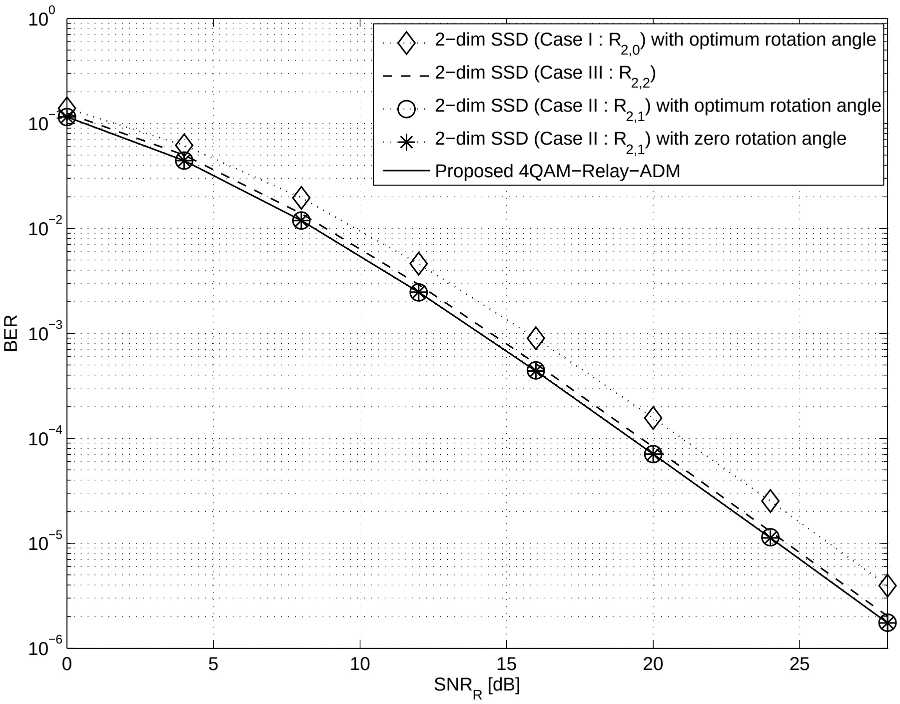

Figure 3.4: Average BER performance comparison between the proposed ADM scheme with 4QAM in the $\mathcal{R}\rightarrow\mathcal{D}$ link (4QAM-Relay-ADM) and the 2-dimensional SSD schemes. We can see that the proposed 4QAM-Relay-ADM achieves the best BER performance compared to the other 2-dimensional SSD schemes.

same second-order time diversity as the other cases of 2-dimensional SSD schemes do, while outperforming the coding gain. SSD Case I shows the worst coding gain due to its limited utilization of the signal space. Surprisingly, SSD Case II performs better than SSD Case III. In addition, SSD Case II not only achieves exactly the same BER performance as 4QAM-Relay-ADM but is also invariant to the rotation angle. This is due to the fact that, in the 2-dimensional case, there is a limitation on the dispersion of the constellation points. Thus, a simple repetition scheme such as the proposed 4QAM-Relay-ADM can show the best BER performance of all.

3.7.4 4-dimensional Case

For the 4-dimensional case, we consider cases similar to the 2-dimensional SSD schemes, as shown in Fig. 3.5. For SSD Cases I and II, we use the real 4×4 rotation matrix, $\mathrm{R}_{4,0}$, from $\mathrm{R}_{2,2}$ by replacing each complex entry $a+jb$ of $\mathrm{R}_{2,2}$ by a 2×2 matrix $\begin{pmatrix} a & -b \\ b & a \end{pmatrix}$ [4]. Instead, Case II also utilizes quadrature space by repeating the codeword, and is defined as the complex rotation matrix of $\mathrm{R}_{4,1}=\begin{bmatrix} 1 & 0 & j & 0 \\ 0 & 1 & 0 & j \\ j & 0 & 1 & 0 \\ 0 & j & 0 & 1 \end{bmatrix}\mathrm{R}_{4,0}$. For SSD Case III, we use a simple 4-point Discrete Fourier Transform (DFT) matrix and $\mathrm{R}_{4,2}$.

Fig. 3.6 shows the average BER performances of the above schemes. First, we can see that the real rotation matrix of $\mathrm{R}_{4,0}$ and its complex-version $\mathrm{R}_{4,1}$, which are derived from $\mathrm{R}_{2,2}$, limit their diversity performance to the second order due to the second-order diversity property of their origin $\mathrm{R}_{2,2}$. Repeating through the quadrature space in SSD Case II gives more than 3dB coding gain since complex

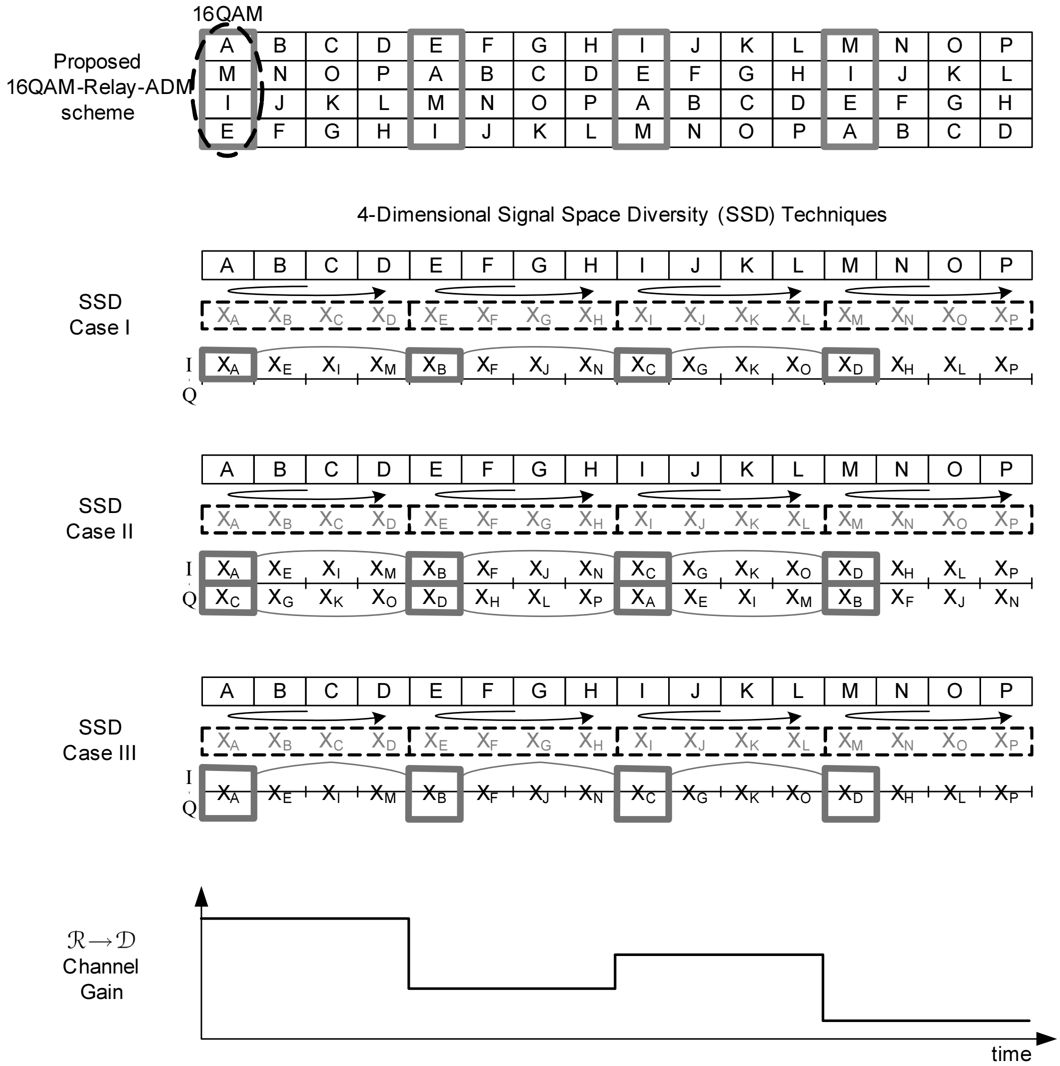

Figure 3.5: Proposed ADM schemes with 16QAM in the $\mathcal{R} \rightarrow \mathcal{D}$ link and 4-dimensional SSD schemes for average BER performance comparison. These schemes are considered in the block fading channel model with 2 independent channel gains.

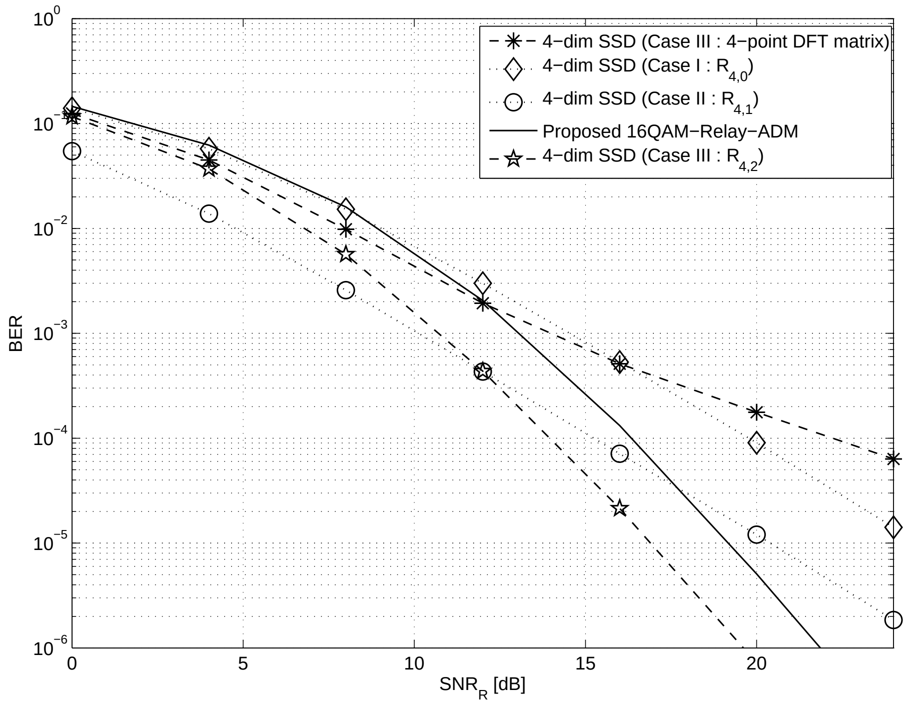

Figure 3.6: Average BER performance comparison between the proposed 16QAM-Relay-ADM scheme in the $\mathcal{R}\rightarrow\mathcal{D}$ link and the 4-dimensional SSD schemes. Fig. 3.6 clearly shows that, unlike the 4-point DFT matrix, 16QAM-Relay-ADM achieves fairly good BER performance without losing any diversity, even while still in its simplest coding structure.

codewords clearly provide larger product distance between each other than real codewords, thereby maximizing the minimum product distance. Moreover, we can see that SSD Case III with $\mathrm{R}_{4,2}$ not only obtains 4^{th}-order diversity performance as 16QAM-Relay-ADM but also provides a larger coding gain, which shows that $\mathrm{R}_{4,2}$ fully disperses the codewords over the 4-dimensional complex spaces. However, SSD Case III with the 4-point discrete Fourier transform (DFT) matrix exhibits diversity-limited BER performance due to its poor rotation, whereas 16QAM-Relay-ADM achieves fairly good BER performance without losing any diversity. While we can be certain that there are better performance-achieving rotation matrices than the proposed ADM method in 4-dimensional cases or higher, we can nevertheless conclude from these comparisons that the proposed ADM scheme provides comparable BER performances with its simplest coding structure, especially in regard to the simple rotation matrices of the existing SSD schemes.

Chapter 4

Diversity-Enhancing with XOR-Relay in the ADM scheme

IN this chapter, diversity-enhancing technique combined with the asymmetric diversity modulation (ADM) is further proposed. The proposed method is a simple encoding scheme that XORs the consecutive bits at the relay. Without wasting additional resources, an XOR-relay provides an increased number of diversity paths, enabling the destination to construct a cycle-free decoding structure, and resulting in an enhanced diversity order. Section 4.1 describes the system and transmission model of the proposed scheme. Section 4.2 analyzes in terms of theoretical bit error rate (BER) via numerical analysis and compares the proposed XOR-Relay scheme to the conventional case in the situation of asymmetric diversity modulation (ADM). Closed form bit error rate (BER) of the proposed scheme is presented in Section

4.3. Section 4.4 explains the iterative maximum a porsteriori (MAP) decoding at the destination and discusses its complexity burden. In Section 4.5, we confirm the increased diversity order of the proposed XOR-Relay scheme via the simulation results.

4.1 The system and transmission model of the proposed XOR-Relay scheme

As shown in Fig. 4.1, the considered system and transmission models are the same as those considered in Chapter 3. $\mathcal{S}$ delivers its packet to $\mathcal{D}$ over $\mathcal{R}$ with the time-divided transmission of $\mathcal{S}$ and $\mathcal{R}$ as explained in the previous chapter.

The conventional scheme is transmitting the same packet from $\mathcal{S}$ and $\mathcal{R}$ to $\mathcal{D}$ at each time slot (repetition-coding of the relay). It does not matter whether the relay employs higher modulation order to utilize the ADM or not. What matters is that the relay of the conventional scheme just retransmits the received packet from the source without any processing on those bits.

However, the relay of the proposed scheme XORs the consecutive bits of the received packet and transmits, as depicted in Fig. 4.1. The basic idea underlying accomplishing increased diversities in BER, stems from the fact that the number of independent fading channels which the bits are carried over determines the diversity order. Bit B in the proposed XOR-Relay experiences three fading channels through $x_{\mathcal{S}_2}$, $x_{\mathcal{R}_2}$ and $x_{\mathcal{R}_3}$, even though it is XORed with other bits. Bit C suffers the same through $x_{\mathcal{S}_3}$, $x_{\mathcal{R}_3}$ and $x_{\mathcal{R}_4}$, and so on. Proceeding in this way, we can force each bit to undergo increased channel fadings in the $\mathcal{R}\to\mathcal{D}$ link with the aid of XORing.

What is important is that the combination with the ADM does not limit its potential to increase diversity order. In the ADM scheme, the relay packet is duplicated and rearranged to be carried by higher modulation symbol in order to give addition

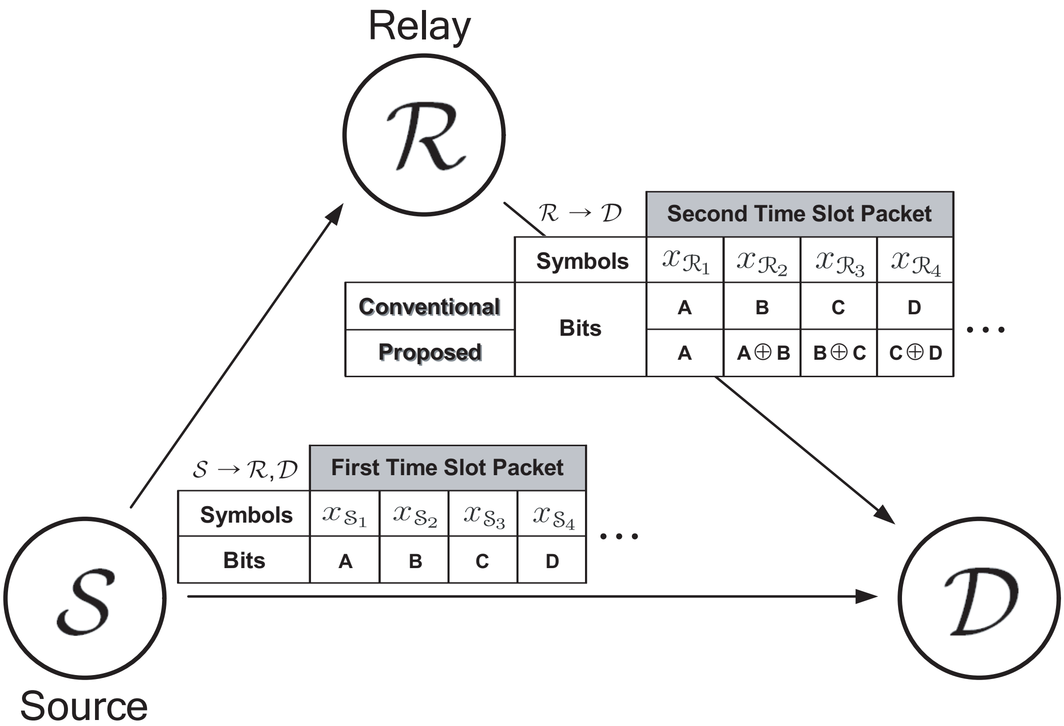

Figure 4.1: Conventional scheme and Proposed XOR-Relay scheme in a wireless fading channels with one relay.

time diversity in the $\mathcal{R}\rightarrow\mathcal{D}$ link. Since additional time diversities are provided no matter what kind of bits are carried through the ADM, the XORed bits of the relay packet will have successfully time diversities in combination with the ADM.

Therefore, compared to the conventional case, the proposed XOR-Relay scheme can provide more reliabilities from the $\mathcal{R}\rightarrow\mathcal{D}$ link to each bit transmitted from the source. If we suppose that the original bits of the conventional case and the XORed bits of the proposed scheme have the same time diversities, the protection from the $\mathcal{R}\rightarrow\mathcal{D}$ link is doubled since each bit is consecutively XORed by the relay in the proposed XOR-Relay scheme. In the next section, we will investigate this aspect through the numerical analysis in terms of BER.

4.2 Performance Analysis of the Proposed XOR-Relay scheme

Let us consider the system model where the source is using BPSK and the relay is using $K_{\mathcal{R}}^{\text{th}}$-order modulation. Since in the previous chapter we proposed 'bit'-based channel combining method in deriving the theoretical BER of the ADM scheme, we are also using this method in deriving the BER of the proposed scheme.

The key point of the proposed 'bit'-based channel combining method is that we fix the basis of diversity combining as a 'bit' itself, which is represented as a bit error probability(BEP), $Q\left(\sqrt{2zSNR}\right)$, of an BPSK symbol. Thus, the channel fading effects on the 'bit'-basis ("bit-based channel fading effects" as described in section

3.2) can be thought as if BPSK symbol has gone through this translated channel. Therefore, if we use this bit-based channel fading effects, we can translate our system model such that all symbols are transmitted using BPSK, no matter what its actual modulation order is.

Utilizing this point of view, we derive the theoretical BER of the conventional and the proposed scheme in the next subsections.

4.2.1 Conventional Case

As solved in the previous chapter, the probability density functions (PDF) of bit-based channel fading effects from the $\mathcal{S}\rightarrow\mathcal{D}$ and the $\mathcal{R}\rightarrow\mathcal{D}$ link when the source is using BPSK and the relay is using $K_{\mathcal{R}}^{\text{th}}$-order modulation are $\mathrm{z}_{\mathcal{S}\mathcal{D}} = \chi_2^2(\sigma^2)$ and $\mathrm{z}_{\mathcal{R}\mathcal{D}} = \chi_{2K_{\mathcal{R}}}^2\left(\frac{3\sigma^2\mathrm{G}}{2(M-1)}\right)$, respectively.

Since the conventional case is exactly the same as the ADM scheme, we rewrite its theoretical BER of (3.9) and (3.12) as

$$\begin{aligned} P_{BER,\,CONV} &= \mathbb{E}_{\mathrm{z}_{\mathcal{S}\mathcal{D}},\,\mathrm{z}_{\mathcal{R}\mathcal{D}}}\left[Q\Big(\sqrt{2(z_{\mathcal{S}\mathcal{D}}+z_{\mathcal{R}\mathcal{D}})\mathtt{SNR}_{\mathcal{S}}}\,\Big)\right] \\ &= \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\left(\frac{\sin^2\phi}{\sin^2\phi+2\sigma^2\mathtt{SNR}_{\mathcal{S}}}\right)\left(\frac{\sin^2\phi}{\sin^2\phi+\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}}\right)^{K_{\mathcal{R}}} d\phi\ . \end{aligned} \tag{4.1}$$

4.2.2 Proposed XOR-Relay Scheme

The proposed scheme can be seen as a linear block code, with the systematics from the source and parities from the relay. Fig. 4.1 shows that the destination can separate the received bits into systematics (A, B, C) and parities (A⊕B, B⊕C) as (3,5) linear block code.

To derive its increased diversity order in the theoretical BER, let us focus on bit B, assuming that bits A and C are perfectly decoded at the destination. If so, the BER of bit B is transferred to pairwise error probability, confusing the codeword $\mathbf{X}_{B0}$ when bit B is zero for the codeword $\mathbf{X}_{B1}$ when it is one, i.e., $\mathbf{X}_{B0} = [0\ 0\ 0]$ and $\mathbf{X}_{B1} = [1\ 1\ 1]$ if bit A and C are decided as all zeros.

Since we translate the typical Rayleigh fadings into the bit-based channel fading effects in the ADM scheme, we can think that all bits are transmitted through BPSK symbols. In this case, the codeword symbols depending on bit B are always opposite each other in any decided bit combination of bits A and C. Hence, the conditional BER for bit B detected as one when zero is transmitted over three translated channel fadings, $\mathbf{h}' = [h_{A\oplus B}\ h_B\ h_{B\oplus C}]$, is

$$Pr\{\ B\ :\ 0 \rightarrow 1 \mid \mathbf{h}'\ \} = Pr\{\ \mathbf{X}_{B0}\ \rightarrow\ \mathbf{X}_{B1} \mid \mathbf{h}'\ \} = Q\left(\frac{\|\mathbf{u}_{B0} - \mathbf{u}_{B1}\|}{2\sqrt{\frac{N_0}{2}}}\right) \tag{4.2}$$

where $\mathbf{u}_{B0}$ and $\mathbf{u}_{B1}$ are received vectors over fadings in a complex-white Gaussian process, and the quantity $\frac{\|\mathbf{u}_{B0}-\mathbf{u}_{B1}\|}{2}$ is the Euclidean distance from each vector to the decision boundary [7].

Averaging (4.2) over $\mathbf{h}'$ yields

$$P_{BER,\,PROP} = \mathbb{E}_{\mathbf{h}'}\left[Q\left(\frac{\|\mathbf{u}_{B0} - \mathbf{u}_{B1}\|}{2\sqrt{\frac{N_0}{2}}}\right)\right] = \mathbb{E}_{\mathbf{h}'}\left[Q\left(\sqrt{2(|h_{A\oplus B}|^2+|h_B|^2+|h_{B\oplus C}|^2)\mathtt{SNR}_S}\right)\right] \tag{4.3}$$

Since the systematics are transmitted from the source and the parities from the relay, $|h_B|^2 = \mathrm{z}_{\mathcal{SD}}$ and $|h_{A\oplus B}|^2 = |h_{B\oplus C}|^2 = \mathrm{z}_{\mathcal{RD}}$, respectively. Utilizing this relationship, the average BER of the proposed XOR-Relay scheme is expressed as

$$\begin{aligned} P_{BER,\,PROP} &= \mathbb{E}_{\mathbf{h}'}\left[Q\left(\sqrt{2(|h_{A\oplus B}|^2+|h_B|^2+|h_{B\oplus C}|^2)\mathtt{SNR}_{\mathcal{S}}}\right)\right] \\ &= \mathbb{E}_{\mathrm{z}_{\mathcal{SD}},\,\mathrm{z}_{\mathcal{RD}}}\left[Q\left(\sqrt{2(z_{\mathcal{SD}}+2\,z_{\mathcal{RD}})\mathtt{SNR}_{\mathcal{S}}}\,\right)\right] \\ &= \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\left(\frac{\sin^2\phi}{\sin^2\phi+2\sigma^2\mathtt{SNR}_{\mathcal{S}}}\right)\left(\frac{\sin^2\phi}{\sin^2\phi+\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}}\right)^{2K_{\mathcal{R}}} d\phi\,. \end{aligned} \tag{4.4}$$

Hence, the proposed scheme is able to achieve the increased BER performance from $(K_{\mathcal{R}}+1)^{\mathrm{th}}$-order to $(2K_{\mathcal{R}}+1)^{\mathrm{th}}$-order diversity. Due to the XORing at the relay, we can clearly see in the above equation that the proposed XOR-Relay scheme provides double protections from the $\mathcal{R}\rightarrow\mathcal{D}$ link on each bit transmitted from the source.

4.3 Closed Form Analysis

Using [12, eqs. (5A.58), (5A.59) and (5A.60)], the integral of equation (4.4) can be rendered in a closed form as

$$\begin{aligned} P_{BER,\,PROP} = \;& c\left[1-\sqrt{\tfrac{2\sigma^2\mathtt{SNR}_{\mathcal{S}}}{1+2\sigma^2\mathtt{SNR}_{\mathcal{S}}}}-\tfrac{3\mathrm{G}}{2(M-1)}\textstyle\sum_{k=0}^{2K_{\mathcal{R}}-1}\left(1-\tfrac{3\mathrm{G}}{2(M-1)}\right)^k\right. \\ & \left.\left(1-\sqrt{\tfrac{\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}}{1+\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}}}\left\{1+\textstyle\sum_{n=1}^{k}\tfrac{(2n-1)!!}{n!2^n\left(1+\frac{3\sigma^2\mathrm{G}}{(M-1)}\mathtt{SNR}_{\mathcal{S}}\right)^n}\right\}\right)\right], \end{aligned} \tag{4.5}$$

where ! denotes the factorial and !! the double factorial notation denoting the product of only odd integers from 1 to $2n-1$. The coefficient c in (4.5) is calculated as $\frac{\left(\frac{2}{K_{\mathcal{R}}}\sum_{k=1}^{K_{\mathcal{R}}/2}\frac{2^k}{\sqrt{M}}\right)^4}{2\left(1-\frac{3\mathrm{G}}{2(M-1)}\right)^{2K_{\mathcal{R}}}}$.

4.4 Iterative MAP Decoder

A refined decoder can guide us to approach the lower bound of the BER, which in this thesis is diversity order $(2K_{\mathcal{R}}+1)^{\text{th}}$. Decoding a linear block code iteratively following the maximum a posteriori (MAP) rule can achieve a high performance gain [14], and the systematics and parities in the linear block code can be represented as a factor graph to be decoded iteratively using a Sum-Product algorithm which utilizes the LLR of each bit [15]. Fig. 4.2 shows a factor graph representation of the proposed XOR-Relay scheme. Each bit node ($\mathcal{S}\rightarrow\mathcal{D}$ systematics) is connected through the variable nodes ($\mathcal{R}\rightarrow\mathcal{D}$ parities) by a line called an 'edge'.

Each node has its own intrinsic LLR channel output message, represented as '**i**', and each bit node extracts an a posteriori LLR message '**P**' via the decoding process. All connected nodes exchange messages among each other, following the rule such that the output message along one edge of a node is a function of the inputs along all the other edges of the node. The definition of message-calculation at each node and the rule following the Sum-Product Algorithm are explained fully in [15].

Note that this node-distributed message-passing computation of soft decoding can succeed when the graph is cycle-free. Our relay-encoding design (XORing consecutive bits) not only provides the transmitted bits with an increased number of independently faded paths, but also enables the destination to construct the cycle-free factor graph for iterative MAP decoding that is shown in Fig 4.2.

However, the diversity performance of the proposed scheme is rather limited compared to the high-complex Forward Error Correcting (FEC) codes' redundancies

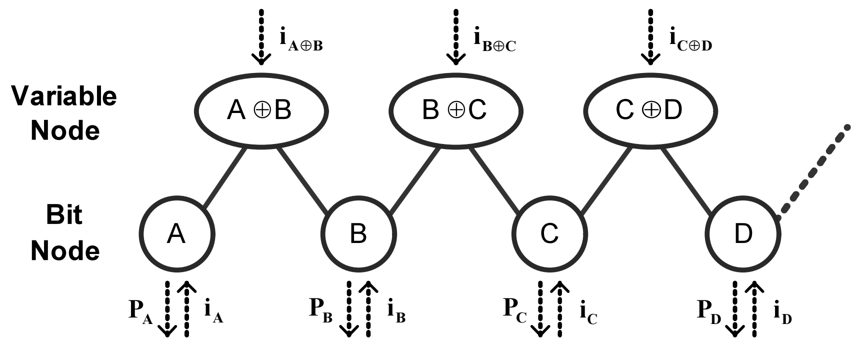

Figure 4.2: Factor graph representation of the proposed scheme, constructed at the destination for iterative MAP decoding. The parities XORing the consecutive bits make this graph cycle-free.

such as an Turbo code and Low Density Parity Check (LDPC) code [16,17]. Fig 4.3 shows the BER performance when the relay is transmitting the parities of the rate 1/2 LDPC code (576,288) instead of the consecutively XORed redundancies of the proposed scheme. When the relay is using BPSK, LDPC-parities relaying obtains nearly 10dB coding gain with 10 iterations compared to our proposed scheme. But there is no doubt that the iterative decoding entails some added complexity at the destination. If we are relaying the parities of the powerful FEC codes, the encoding and decoding complexity can be significantly large enough to be a hugh burden to the system, since the burden at each node arises from the number of connected edges to which the calculated messages have to be passed. However, the added complexity of the proposed XOR-Relay seems reasonable in connection with the simplest edge connection thanks to the cycle-free decoding structure.

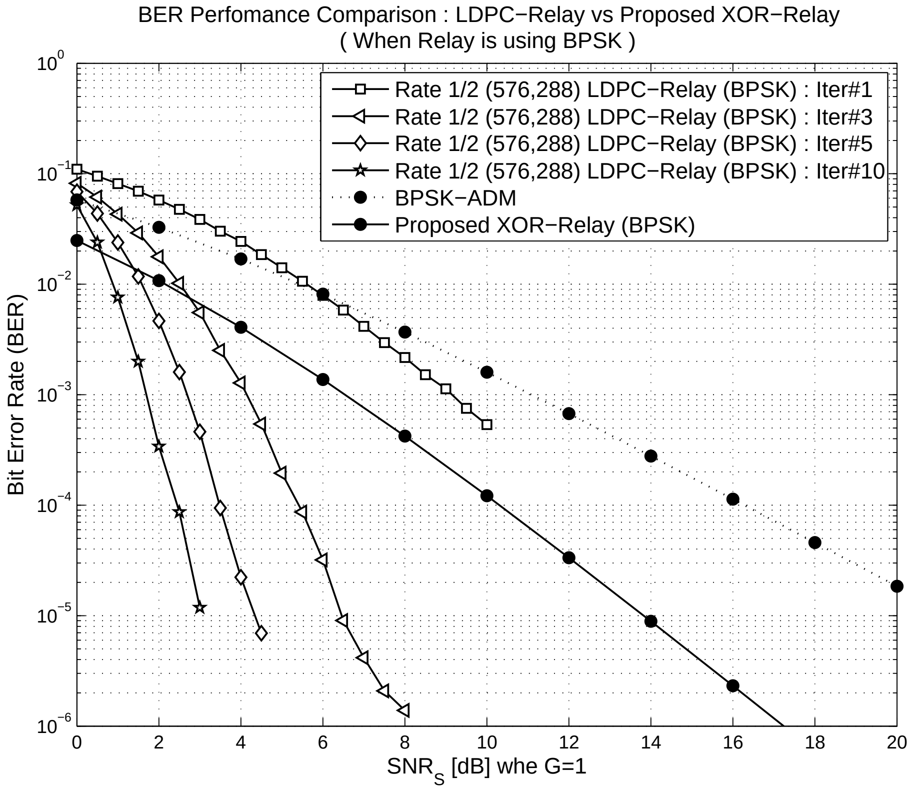

Figure 4.3: BER performance comparison between rate 1/2 LDPC (576,288) parities and the proposed XOR-Relay redundancies. We can see that the coding gain of LDPC-parities relaying is much better than that of the proposed scheme with the higher encoding and decoding complexity.

4.5 Simulation Results

Fig. 4.4, Fig. 4.5, and Fig. 4.6 depict the average BER of the proposed XOR-Relay scheme and its theoretical performance under (4.5) when $\mathrm{G} = 1$. A packet

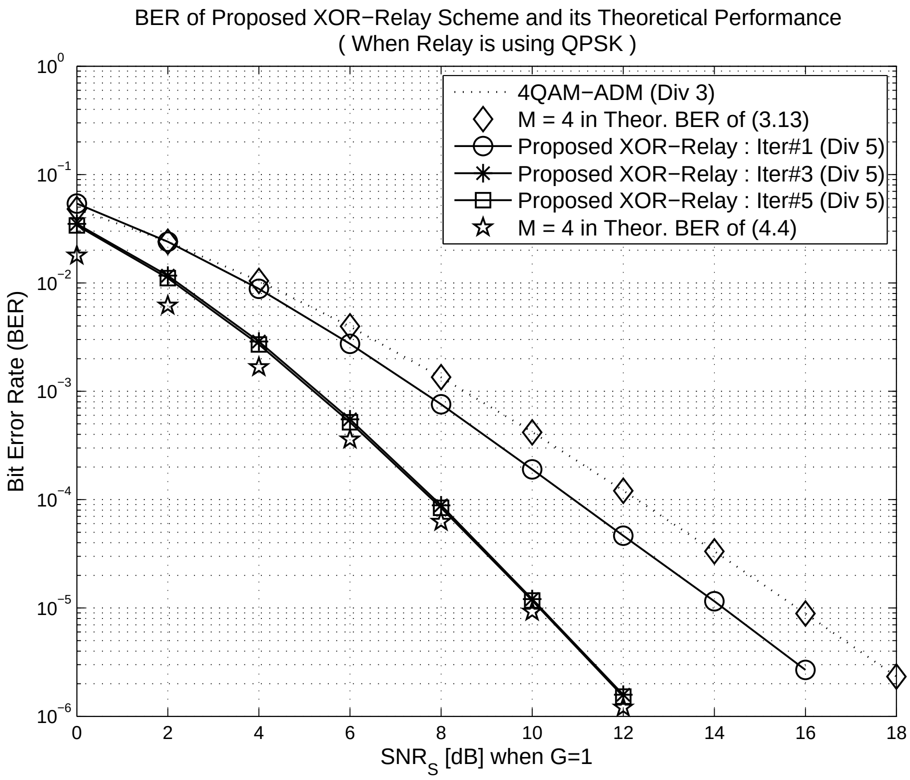

Figure 4.4: BER for the conventional scheme and the proposed XOR-Relay scheme in a wireless fading channel with one relay using QPSK. Iterative MAP decoding of the proposed scheme clearly derives improved diversity performance approaching the theoretical 5^{th}-order diversity without using any more resources than the conventional system.

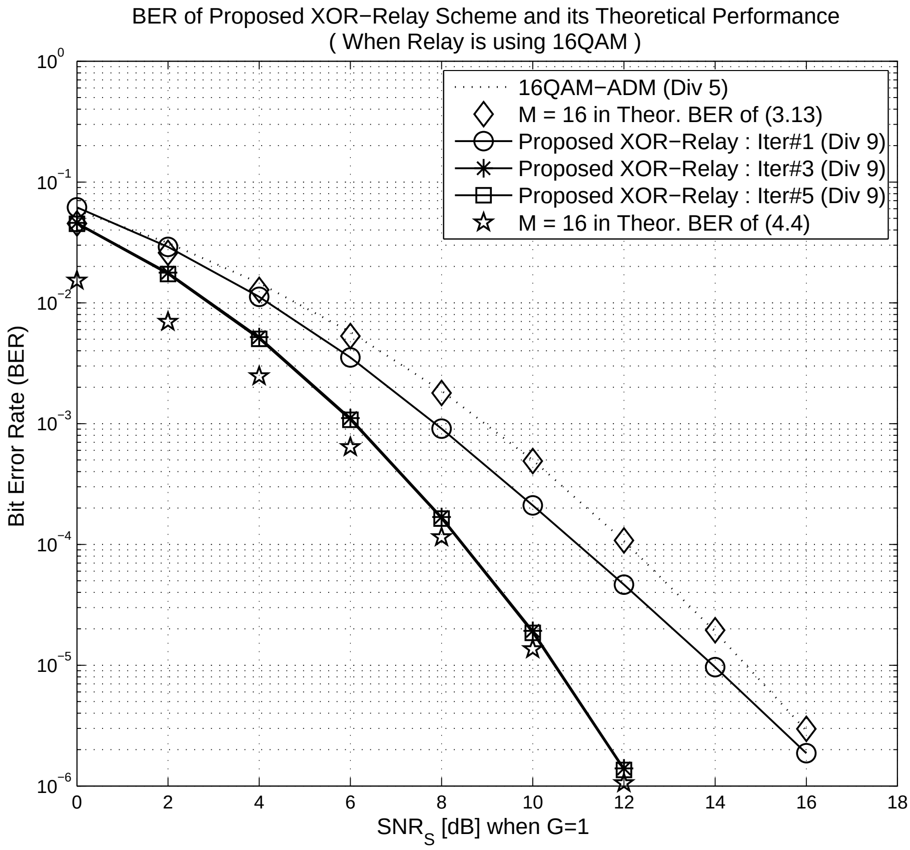

Figure 4.5: BER for the conventional scheme and the proposed XOR-Relay scheme in a wireless fading channel with one relay using 16QAM. Iterative MAP decoding of the proposed scheme clearly derives improved diversity performance approaching the theoretical 9$^{\text{th}}$-order diversity without using any more resources than the conventional system.

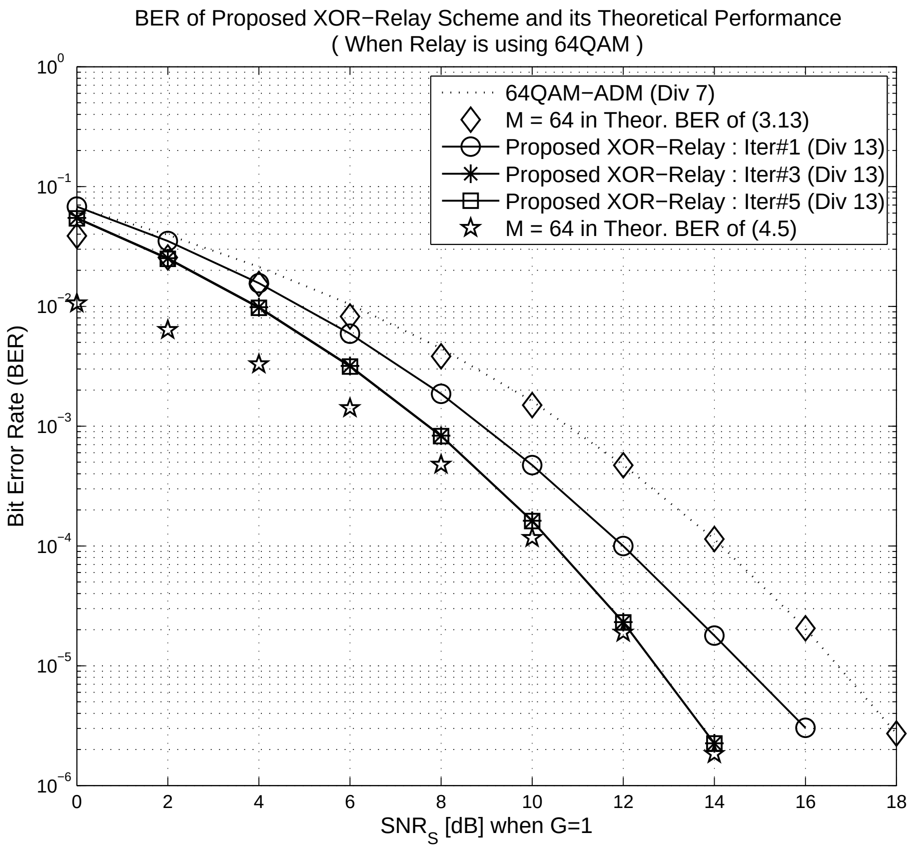

Figure 4.6: BER for the conventional scheme and the proposed XOR-Relay scheme in a wireless fading channel with one relay using 64QAM. Iterative MAP decoding of the proposed scheme clearly derives improved diversity performance approaching the theoretical 13^{th}-order diversity without using any more resources than the conventional system.

size of 240 symbols (conveying 240 bits per packet) was used in this simulation, although the packet size does not greatly affect the performance with the proposed scheme, since the iterative decoder operates in a parallel fashion. The simulated block channel fading model has the same number of independent fadings as the order of QAM modulation employed at the relay. We can see that these figures confirm that the simulation results approaches to the maximum lower bound of the proposed XOR-Relay scheme, which is from the numerical analysis of (4.5).

Since the relay provides an increased number of paths for the transmitted bits by XORing, the BER performance of the proposed scheme approaches the maximum $(2K_{\mathcal{R}}+1)^{\text{th}}$ diversity level that it could potentially reach as it iterates, especially in the high SNR region. As shown in the analysis, an enhanced diversity order can be accomplished based on exact detection of adjacent bits. Thus, performance degrades to a point equivalent to the conventional scheme as the SNR falls, since many of the LLR messages are not reliable in the iterative decoding process.

However, the proposed XOR-Relay scheme clearly obtains enhanced diversity performance even though it consumes the same resources as the conventional scheme. And thanks to its simple and cycle-free decoding structure, three iterations are enough to approach the maximum diversity performance of $(2K_{\mathcal{R}}+1)^{\text{th}}$-order.

For more clear comparison between the conventional case of the ADM scheme and the proposed XOR-Relay scheme, theoretical results of (3.13) and (4.5) with different modulation orders are depicted in Fig. 4.7. A scheme where the relay uses BPSK without extracting additional time diversity in the $\mathcal{R}\rightarrow\mathcal{D}$ link is also depicted for

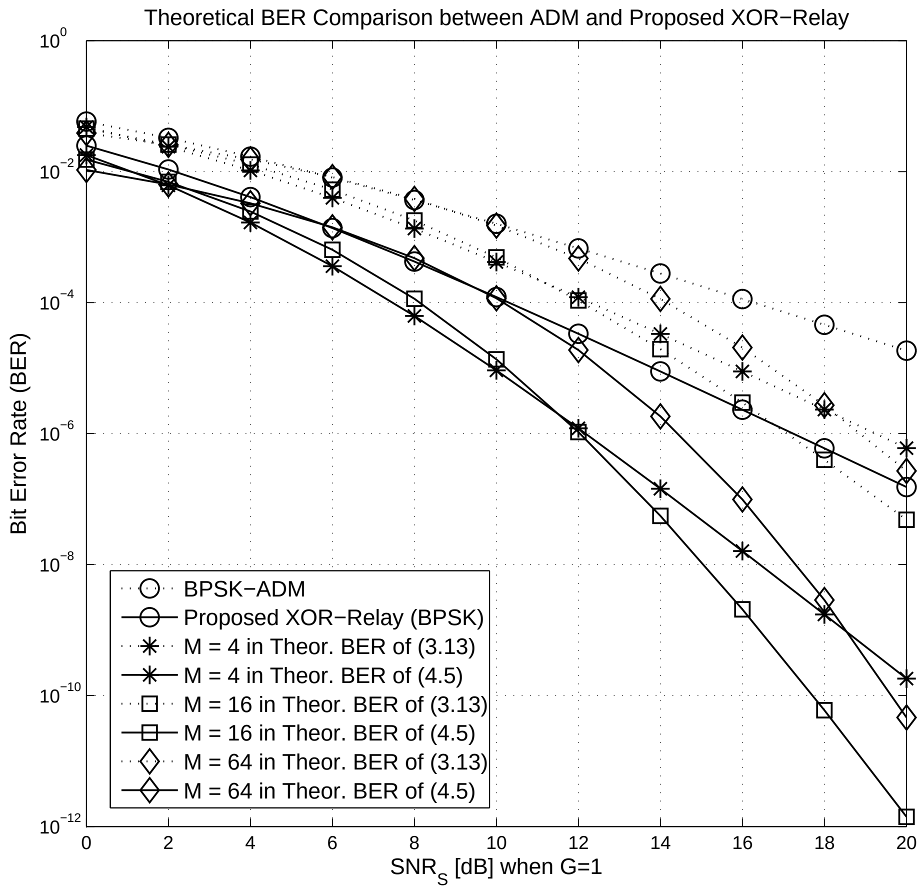

Figure 4.7: Theoretical BER comparison between the conventional scheme (3.13) and the proposed XOR-Relay scheme (4.5) with different modulation orders. It shows that the proposed XOR-Relay can clearly achieve the increased diversity performance with a few iterations, no matter what modulation order is used at the relay.

comparison. We can clearly see in this figure that the proposed methods successfully derives the increased diversity performance without wasting additional resources, no matter what modulation order is employed at the relay.

Chapter 5

Conclusion

IN this thesis, we proposed a new diversity scheme which extracts additional time diversity in wireless fading relay channels. We also derived the theoretical BER performance numerically based on a proposed 'bit'-based channel-combining method and analyzed a trade-off relationship between the bit power and time diversity arising from a high-order modulating relay. Finally, we confirmed our analysis via simulation results and compared its performance to existing SSD techniques. The results demonstrated that the proposed method efficiently enhances diversity through relaying in the block fading channel model.

We also presented a simple technique for achieving higher diversity performance, combining an XORing at the relay and an iterative MAP decoder at the destination. Enhanced diversity performance comes from the increased number of fading paths on each bit by the XORing operation at the relay, and the high performance delivered by a cycle-free iterative MAP decoder with a few iterations. Combined with the ADM scheme, we investigated the proposed method and showed its increased diversity performance via numerical analysis and simulation results.

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국 문 요 약

무선 페이딩 채널 환경에서 비대칭적 다이버시티 변조를 이용한 다이버시티 증대 기법

본 논문에서는 단일 소스-릴레이 시스템에서 릴레이의 높은 전송 능력을 다이버시티 형태로 전환하는 비대칭적 다이버시티 변조 기법을 제안한다. 제안된 기법은 릴레이에서 높은 차수의 변조를 사용하여 복수의 소스 비트들을 전송함으로서 부가적인 시간 다이버시티를 제공한다. 그리고 데스티네이션에서는 '비트' 단위로 공간과 시간 다이버시티를 결합한다. 또한 제안된 '비트' 단위의 채널 결합 방법을 사용하여, 이러한 시스템에서의 이론적인 bit error rate (BER)을 구하고, 릴레이에서 높은 차수의 변조를 사용함에 따른 비트 파워와 시간 다이버시티 간의 trade-off 관계를 분석한다.

또한, 본 논문에서는 릴레이에서 XOR을 이용하여 부가적인 자원 손실 없이 다이버시티 차수를 증대시키는 기법을 제안한다. 제안된 XOR-릴레이 기법은 다이버시티 경로 수를 증대시키며, 데스티네이션에서 순환 없는 디코딩 구조를 만듦으로서 증대된 다이버시티 효과를 가져 온다. 앞서 제안된 '비트' 단위의 채널 결합 방법을 사용하여, 이러한 시스템에서의 이론적인 bit error rate (BER)을 구하며, 모의실험 결과를 통해서 제안된 기법은 데스티네이션에서 적은 횟수의 반복적 디코딩만으로도 증대된 다이버시티 차수에 근접하는 것을 확인한다.

핵심되는 말 : Wireless fading relay channel, diversity, modulation, bit error probability, iterative MAP decoding