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ภาคผนวก

ผลงานวิจัยของผู้เขียนที่ได้รับการตีพิมพ์แล้ว

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On Performance of Full-Rate Differentially Encoded Cooperative Communications:Bit Error Rate Bounds

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Abstract—**This paper investigates a performance of differentially encoded cooperative communications achieving full rate in three-user cooperation scenario with a fixed amplify-and-forward protocol. Tight upper and lower bounds on an averaged bit error rate (BER) for such system are also developed. These proposed bounds significantly reduce computational complexity of evaluating an exact averaged BER. Simulation results show that the proposed bounds are tight to the exact one in fairly-high signal to noise ratio (SNR) regimes.**

I. INTRODUCTION

Multiple-input multiple-output (MIMO) systems can provide several desirable advantages to the future wireless communications with broad coverage areas by the virtue of a spatial diversity gain, inherited in such system [1], [2]. Recently, cooperative diversity techniques have been proposed for realizing the benefits of MIMO systems in a distributed fashion by allowing a single antenna at the handset to be shared by the cooperative partners in wireless networks [1], [2]. These communication techniques received significant interest because a single antenna at the handset consumes less energy than the conventional MIMO system.

Various communication protocols have been proposed for the cooperative wireless networks, in which one or more neighbor users may act as a relay by amplify-and-forward (AF) or decode-and-forward (DF) the received signal to the destination [1]. In [3], the two-user cooperative diversity systems using the differential modulation technique with the AF protocol and the quadrature differential phase shift keying modulation (Q-DPSK) have been investigated. In [4], the DF and selection relaying protocols (SR) for the two-user cooperative communication system employing (Q-DPSK) modulation have been studied. However, there is a lack of studies for the differentially encoded cooperation systems, where more than two users are active, with a full transmission rate constraint. This is the motivation of this paper. The main contributions of this paper are as follows,

- The simple tight upper and lower bounds on an averaged bit-error-rate (BER) are proposed.
- we can show that in the full-rate differentially encoded AF cooperation system, the proper cooperation strategy solely depends on a signal to noise ratio (SNR)of the systems, taking into account a channel gain.

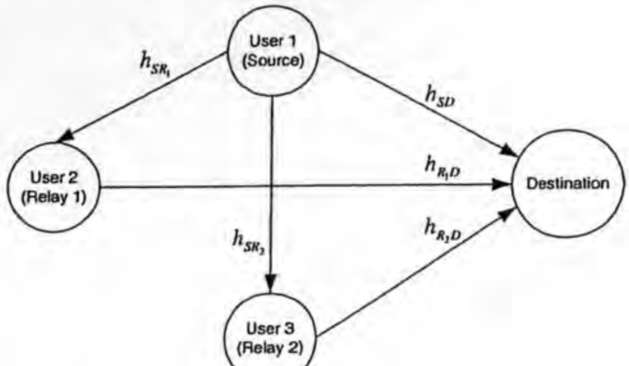

Fig. 1: The cooperative wireless networks with three active users and one destination.

This paper is organized as follows. In section II, the channel and system models to be studied are described. In section III, we analyze the averaged BER together with the upper and lower bounds for such system, including a noncooperation strategy with B-DPSK modulation, a 1-relay cooperation strategy with Q-DPSK modulation, and a 2-relay cooperation strategy with 8-DPSK modulation. In section IV, the simulation results and discussion are given. Finally, conclusions are given in section V.

II. CHANNEL AND SYSTEM MODELS

For the sake of exposition, let us consider the cooperative wireless networks with three active users and one destination, i.e. one user acts as a source node and the remaining two users act as relay nodes, as illustrated on Fig.1.

In addition, we consider flat Rayleigh fading channels in this study, We also assume that all users transmit their signal through orthogonal channels by using time-division multiplexing (TDMA), in which the transmission is divided into two phases [1]. In phase I, user 1 acts as a source node and transmits its modulated signal to the destination. By the broadcast nature of wireless channels, the transmitted signal is also broadcast to user 2 and user 3. In phase II, user 2 and user 3 act as relay nodes that amplify and forward the received signal to user 1's destination, respectively.

Since the full-rate differentially encoded cooperation system

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is considered in this paper, we employ the differential modulation scheme with adaptive modulation in order to maintain the full rate, i.e. rate 1. Specifically, the B-DPSK modulation is used for direct transmission, i.e. a noncooperation strategy. For a 1-relay cooperation strategy, where either relay1 or relay 2 is employed, Q-DPSK modulation is used. Finally, for a 2-relay cooperation strategy, where both relay 1 and relay 2 are employed, 8-DPSK modulation is used.

The basic principle of differential modulation is to use the previous symbol as a phase reference for the current symbol under an assumption that the channels for two consecutive information symbols are almost the same [5], i.e. slow fading channels. Specifically, the information bits are encoded as the differential phase between the current symbol and the previous symbol. The source node first differentially encodes the symbol *S(n)* as follows,

$$S(n) = e^{j\phi_i} S(n-1), \tag{1}$$

where $\phi_i = \frac{2\pi i}{M}, i=0,1\ldots,M-1$ with *M* being the constellation size of the adaptive modulation and $S(n-1)$ is the previous symbol. For the sake of simplicity, we omit a time parameter *n* in the following. In addition, all cooperation strategies have a fixed total transmit power *P* per symbol. In phase I, the user 1 broadcasts the symbol S to the destination and both relay nodes. The received signal at the destination Y_{SD}, at the relay 1 Y_{SR_1}, and at the relay 2 Y_{SR_2}, can be expressed as follows, respectively,

$$Y_{SD} = \sqrt{P_1} h_{SD} S + n_{SD} \tag{2}$$

$$Y_{SR_1} = \sqrt{P_1} h_{SR_1} S + n_{SR_1} \tag{3}$$

$$Y_{SR_2} = \sqrt{P_1} h_{SR_2} S + n_{SR_2} \tag{4}$$

where h_{SD}, h_{SR_1} and h_{SR_2} are the channel coefficients from source to destination, source to relay 1, and source to relay 2, respectively; n_{SD}, n_{SR_1} and n_{SR_2} are the additive noise with zero-mean and variance N_0 at the destination, relay 1, and relay 2, respectively; is the transmit power of the source node. In phase II, the relay 1 and relay 2 amplify the received signal and forward it to the destination with the transmit power $\tilde{P}_2$ and $\tilde{P}_3$, respectively [2]. The received signal at the destination Y_{R_1D}, transmitted from the relay 1, and Y_{R_2D}, transmitted from the relay 2, can be expressed as follows,

$$Y_{R_1D} = \sqrt{\tilde{P}_2} h_{R_1D} Y_{SR_1} + n_{R_1D}, \tag{5}$$

$$Y_{R_2D} = \sqrt{\tilde{P}_3} h_{R_2D} Y_{SR_2} + n_{R_2D}, \tag{6}$$

$$\tilde{P}_2 = \begin{cases} \frac{P_2}{P_1 \delta^2_{SR_1} + N_0} & \text{; If relay 1 is selected} \\ 0 & \text{; Other} \end{cases}$$

$$\tilde{P}_3 = \begin{cases} \frac{P_3}{P_1 \delta^2_{SR_2} + N_0} & \text{; If relay 2 is selected} \\ 0 & \text{; Other} \end{cases}$$

where h_{R_1D} and h_{R_2D} are the channel coefficients from relay 1 and relay 2 to the destination and relay 2 to the destination, respectively; n_{R_1D} and n_{R_2D} are the additive noises with zero-mean and variance N_0. In the Reyleigh fading channels, the channel coefficients h_{SD}, h_{SR_1}, h_{SR_2}, h_{R_1D} and h_{R_2D} are modelled as independent zero-mean, complex Gaussian random variables with variances δ^2_{SD}, $\delta^2_{SR_1}$, $\delta^2_{SR_2}$, $\delta^2_{R_1D}$ and $\delta^2_{R_2D}$, respectively.

At the destination, all received signals are combined using the maximum ratio combining (MRC) technique [6] expressed as follows,

$$Z = a_1 Y_{SD} + a_2 Y_{R_1D} + a_1 Y_{R_1D}, \tag{7}$$

where $a_1 = \frac{P_1|h_{SD}|^2}{N_0}$, $a_2 = \frac{P_1P_2|h_{SR_1}|^2|h_{R_1D}|^2}{N_0[P_1\delta^2_{SR_1}+P_2|h_{R_1D}|^2+N_0]}$, and $a_3 = \frac{P_1P_3|h_{SR_2}|^2|h_{R_2D}|^2}{N_0[P_1\delta^2_{SR_2}+P_3|h_{R_2D}|^2+N_0]}$. Since we assume that the knowledge of channel coefficients is not available to the receiver except the channel variances, we propose to replace all instantaneous channel gain, i.e. $|h_{SD}|^2$, $|h_{SR_1}|^2$, $|h_{SR_2}|^2$, $|h_{R_1D}|^2$ and $|h_{R_2D}|^2$, in a_1, a_2, and a_3 by their channel variances, i.e. δ^2_{SD}, $\delta^2_{SR_1}$, $\delta^2_{SR_2}$, $\delta^2_{R_1D}$ and $\delta^2_{R_2D}$. Hence, we propose to linearly combine all received signals in a similar way to (7) with $a_1 = \frac{P_1\delta^2_{SD}}{N_0}$, $a_2 = \frac{P_1P_2\delta^2_{SR_1}\delta^2_{R_1D}}{N_0[P_1\delta^2_{SR_1}+P_2\delta^2_{R_1D}+N_0]}$, and $a_3 = \frac{P_1P_3\delta^2_{SR_2}\delta^2_{R_2D}}{N_0[P_1\delta^2_{SR_2}+P_3\delta^2_{R_2D}+N_0]}$. The decoder uses the sufficient statistics to decode information symbol $\hat{S}(n)$ as follow,

$$\hat{S}(n) = \arg_{m=0,1,\ldots,M-1} \max Re[(e^{-j\phi_i}) Z(n)], \tag{8}$$

where *Re*[ ] is the real operator.

III. BER ANALYSIS FOR FLAT RAYLEIGH FADING CHANNEL

In this section, we derive the exact averaged BER of the cooperation systems with the M-DPSK modulation. The analysis here is based on the MRC combiner, and then the approximate SNR of the combine output is $\gamma \approx \gamma_1 + \gamma_2 + \gamma_3$ [6] where $\gamma_1 = \frac{P_1|h_{SD}|^2}{N_0}$, $\gamma_2 = \frac{P_1P_2|h_{SR_1}|^2|h_{R_1D}|^2}{N_0[P_1\delta^2_{SR_1}+P_2|h_{R_1D}|^2+N_0]}$, and $\gamma_3 = \frac{P_1P_3|h_{SR_2}|^2|h_{R_2D}|^2}{N_0[P_1\delta^2_{SR_2}+P_3|h_{R_2D}|^2+N_0]}$. The results of this calculation serve as the performance lower bound to our proposed scheme in the previous section. However, as we will see in the simulation section, the analytical results and the simulated results are almost the same. The conditional BER of M-DPSK modulation for the cooperation system given the channel coefficients h_{SD}, h_{SR_1}, h_{SR_2}, h_{R_1D} and h_{R_2D} can be expressed as follows [7],[8],

$$P_{(M-DPSK)} = \frac{\xi}{2^{2L}\pi} \int_{-\pi}^{\pi} f(\theta) \exp[-\gamma\alpha(\theta)] d(\theta) \tag{9}$$

where $\xi = 1$ when B-DPSK is employed, $\xi = \frac{4}{\log_2 M}$ with M being the modulation constellation size when M-DPSK is employed, i.e. $M > 2$, $\alpha(\theta) = \frac{b^2(1+2\beta\sin(\theta)+\beta^2)}{2}$. Here, $0 < \beta = \frac{a}{b} \leq 1$ is a constant which a and b depended on the modulation size, L is a number of independent identically distributed Rayleigh fading paths of the received signal, γ is a total SNR per bit, and $f(\theta)$ is a modulation function depending on β and L.

We can derive the averaged BER of such systems, averaged over all channels, as follows. For the noncooperation strategy with B-DPSK modulation, the averaged BER can be express as follows,

$$P_{(B-DPSK)}^{Noncoop} = \frac{1}{4\pi}\int_{-\pi}^{\pi} f(\theta)M\gamma_1 d(\theta), \qquad (1$$

where $f(\theta) = \frac{b^2(1-\beta^2)}{2\alpha(\theta)}$ and $M\gamma_1(\theta) = \frac{1}{1+k_{SD}(\theta)}$ wi $k_{SD}(\theta) = \alpha(\theta)\frac{P_1\delta_{SD}^2}{N_0}$ [3] being the moment generating fun tion (MGF) of γ_1. For the noncooperation strategy, we ha $a = 10^{-3}$ and $b = \sqrt{2}$ [5].

For the 1-Relay cooperation strategy with Q-DPSK modul tion, the averaged BER can be expressed as follows,

$$P_{(Q-DPSK)}^{Relay1,2} = \frac{1}{8\pi}\int_{-\pi}^{\pi} f(\theta)M\gamma_1 M\gamma_2 d(\theta), \qquad (1$$

where

$$f(\theta) = \frac{b^2}{2\alpha(\theta)}[(1-\beta^2)(3+\cos(2\theta)) - (\beta^{-1}-\beta^3)\sin(\theta)],$$

$M\gamma_2(\theta) = \frac{1}{1+k_{SR_1}(\theta)}\left[1 + \frac{P_1\delta_{SR_1}^2}{P_2\delta_{R_1D}^2}\frac{k_{SR_1}(\theta)}{1+k_{SR_1}(\theta)}Z\gamma_2\right]$ denotes the MGF of γ_2, $k_{SR_1}(\theta) = \alpha(\theta)\frac{P_1\delta_{SR_1}^2}{N_0}$, $Z\gamma_2 = \int_0^\infty \exp\left(\frac{u}{\delta_{R_1D}^2}\right)\left[u + R\gamma_2(\theta)\right]^{-1}du$, and $R\gamma_2(\theta) = \frac{P_1\delta_{SR_1}^2+N_0}{P_2}\left[1 \frac{P_1\delta_{SR_1}^2 b^2(1+2\beta\sin(\theta)+\beta^2)}{2N_0}\right]^{-1}$, respectively [3]. In addition, t $M\gamma_1$ is similar to (10). For 1-relay cooperation strategy, have $a = \sqrt{2-\sqrt{2}}$ and $a = \sqrt{2+\sqrt{2}}$ [5]. We can compute approximate averaged BER by upper and lower bounding t $(1+2\beta\sin(\theta)+\beta^2)$ term in $R\gamma_2(\theta)$ by minimum or maximu values corresponding to $\theta = \frac{\pi}{2}$ or $\theta = -\frac{\pi}{2}$, respectively[8], follows,

$$R\gamma_2(\min) = \frac{P_1\delta_{SR_1}^2+N_0}{P_2}\left[1 + \frac{P_1\delta_{SR_1}^2 b^2(1+\beta)^2}{2N_0}\right]^{-1} \text{ and}$$

$$R\gamma_2(\max) = \frac{P_1\delta_{SR_1}^2+N_0}{P_2}\left[1 + \frac{P_1\delta_{SR_1}^2 b^2(1-\beta)^2}{2N_0}\right]^{-1}.$$

The upper bound and lower bound on the averaged can be computed by substituting $R\gamma_2(\min)$ and $R\gamma_2(\max)$, respectively, into $Z\gamma_2$.

For the 2-Relay cooperation strategy with 8-DPSK modulation, the averaged BER can be expressed as follows,

$$P_{(8-DPSK)}^{2-Relay} = \frac{1}{48\pi}\int_{-\pi}^{\pi} f(\theta)M\gamma_1 M\gamma_2 M\gamma_3 d(\theta) \qquad (12)$$

where $f(\theta) = \frac{b^2}{2\alpha(\theta)}[10(1-\beta^2) + (5 - 5\beta^2 - \beta^{-2} + \beta^4) \cdot\cos(2\theta) - 5(\beta^{-1}-\beta^3)\sin\theta - (\beta^{-1} - \beta^3)\sin(3\theta)]$, $M\gamma_3(\theta) = \frac{1}{1+k_{SR_2}(\theta)}\left[1 + \frac{P_1\delta_{SR_2}^2}{P_3\delta_{R_2D}^2}\frac{k_{SR_2}(\theta)}{1+k_{SR_2}(\theta)}Z\gamma_3\right]$ being the MGF of γ_3, $k_{SR_2}(\theta) = \alpha(\theta)\frac{P_1\delta_{SR_2}^2}{N_0}$, $Z\gamma_3 = \int_0^\infty \exp\left(-\frac{u}{\delta_{R_2D}^2}\right)\left[u + R\gamma_3(\theta)\right]^{-1}du$, and $R\gamma_3(\theta) = \frac{P_1\delta_{SR_2}^2+N_0}{P_3}\left[1 + \frac{P_1\delta_{SR_2}^2 b^2(1+2\beta\sin(\theta)+\beta^2)}{2N_0}\right]^{-1}$, respectively. In addition, the $M\gamma_1$ and $M\gamma_2$ are similar to (10) and (11), respectively. For 2-relay cooperation strategy, we

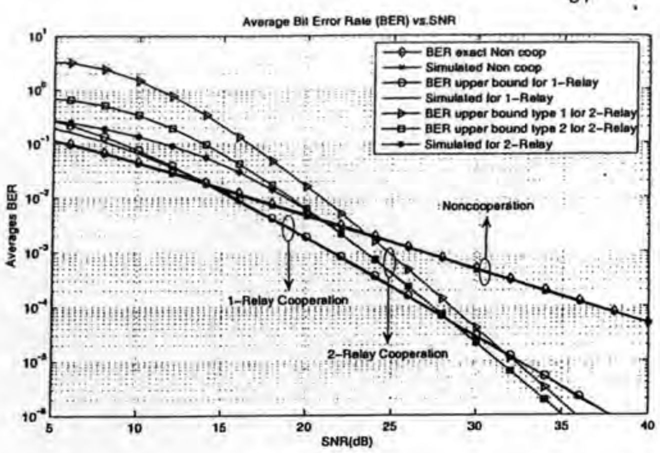

Fig. 2: The curves of the averaged BER and the upper bound for differentially encoded cooperative communications with three active users.

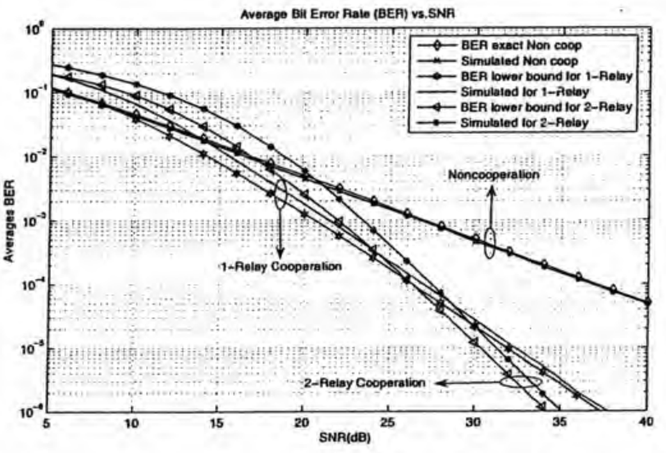

Fig. 3: The curves of the averaged BER and the lower bound for differentially encoded cooperative communications with three active users.

have $a = \sqrt{2-\sqrt{2-\sqrt{2}}}$ and $b = \sqrt{2+\sqrt{2-\sqrt{2}}}$ [8]. We can compute an approximate averaged BER by upper and lower bounding the $(1+2\beta\sin(\theta)+\beta^2)$ term in $R\gamma_3(\theta)$ by minimum or maximum values corresponding to $\theta = \frac{\pi}{2}$ or $\theta = -\frac{\pi}{2}$, respectively[8] as follows,

$$R\gamma_3(\min) = \frac{P_1\delta_{SR_2}^2+N_0}{P_3}\left[1 + \frac{P_1\delta_{SR_2}^2 b^2(1+\beta)^2}{2N_0}\right]^{-1} \text{ and}$$

$$R\gamma_3(\max) = \frac{P_1\delta_{SR_2}^2+N_0}{P_3}\left[1 + \frac{P_1\delta_{SR_2}^2 b^2(1-\beta)^2}{2N_0}\right]^{-1}.$$

There are two ways to upper bound (12): Type 1) by substituting $Z\gamma_2$ and $Z\gamma_3$ with $R\gamma_2(\min)$ and $R\gamma_3(\min)$, respectively; Type 2) by substituting $Z\gamma_2$ with $R\gamma_2(\min)$ and $Z\gamma_3$ by $R\gamma_3(\max)$. Furthermore, the lower bound can be obtained by substituting

$Z\gamma_2$ and $Z\gamma_3$ with $R\gamma_2(\max)$ and $R\gamma_3(\max)$, respectively.

IV. Simulation Results

In this section, we perform a computer simulation for the system in which the B-DPSK, Q-DPSK and 8-DPSK adaptive modulation and the AF protocol are employed. The fading channel coefficients are taken from Jake's model with Doppler shift of 3.5KHz, baud rate of 1 Mbps, and bandwidth efficiency of 1 bit/s/Hz [2]. For fair comparison, we plot the averaged BER curves as a function of P/N_0, where P is the total transmit power, that is fixed to be 1 W. We assume that the variance of all noises is 1, i.e. $N_0 = 1$. In this study, the variance of the channel links between source node and destination δ^2_{SD}, source node and relay 1, $\delta^2_{SR_1}$, and source node and relay 2 $\delta^2_{SR_2}$ are 1, 5, and 10, respectively. The variance of the channel links between relay 1 and destination $\delta^2_{R_1D}$ is 1, and between relay 2 and destination $\delta^2_{R_2D}$ is 1.

In Fig.2, we illustrate the curves of the simulated averaged BER, the exact averaged BER, and the proposed upper bound on the averaged BER for the noncooperative scheme, 1-relay cooperative schems, and 2-relay cooperative scheme. For the noncooperation and 1-relay cooperation schemes, the proposed upper bound is close to the exact one throughout the SNR range. For the 2-relay cooperation schems, the proposed upper bound type 2 is more tight than type 1. For example, the upper bound type 2 is close to the exact one at SNR of 20 dB, whereas the upper bound type 1 is parallel to the exact one with the SNR gap of 1.1 dB.

In Fig.3, we illustrate the curves of the simulated averaged BER, the exact averaged BER, and the proposed lower bound on the averaged BER for the noncooperative scheme, 1-relay cooperative schems, and 2-relay cooperative scheme. It is worth noticing that the proposed lower bound is tight to the exact one except for the 2-relay cooperation scheme, where the gap of 0.8 dB between such bound and the exact one is observed.

The proposed upper bound is very useful because it is tight and it requires less computational complexity for evaluation than the exact averaged BER. By assuming the knowledge of all channel variances at the transmitting node, we can compute this bound and use it as a performance index to determine the optimum cooperation strategy, i.e. the number of cooperative partners and the type of modulation constellation. This would be our future work.

V. Conclusion

In this paper, we have analyzed the performance of the differentially encoded cooperative communication systems achieving full rate with three active users in term of the averaged BER. We have also proposed the tight and simple upper and lower bounds on such averaged BER. Simulation results revealed that the proposed bounds are tight to the exact one in fairly high SNR regimes. In addition, with full rate constraint, it is worth noticing that the proper cooperation strategy, in the sense of the minimum probability of detection error, solely depends on the SNR of the systems taking into account the channel gain, i.e. a small number of cooperative users is preferred in low SNR regimes, whereas a large number of cooperative users is preferred in high SNR regimes.

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Tight Approximate Bounds on Bit Error Rate for Full-Rate Differentially Encoded Cooperative Communications

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Abstract—**In this paper, we propose tight approximate bounds on the averaged bit error rate (BER) for differentially cooperative communications achieving full rate in three-user scenario. Under a fixed amplify-and-forward (AF) protocol with fixed transmission rate and power constraints, the proposed approximate bounds are tight to the exact BER in fairly-high signal-to-noise ration (SNR) regimes. Furthermore, the computational complexity of the proposed bounds are significantly lower than the existing bounds. A sub-optimum power allocation for such system is also studied, based on a guideline of an optimum power allocation for a coherent cooperative communication system. Simulation results show superior performance for the sub-optimum power allocation scheme to the equal power allocation one.**

I. Introduction

Multiple-input multiple-output (MIMO) techniques can significantly enhance performance of wireless communication systems through spatial diversity gain. Recently, cooperative diversity techniques have emerged as an alternative way for achieving such gain in such a way that a single antenna at the handset in the multiuser wireless network is shared to form a virtual MIMO system [1],[2].

Various communication protocols have been proposed for the cooperative communication systems [1]-[6]. In [1], the outage probability for several relaying protocols has been investigated, including fixed amplify-and-forward (AF) and decode-and-forward (DF) protocols and an adaptive protocol. In [3], the adaptive AF protocol and optimum power allocation for the three-user cooperative communication system, in which fixed power and full transmission rate constraints are taken into account, have been proposed. In [4], the two-user cooperative communication system employing the quadrature differential Phase shift keying (Q-DPSK) modulation with the AF protocol has been investigated. In [5], the DF and selection relaying (SR) protocols for the two-user cooperative communication system employing Q-DPSK modulation have been studied, for both symmetric and asymmetric inter-user channels. Recently, tight upper and lower bounds on the averaged bit error rate (BER) for full-rate differentially encoded cooperative communications have been proposed [6]. However, such bounds impose high computational complexity, i.e. an integration operator is needed, and the closed-form expression is not obtained. Furthermore, the effect of power allocation on the averaged BER of the full-rate differentially encoded cooperative communications is not fully investigated. These are the motivation of this paper. The main contributions of this paper are as follows,

- Tight approximate bounds, with low computational complexity, on the average BER is developed.
- A sub-optimum power allocation strategy for the differentially encoded cooperative communications system is studied, based on a guideline of an optimum power allocation for a coherent cooperative communication system [3]. In addition, the sub-optimum power allocation scheme is superior to the equal power allocation scheme.
- We can show that the proper cooperation strategy solely depends on a signal to noise ration (SNR) of the system.

This paper is organized as follows. In section II, the channel and system models are described. In section III, we analyze the approximate bound on the averaged BER for three different cooperation strategies, including a noncooperative strategy with B-DPSK modulation, a 1-relay cooperation strategy with Q-DPSK modulation, and a 2-relay cooperation strategy with 8-DPSK modulation. In section IV, the sub-optimum power allocation is presented. In section V, the simulation results and discussion are presented. Finally, a conclusion remark is given in section VI.

II. Channel and System Models

For the sake of exposition, let us consider the cooperative wireless networks with three active users and one destination, i.e. one user acts as a source node and the remaining two users act as relay nodes, as illustrated in Fig. 1.

In addition, we consider flat Rayleigh fading channels in this study. We also assume that all users transmit their signal through orthogonal channels by using time division multiple access (TDMA), in which the transmission is divided into two phases [1]. In phase I, user 1 acts as a source node and transmits its modulated signal to the destination. By the broadcast nature of wireless channels, the transmitted signal is also broadcast to user 2 and user 3. In phase II, user 2 and user 3 act as relay nodes that amplify and forward the received signal to user 1's destination, respectively.

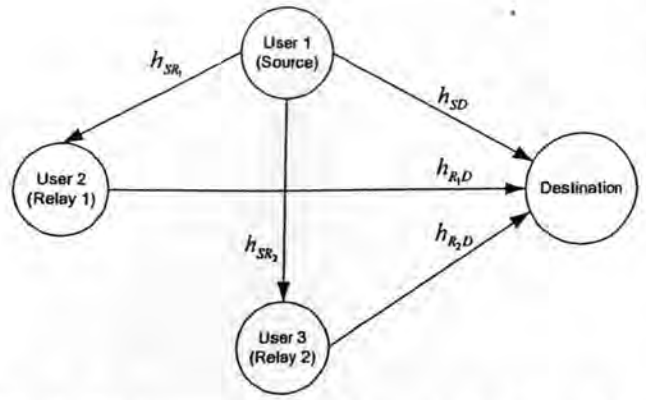

Fig. 1: The cooperative wireless networks with three active users and one destination.

Since the full-rate differentially encoded cooperation system is considered in this paper, we employ the differential modulation scheme with adaptive modulation in order to maintain the full rate, i.e. rate 1. Specifically, the B-DPSK modulation is used for direct transmission, i.e. a noncooperation strategy. For a 1-relay cooperation strategy, where either relay1 or relay 2 is employed, Q-DPSK modulation is used. Finally, for a 2-relay cooperation strategy, where both relay 1 and relay 2 are employed, 8-DPSK modulation is used. The basic principle of differential modulation is to use the previous symbol as a phase reference for the current symbol under an assumption that the channels for two consecutive information symbols are almost the same [7], i.e. slow fading channels. Hence, by employing the differential communication system, there is no need to use the channel state information (CSI) for data detection [7], resulting in the efficient communications with a penalty of 3-dB loss the probability of detection error performance. Specifically, the information bits are encoded as the differential phase between the current symbol and the previous symbol. The source node first differentially encodes the symbol *S(n)* as follows,

$$S(n) = e^{j\phi_i} S(n-1), \qquad (1)$$

where $\phi_i = \frac{2\pi i}{M}$, $i=0,1\ldots,M-1$ with M being the constellation size of the adaptive modulation and $S(n-1)$ is the previous symbol. For the sake of simplicity, we omit a time parameter n in the following. In addition, all cooperation strategies have a fixed total transmit power P per symbol. In phase I, the user 1 broadcasts the symbol S to the destination and both relay nodes. The received signal at the destination Y_{SD}, at the relay 1 Y_{SR_1}, and at the relay 2 Y_{SR_2} , can be expressed as follows, respectively,

$$Y_{SD} = \sqrt{P_1} h_{SD} S + n_{SD} \qquad (2)$$

$$Y_{SR_1} = \sqrt{P_1} h_{SR_1} S + n_{SR_1} \qquad (3)$$

$$Y_{SR_2} = \sqrt{P_1} h_{SR_2} S + n_{SR_2} \qquad (4)$$

where h_{SD}, h_{SR_1} and h_{SR_2} are the channel coefficients from source to destination, source to relay 1, and source to relay 2, respectively; n_{SD}, n_{SR_1} and n_{SR_2} are the additive white Gaussian noises with zero-mean and variance N_0 at the destination, relay 1, and relay 2, respectively; P_1 is the transmit power of the source node. In phase II, the relay 1 and relay 2 amplify the received signal and forward it to the destination with the transmit power $\tilde{P}_2$ and $\tilde{P}_3$, respectively [3],[6]. The received signal at the destination Y_{R_1D}, transmitted from the relay 1, and Y_{R_2D}, transmitted from the relay 2, can be expressed as follows,

$$Y_{R_1D} = \sqrt{\tilde{P}_2} h_{R_1D} Y_{SR_1} + n_{R_1D}, \qquad (5)$$

$$Y_{R_2D} = \sqrt{\tilde{P}_3} h_{R_2D} Y_{SR_2} + n_{R_2D}, \qquad (6)$$

$$\tilde{P}_2 = \begin{cases} \frac{P_2}{P_1 \delta^2_{SR_1} + N_0} & ; \text{If relay 1 is selected} \\ 0 & ; \text{Other} \end{cases}$$

$$\tilde{P}_3 = \begin{cases} \frac{P_3}{P_1 \delta^2_{SR_2} + N_0} & ; \text{If relay 2 is selected} \\ 0 & ; \text{Other} \end{cases}$$

where h_{R_1D} and h_{R_2D} are the channel coefficients from the relay 1 to the destination and the relay 2 to the destination, respectively; n_{R_1D} and n_{R_2D} are the additive white Gaussian noises with zero-mean and variance N_0. In the Reyleigh fading channels, the channel coefficients h_{SD}, h_{SR_1}, h_{SR_2}, h_{R_1D} and h_{R_2D} are modelled as independent zero-mean, complex Gaussian random variables with variances δ^2_{SD}, $\delta^2_{SR_1}$, $\delta^2_{SR_2}$, $\delta^2_{R_1D}$ and $\delta^2_{R_2D}$, respectively.

At the destination, all received signals are combined using the maximum ratio combining (MRC)technique [8] expressed as follows,

$$Z(n) = a_1 Y_{SD} + a_2 Y_{R_1D} + a_3 Y_{R_2D}, \qquad (7)$$

where $a_1 = \frac{P_1|h_{SD}|^2}{N_0}$, $a_2 = \frac{P_1P_2|h_{SR_1}|^2|h_{R_1D}|^2}{N_0[P_1\delta^2_{SR_1}+P_2 h_{R_1D}|^2+N_0]}$, and $a_3 = \frac{P_1P_3|h_{SR_2}|^2|h_{R_2D}|^2}{N_0[P_1\delta^2_{SR_2}+P_3 h_{R_2D}|^2+N_0]}$. Since we assume that the knowledge of channel coefficients is not available to the receiver except the channel variances, we propose to replace all instantaneous channel gain, i.e. $|h_{SD}|^2$, $|h_{SR_1}|^2$, $|h_{SR_2}|^2$, $|h_{R_1D}|^2$ and $|h_{R_2D}|^2$, in a_1, a_2, and a_3 by their channel variances, i.e. δ^2_{SD}, $\delta^2_{SR_1}$, $\delta^2_{SR_2}$, $\delta^2_{R_1D}$ and $\delta^2_{R_2D}$, respectively. Hence, we propose to linearly combine all received signals in a similar way to (7) with $a_1 = \frac{P_1\delta^2_{SD}}{N_0}$, $a_2 = \frac{P_1P_2\delta^2_{SR_1}\delta^2_{R_1D}}{N_0[P_1\delta^2_{SR_1}+P_2\delta^2_{R_1D}+N_0]}$, and $a_3 = \frac{P_1P_3\delta^2_{SR_2}\delta^2_{R_2D}}{N_0[P_1\delta^2_{SR_2}+P_3\delta^2_{R_2D}+N_0]}$. The decoder uses the sufficient statistics to decode the information symbol $\hat{S}(n)$ as follow,

$$\hat{S}(n) = \arg_{m=0,1,\ldots,M-1} \max Re[(e^{-j\phi_i})Z(n)], \qquad (8)$$

where *Re*[ ] is the real operator.

III. APPROXIMATE BOUNDS ON AVERAGED BER FOR RAYLEIGH FADING CHANNELS

In this section, we derive the approximate bound on averaged BER for the cooperation systems with the M-DPSK modulation. The analysis here is based on the MRC combiner, and then the approximate SNR of the combined output is $\gamma \approx \gamma_1+\gamma_2+\gamma_3$

[8] where $\gamma_1 = \frac{P_1|h_{SD}|^2}{N_0}$, $\gamma_2 = \frac{P_1P_2|h_{SR_1}|^2|h_{R_1D}|^2}{N_0[P_1\delta^2_{SR_1}+P_2h_{R_1D}|^2+N_0]}$, and $\gamma_3 = \frac{P_1P_3|h_{SR_2}|^2|h_{R_2D}|^2}{N_0[P_1\delta^2_{SR_2}+P_3h_{R_2D}|^2+N_0]}$. The results of this calculation serve as the performance lower bound to our proposed scheme in the previous section. However, as we will see in the simulation section, the analytical results and the simulated results are almost the same. The conditional BER of M-DPSK modulation for the cooperation system given the channel coefficients h_{SD}, h_{SR_1}, h_{SR_2}, h_{R_1D} and h_{R_2D} can be expressed as follows [9],

$$P_{(DPSK)} = \frac{\xi}{2^{2L}\pi}\int_{-\pi}^{\pi} f(\theta)\exp[-\gamma\alpha(\theta)]d(\theta), \quad (9)$$

where $\xi = 1$ when B-DPSK is employed, $\xi = \frac{4}{\log_2 M}$ with M being the modulation constellation size when M-DPSK is employed, i.e $M > 2$, and $\alpha(\theta) = \frac{b^2(1+2\beta\sin(\theta)+\beta^2)}{2}$. Here, $0 < \beta = \frac{a}{b} \leq 1$ is a constant with a and b depending on the modulation size, L is a number of independent identically distributed Rayleigh fading paths of the received signals, γ is a total SNR per bit, and $f(\theta)$ is a modulation function depending on β and L.

We can derive the averaged BER of such systems, averaged over all channels, as follows. For the noncooperation strategy with B-DPSK modulation, the averaged BER can be expressed as follows,

$$P^{Noncoop}_{(B-DPSK)} = \frac{1}{4\pi}\int_{-\pi}^{\pi} f(\theta)M\gamma_1 d(\theta), \quad (10)$$

where $f(\theta) = \frac{b^2(1-\beta^2)}{2\alpha(\theta)}$ and $M\gamma_1(\theta) = \frac{1}{1+k_{SD}(\theta)}$ with $k_{SD}(\theta) = \alpha(\theta)\frac{P_1\delta^2_{SD}}{N_0}$ [4] being the moment generating function (MGF) of γ_1. We can derive the approximate upper bound on the averaged BER in (10) by ignoring all 1's in the denominator of $M\gamma_1$. For the noncooperation strategy with B-DPSK modulation, the approximate bound on the averaged BER can be written as,

$$P^{Noncoop}_{(B-DPSK)} \approx \frac{N_0}{P_1\delta^2_{SD}} I(\beta,\theta), \quad (11)$$

where $I(\beta,\theta) = \frac{1}{2\pi b^2}\int_{-\pi}^{\pi}\frac{f_B(\theta)}{(1+2\beta sin\theta+\beta^2)^2}d(\theta)$, $f_B(\theta) = (1-\beta^2)$, and $P_1 = P$. For the noncooperation strategy, we have $a = 10^{-3}$ and $b = \sqrt{2}$ [9].

For the 1-Relay cooperation strategy with Q-DPSK modulation, the averaged BER can be expressed as follows,

$$P^{Relay1,2}_{(Q-DPSK)} = \frac{1}{8\pi}\int_{-\pi}^{\pi} f(\theta)M\gamma_1 M\gamma_2 d(\theta), \quad (12)$$

where

$$f(\theta) = \frac{b^2}{2\alpha(\theta)}[(1-\beta^2)(3+\cos(2\theta)) - (\beta^{-1}-\beta^3)\sin(\theta)]$$

, $M\gamma_2(\theta) = \frac{1}{1+k_{SR_1}(\theta)}\left[1 + \frac{P_1\delta^2_{SR_1}}{P_2\delta^2_{R_1D}}\frac{k_{SR_1}(\theta)}{1+k_{SR_1}(\theta)}Z\gamma_2\right]$ denotes the MGF of γ_2, $k_{SR_1}(\theta) = \alpha(\theta)\frac{P_1\delta^2_{SR_1}}{N_0}$, $Z\gamma_2 = \int_0^\infty \exp\left(-\frac{u}{\delta^2_{R_1D}}\right)[u + R\gamma_2(\theta)]^{-1}du$, and $R\gamma_2(\theta) = \frac{P_1\delta^2_{SR_1}+N_0}{P_2}\left[1 + \frac{P_1\delta^2_{SR_1}b^2(1+\beta)^2}{2N_0}\right]^{-1}$ [4]. In addition, the $M\gamma_1$ is similar to (10). We can derive the approximate upper bound on the averaged BER in (12) by ignoring all 1's in the denominator of $M\gamma_1$ and $M\gamma_2$. For the 1-relay cooperation strategy with Q-DPSK modulation where relay 1 is selected, the approximate upper bound on the averaged BER can be written as,

$$P^{Relay1}_{(Q-DPSK)} \approx \frac{P_2\delta^2_{R_1D} + (P_1\delta^2_{SR_1}+1)Z\gamma_2}{P_1^2P_2\delta^2_{SD}\delta^2_{SR_1}\delta^2_{R_1D}}N_0^2 I(\beta,\theta), \quad (13)$$

where $I(\beta,\theta) = \frac{1}{2\pi b^4}\int_{-\pi}^{\pi}\frac{f_Q(\theta)}{(1+2\beta sin\theta+\beta^2)^3}d(\theta)$, $f_Q(\theta) = (1-\beta^2)(3+\cos(2\theta)) - (\beta^{-1}-\beta^3)\sin(\theta)$, and $P = P_1 + P_2$ [4]. Likewise, for the 1-relay cooperation strategy where the relay 2 is selected, we can show that the approximate upper bound on the averaged BER with the Q-DPSK modulation can be written as ,

$$P^{Relay2}_{(Q-DPSK)} \approx \frac{P_3\delta^2_{R_2D} + (P_1\delta^2_{SR_2}+1)Z\gamma_2}{P_1^2P_3\delta^2_{SD}\delta^2_{SR_2}\delta^2_{R_2D}}N_0^2 I(\beta,\theta), \quad (14)$$

where $Z\gamma_2 = \int_0^\infty \exp\left(-\frac{u}{\delta^2_{R_2D}}\right)[u + R\gamma_2(\theta)]^{-1}du$, $R\gamma_2(\theta) = \frac{P_1\delta^2_{SR_2}+N_0}{P_3}\left[1 + \frac{P_1\delta^2_{SR_2}b^2(1+\beta)^2}{2N_0}\right]^{-1}$ and $P = P_1 + P_3$. For 1-relay cooperation strategy, we have $a = \sqrt{2-\sqrt{2}}$ and $b = \sqrt{2+\sqrt{2}}$ [9]. For the 2-Relay cooperation strategy with 8-DPSK modulation, the averaged BER can be expressed as follows,

$$P^{2-Relay}_{(8-DPSK)} = \frac{1}{48\pi}\int_{-\pi}^{\pi} f(\theta)M\gamma_1 M\gamma_2 M\gamma_3 d(\theta) \quad (15)$$

where $f(\theta) = \frac{b^2}{2\alpha(\theta)}[10(1-\beta^2) + (5-5\beta^2-\beta^{-2}+\beta^4)\cdot\cos(2\theta) - 5(\beta^{-1}-\beta^3)\sin\theta - (\beta^{-1}-\beta^3)\sin(3\theta)]$, $M\gamma_3(\theta) = \frac{1}{1+k_{SR_2}(\theta)}\left[1 + \frac{P_1\delta^2_{SR_2}}{P_3\delta^2_{R_2D}}\frac{k_{SR_2}(\theta)}{1+k_{SR_2}(\theta)}Z\gamma_3\right]$ denotes the MGF of γ_3, $k_{SR_2}(\theta) = \alpha(\theta)\frac{P_1\delta^2_{SR_2}}{N_0}$, $Z\gamma_3 = \int_0^\infty \exp\left(-\frac{u}{\delta^2_{R_2D}}\right)[u + R\gamma_3(\theta)]^{-1}du$, and $R\gamma_3(\theta) = \frac{P_1\delta^2_{SR_2}+N_0}{P_3}\left[1 + \frac{P_1\delta^2_{SR_2}b^2(1-\beta)^2}{2N_0}\right]^{-1}$. . In addition, the $M\gamma_1$ and $M\gamma_2$ are similar to (10) and (12), respectively. We can derive the approximate upper bound on the averaged BER in (15) by ignoring all 1's in the denominator of $M\gamma_1$, $M\gamma_2$ and $M\gamma_3$. For the 2-relay cooperation strategy with 8-DPSK modulation, the approximate upper bound on BER can be written as,

$$P^{2-Relay}_{(8-DPSK)} \approx \frac{[P_3\delta^2_{R_2D} + (P_1\delta^2_{SR_2}+1)Z\gamma_3]C_0}{P_1^3P_2P_3\delta^2_{SD}\delta^2_{SR_1}\delta^2_{SR_2}\delta^2_{R_1D}\delta^2_{R_2D}}N_0^3 I(\beta,\theta), \quad (16)$$

where $I(\beta,\theta) = \frac{1}{6\pi b^6}\int_{-\pi}^{\pi}\frac{f_8(\theta)}{(1+2\beta sin\theta+\beta^2)^4}d(\theta)$, $f_8(\theta) = [10(1-\beta^2) + (5-5\beta^2-\beta^{-2}+\beta^4)\cos(2\theta) - 5(\beta^{-1}-\beta^3)\sin\theta - (\beta^{-1}-\beta^3)\sin(3\theta)]$, $C_0 = P_2\delta^2_{R_1D} + (P_1\delta^2_{SR_1}+1)Z\gamma_2$ and $P = P_1 + P_2 + P_3$. In addition, $Z\gamma_2$ is similar to 1-relay cooperation strategy where the relay 1 is selected. For 2-relay cooperation strategy, we have $a = \sqrt{2-\sqrt{2-\sqrt{2}}}$ and $b = \sqrt{2+\sqrt{2-\sqrt{2}}}$ [9].

IV. SUB-OPTIMUM POWER ALLOCATION

From (11), (13), and (16), one can see that these expressions depend on the power factors, i.e. P_1, P_2, and P_3, in a non-linear fashion. It is difficult to optimize these expressions for the optimum power allocation in a closed form manner. Hence, we need to find such sub-optimum approach to be used as a guideline for obtaining the optimum power allocation. It is well-known that the differential communication system is 3-dB worse than the coherent communication system in term of the probability of detection error. Since our system shares the same environment as in [3], in which the coherent communication systems is investigated, the diversity behavior of the probability of error of both system is the same. Hence, we propose to exploit the optimum power allocation strategy,that is based on a minimum probability of error approach, in [3] as the sub-optimum power allocation strategy for our system. This sub-optimum power allocation is only the guideline for obtaining the optimum power allocation. Then, a numerically exhaustive search is employed as the last step for obtaining such optimum power allocation. Interestingly, it has been shown in the simulation results that the sub-optimum power allocation scheme performs better than the equal power allocation scheme.

V. SIMULATION RESULTS

In this section, we perform a computer simulation to examine the differentially encoded cooperative communication system in which the B-DPSK, Q-DPSK and 8-DPSK adaptive modulation and the AF protocol are employed. The fading channel coefficients are taken from Jake's model with Doppler shift of 3.5 KHz, baud rate of 1 Mbps, and bandwidth efficiency of 1 bit/s/Hz. For fair comparison, we plot the averaged BER curves as a function of P/N_0, where P is the total transmit power, that is normalized to be 1 W. In this study, the variance of the channel links between source node and destination δ^2_{SD}, source node and relay 1 $\delta^2_{SR_1}$, and source node and relay 2 $\delta^2_{SR_2}$ are 1. The variance of the channel links between relay 1 and destination $\delta^2_{R_1D}$ is 10, and relay 2 and destination $\delta^2_{R_2D}$ is 5.

In Fig.2, we illustrate the curves of the simulated averaged BER, the upper bounds on the averaged BER in [6], and the proposed approximate bound on the averaged BER for the noncooperative scheme, 1-relay cooperative scheme, and 2-relay cooperative scheme. The proposed approximate bound on the averaged BER are loose at low SNR, but asymptotically tight to the averaged BER at the high SNR regimes.

In Fig.3, we illustrate the curves of the simulated averaged BER and the proposed approximate bound on the averaged BER with equal and sub-optimum power allocations for the noncooperative strategy , 1-relay cooperative strategy, the and the 2-relay cooperative strategy. In additional, the sub-optimum power allocation scheme is based on the guideline of an optimum power allocation in [3]. The power ration is given by $P_1 = 0.65P$ and $P_2 = 0.35P$ for the 1-relay cooperation strategy; $P_1 = 0.6P$, $P_2 = 0.15P$ and $P_3 = 0.25P$ for the 2-relay cooperation strategy. It is worth noticing that the

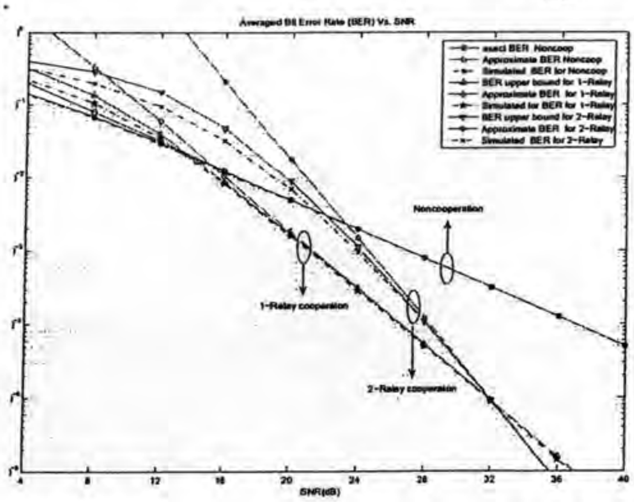

Fig. 2: The curves of the upper bounds and approximate bounds on average BER for the differentially encoded cooperative communication system with three active users.

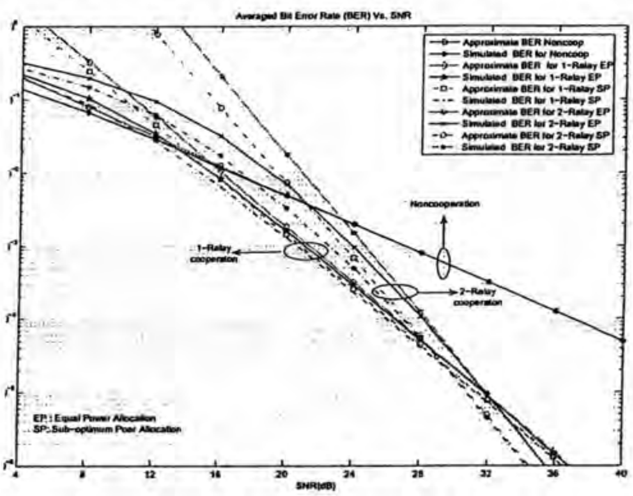

Fig. 3: The curves of the approximate bounds on averaged BER for the differentially encoded cooperative communication system with equal and sub-optimum power allocation schemes.

proposed approximate bound on BER with sub-optimum power allocation can improve the BER performance, where a SNR difference between the sub-optimum power allocation scheme and the equal power allocation scheme of 0.6 dB for the 1-relay cooperation strategy and 1.3 dB for 2-relay cooperation strategy, at BER of 10^{-4}, are observed.

The proposed approximate bound are very useful because they are tight and it requires less computational complexity for evaluation than the exact averaged BER and the upper bounds on the averaged BER in [6]. By assuming the knowledge of all channel variances at the transmitting node, we can compute this bound and use it as a performance index to determine the optimum cooperation strategy, i.e. the number of cooperative partners and the type of modulation constellation. This would be our future work.

VI. CONCLUSION

In this paper, we have analyzed the performance of the differentially encoded cooperative communication system in term of the averaged BER. We have also proposed the tight and low-computational-complexity approximate bounds on the averaged BER. In addition, the proposed approximate bounds on the averaged BER with sub-optimum power allocation strategy can improve the BER performance in comparison with the equal power allocation scheme. It is worth noticing that these proposed approximate bounds are tight to the averaged BER in fairly-high SNR regimes. In addition, with a full transmission rate constraint, it is worth noticing that the proper cooperation strategy, in the sense of the minimum probability of detection error, solely depends on the SNR of the system taking into account the channel gain, i.e. a small number of cooperative users is preferred in low SNR regimes, whereas a large number of cooperative users is preferred in high SNR regimes. Further more, the proposed approximate bound can be used as a metric for choosing the proper cooperative partner.

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ประวัติผู้เขียนวิทยานิพนธ์

นายชูชาติ มูลคำ เกิดวันที่ 27 กุมภาพันธ์ พ.ศ. 2520 ที่จังหวัดเชียงใหม่ เข้ารับการศึกษาในหลักสูตรวิศวกรรมศาสตรบัณฑิต สาขาวิศวกรรมโทรคมนาคม สถาบันเทคโนโลยีพระจอมเกล้าเจ้าคุณทหารลาดกระบัง ในปีการศึกษา 2542 สำเร็จการศึกษาปริญญาวิศวกรรมศาสตรบัณฑิต สาขาวิชาวิศวกรรมโทรคมนาคม จากสถาบันเทคโนโลยีพระจอมเกล้าเจ้าคุณทหารลาดกระบัง ในปีการศึกษา 2544 และเข้าศึกษาต่อในหลักสูตรวิศวกรรมศาสตรมหาบัณฑิต สาขาวิชาวิศวกรรมไฟฟ้า ที่จุฬาลงกรณ์มหาวิทยาลัย ในปีการศึกษา 2547