An Exact Closed-Form Expression for Bit Error Rate of Decode-and-Forward Relaying Using Selection Combining over Rayleigh Fading Channels

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Abstract: **Cooperative transmission is an effective solution to improve the performance of wireless communications over fading channels without the need for physical co-located antenna arrays. In this paper, selection combining is used at the destination instead of maximal ratio combing to optimize the structure of destination and to reduce power consumption in selective decode-and-forward relaying networks. For an arbitrary number of relays, an exact and closed-form expression of the bit error rate (BER) is derived for M-PAM, M-QAM, and M-PSK, respectively, in both independent identically distributed and independent but not identically distributed Rayleigh fading channels. A variety of simulations are performed and show that they match exactly with analytic ones. In addition, our results show that the optimum number of relays depend not only on channel conditions (operating SNRs) but also on modulation schemes which to be used.**

Index Terms: **Bit error rate (BER), decode-and-forward (DF), M-PAM, M-PSK, M-QAM, selection combining (SC).**

I. INTRODUCTION

Low power and small size are stringent design criteria for future wireless devices such as wireless sensors [1], [2]. Additionally, reliable communications is another requirement. A feasible solution to these problems is to take full advantage of idle single-antenna users, namely relays, to assist the data transmission of a certain source to its destination. This not only benefits from path-loss reduction but also creates virtual antenna arrays to gain some advantages of spatial diversity [3]–[5]. The ways the relays process signals received from a source are known as cooperative communications protocols [6]–[15].

Various protocols have been proposed to achieve the benefit from cooperative communication such as amplify-and-forward (AF), decode-and-forward (DF), and coded cooperation. In this paper, we focus on regenerative relaying (or hybrid DF [6] or selection relaying [15]). It is one of the simple cooperative communications protocols where the relay must make an independent decision on whether or not to decode and forward the source information [11]–[15]. Therefore, it avoids the noise enhancement in fixed AF relaying and remedies the decoding error retransmission in fixed DF relaying [15] (both drawbacks induced by the relay).

At the destination, the receiver can employ a variety of diversity combining techniques to obtain diversity from the multiple signal replicas from the relays and the source. Although optimum performance is highly desirable, practical systems often sacrifice some performance in order to reduce their complexity. Instead of using maximal ratio combing (MRC) which requires exact knowledge of the channel state information, a system may use selection combining (SC) which is the simplest combining method. It only selects the best signal out of all replicas for further processing and neglects all the remaining ones. This reduces the computational costs and may even lead to a better performance than MRC, because channels with very low signal-to-noise ratio (SNR) cannot be accurately estimated and contribute much noise [16]. In addition, another benefit of using SC as opposed to MRC is reduced hardware complexity at the receiver. It is appropriate for sensor networks which require fixed processing complexity at each node and reduce more cost in implementation.

In the past, relatively few work has been done on the performance analysis of DF relaying protocols with multi relays [17]–[21]. The performance is often evaluated by outage probability and bit error rate. Some previous analyzes always assume that the channels between the source, relays and destination are independent identically distributed (i.i.d.) Rayleigh. However, in real scenarios, it may be more appropriate to consider independent but not identically distributed channels. Under this condition, a closed form expression for outage probability and bit error rate of DF relaying systems that employ MRC at the destination are presented in [17], [20]. In [21], outage probability for the relaying systems that use SC at the destination is also provided for both independent but not identically distributed (i.n.d.) and i.i.d. channels. In this paper, we present exact and closed-form expressions for BER of the DF relaying protocol with an arbitrary number of relays that uses SC at the destination for both cases of i.n.d. and i.i.d. channels. These derivations are done for the system with M-PAM, M-QAM, and M-PSK modulation, respectively. In addition, we also study the impact of combining techniques on the performance of the system by comparing a system that uses SC to one that uses MRC.

The rest of this paper is organized as follows. In Section II, we introduce the model under study. Section III shows the formulas allowing for evaluation of the average BER. Section IV, we contrast the simulations and the results yielded by theory. Finally, the paper is closed in Section V.

II. SYSTEM MODEL

We consider the wireless network illustrated in Fig. 1. It is assumed that every channel between the nodes experiences slow,

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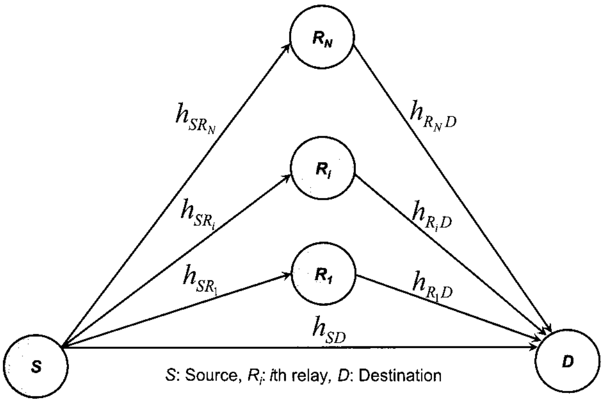

Fig. 1. Selective DF relaying model with N relays.

flat, Rayleigh fading. Due to Rayleigh fading, the channel powers, denoted by $\alpha_0 = |h_{SD}|^2$, $\alpha_{1,i} = |h_{SR_i}|^2$, and $\alpha_{2,i} = |h_{R_iD}|^2$ are independent and exponential random variables whose means are $\lambda_0, \lambda_{1,i}$, and $\lambda_{2,i}$, respectively, where $i = 1, \cdots, N$. The average transmit SNR for the source and the relays are denoted by ρ_S and ρ_{R_i} with $i = 1, \cdots, N$.

To guarantee orthogonal transmissions, we consider a time division multiple access (TDMA) arrangement with $N + 1$ time slots. However, the basic idea and operation of our proposed protocol does not depend on the specifics of the channel access protocol. In the first time slot, the source broadcasts its data to destination and N relays. At the end of the first time slot, relays will demodulate and check whether their received data are right or wrong. We define a decoding set C_D, whose members are relays which decode successfully. In real scenario, the decoding set is determined after receiving one frame from the source by employing cyclic-redundancy-check (CRC). However, in this paper, we assume that the decoding set can be decided by symbol-by-symbol for mathematical tractability of BER calculation [17]. It is obvious that C_D is a subset of $C = \{R_1, \cdots, R_i, \cdots, R_N\}$. During the following N time slots, the members of the decoding set C_D forward the source information to the destination in their respective time slots. It is assumed that the receivers at the destination and relays have perfect channel state information but no transmitter channel state information is available at the source and relays. We further assume that the bit-symbol mappings follow a Gray code.

III. PERFORMANCE ANALYSIS

We first consider the general case of independent and not identically distributed channels and then provide a compact solution for the case when the channels are assumed to be independent and identically distributed.

A. i.n.d. Fading Channels

Using the theorem on total probability, the average BER at the destination can be derived as a weighed sum of the BER for SC at the destination corresponding to each set of decoding relays C_D. Because C_D is a random set, the number of relays in each decoding set C_D is a discrete random variable n, i.e., $|C_D| = n$, $n = 0, 1, \cdots, N$. For each n, there are $\binom{N}{n}$ possible decoding subsets of size n. Thus, the average BER at the destination can be written as

$$\begin{aligned} P_b =& \Pr(C_D = \{\emptyset\}) B_D(C_D = \{\emptyset\}) \\ &+ \sum_{i_1=1}^{N} \Pr(C_D = \{R_{i_1}\}) B_D(C_D = \{R_{i_1}\}) \\ &+ \sum_{\substack{i_1,i_2=1 \\ i_1<i_2}}^{N} \begin{bmatrix} \Pr(C_D = \{R_{i_1}, R_{i_2}\}) \\ \times B_D(C_D = \{R_{i_1}, R_{i_2}\}) \end{bmatrix} \\ &+ \cdots \\ &+ \sum_{\substack{i_1,i_2,\cdots,i_n=1 \\ i_1<i_2<\cdots<i_n}}^{N} \begin{bmatrix} \Pr(C_D = \{R_{i_1}, R_{i_2}, \cdots, R_{i_n}\}) \\ \times B_D(C_D = \{R_{i_1}, R_{i_2}, \cdots, R_{i_n}\}) \end{bmatrix} \\ &+ \cdots \\ &+ \sum_{\substack{i_1,i_2,\cdots,i_N=1 \\ i_1<i_2<\cdots<i_N}}^{N} \begin{bmatrix} \Pr(C_D = \{R_{i_1}, R_{i_2}, \cdots, R_{i_N}\}) \\ \times B_D(C_D = \{R_{i_1}, R_{i_2}, \cdots, R_{i_N}\}) \end{bmatrix} \end{aligned} \quad (1)$$

where $\Pr(C_D = \{R_{i_1}, R_{i_2}, \cdots, R_{i_n}\})$ denotes the probability for decoding set C_D whose cardinality equals to n and $B_D(C_D = \{R_{i_1}, R_{i_2}, \cdots, R_{i_n}\})$ denotes the average conditional BER for the combined signal obtained by using SC after the destination received forwarded signals from the decoding set C_D as well as from the source.

The probability for decoding set C_D can be obtained by

$$\begin{aligned} \Pr(C_D) &= \left[\prod_{R_i \in C_D} (1 - S_i)\right]\left[\prod_{R_i \in C \setminus C_D} S_i\right] \\ &= \left[\prod_{R_i \in C_D} (1 - S_i)\right]\left[\prod_{R_i \notin C_D} S_i\right] \end{aligned} \quad (2)$$

where S_i denotes the average symbol error rate (SER) of modulated symbols transmitted from the source to the ith relay.

It is obvious that we totally have 2^N of decoding subsets of all size from 0 to N. Let us define a $1 \times N$ binary vector $A_k = [\, a_1^k \ \cdots \ a_i^k \ \cdots \ a_N^k \,]$ whose decimal value is k to represent the status of relays relative to the decoding set C_D, then k can take on values from 0 to $2^N - 1$ and a_i^k is defined as follows:

$$a_i^k = \begin{cases} 1, \ R_i \in C_D, \\ 0, \ R_i \notin C_D, \end{cases} i = 1, \cdots, N. \quad (3)$$

With the help of the Matlab function $de2bi(k, N)$, A_k can be calculated as

$$[a_1^k \ \cdots \ a_i^k \ \cdots \ a_N^k \,] = de2bi(k, N) \quad (4)$$

where $\sum_{i=1}^{N} a_i^k = n$. Hence, from (2)–(4), we can rewrite $\Pr(C_D)$ corresponding to each value of k as follows:

$$\Pr(C_D) = \prod_{i=1}^{N} \left[(1 - S_i)^{a_i^k} (S_i)^{1-a_i^k}\right] \quad (5)$$

Given a decoding set C_D, with SC at the destination, the signal with the largest received SNR is always selected. To simplify notation, we define a new set C'_D, which represents all nodes that are involved in the cooperative transmission, i.e., $C'_D = \{S\} \cup C_D$ and $|C'_D| = |C_D|+1 = n+1 = K$. We further define $\gamma_1, \gamma_2, \cdots, \gamma_K$ as the instantaneous SNRs of the paths received by the destination from the set C'_D with their expected values $\bar{\gamma}_1 = \lambda_0 \rho_S$, $\bar{\gamma}_2 = \lambda_{2,i_1}\rho_{R_{i_1}}, \cdots, \bar{\gamma}_K = \lambda_{2,i_{K-1}}\rho_{R_{i_{K-1}}}$, respectively. So the instantaneous SNR at the output of the selection combiner can then be expressed as

$$\gamma = \max(\gamma_1, \gamma_2, \cdots, \gamma_K). \tag{6}$$

If the branches are independently faded then order statistics gives the cumulative distribution function (CDF).

$$F_\gamma(\gamma) = P[\gamma_1 \le \gamma, \cdots, \gamma_K \le \gamma] = \prod_{j=1}^{K} F_{\gamma_j}(\gamma) \tag{7}$$

where $F_{\gamma_j}(\gamma) = P(\gamma_j \le \gamma)$ is the corresponding CDF of γ_j.

We know that when the strongest diversity branch is selected from a total K available i.n.d. diversity branches, the joint pdf of γ for K-branch SC is given by differentiating (7).

$$f_\gamma(\gamma) = \frac{\partial}{\partial\gamma}\prod_{j=1}^{K} F_{\gamma_j}(\gamma) = \sum_{j=1}^{K} f_{\gamma_j}(\gamma) \prod_{\substack{l=1 \\ l\ne j}}^{K} P(\gamma_l \le \gamma) \tag{8}$$

where for the Rayleigh fading channel case, we have

$$f_{\gamma_j}(\gamma) = \frac{1}{\bar{\gamma}_j} e^{-\gamma/\bar{\gamma}_j}, P(\gamma_l \le \gamma) = 1 - e^{-\gamma/\bar{\gamma}_l}. \tag{9}$$

Substituting (9) into (8), we obtain

$$\begin{aligned} f_\gamma(\gamma) &= \sum_{j=1}^{K} \frac{1}{\bar{\gamma}_j} e^{-\gamma/\bar{\gamma}_j} \prod_{\substack{l=1 \\ l\ne j}}^{K} (1 - e^{-\gamma/\bar{\gamma}_l}) \\ &= \sum_{j=1}^{K}\left[(-1)^{j-1} \sum_{\substack{i_1,i_2,\cdots,i_j=1 \\ i_1<i_2<\cdots<i_j}}^{K} \Gamma_j e^{-\Gamma_j \gamma}\right] \end{aligned} \tag{10}$$

where $\Gamma_j = \sum_{l=1}^{j} \bar{\gamma}_{i_l}^{-1}$. The detail of the derivation of (10) is given in Appendix I.

The average conditional bit error rate with SC at the destination for any modulation scheme in slow and flat Rayleigh fading channels can be derived by averaging bit error rate for the AWGN channel over the pdf of the SNR in Rayleigh fading.

$$B_D = \int_0^\infty \tilde{B}_D f_\gamma(\gamma) d\gamma \tag{11}$$

where $\tilde{B}_D$ is the exact bit error rate in AWGN channel, and $f_\gamma(\gamma)$ is the pdf of the instantaneous SNR per bit, γ, from K branches of the set C'_D.

A.1 M-ary PAM

For M-PAM in which $M = 2^m$ with $m = 1, 2, \cdots$, the SER from the source to the ith relay in the Rayleigh fading channel is obtained by averaging the SER for AGWN channel [22] over channel realization.

$$\begin{aligned} S_i &= \mathbb{E}\left[\frac{2(M-1)}{M} Q\left(\sqrt{\frac{6\log_2 M}{M^2-1}\rho_S \alpha_{1,i}}\right)\right] \\ &= \frac{2(M-1)}{M} I_1 \left[6\log_2 M/(M^2-1), \frac{1}{\rho_S \lambda_{1,i}}\right] \\ &= \frac{M-1}{M}\left[1 - \sqrt{\frac{\rho_S\lambda_{1,i} 6\log_2 M/(M^2-1)}{2+\rho_S\lambda_{1,i}6\log_2 M/(M^2-1)}}\right] \end{aligned} \tag{12}$$

where $\mathbb{E}(\cdot)$ is the statistical average operator, $Q(x) = \frac{1}{\pi}\int_0^{\pi/2} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta$ is defined in [23] and I_1 is defined in Appendix II.

The exact BER in the AWGN channel for M-PAM in which $M = 2^m$ with $m = 1, 2, \cdots$, is given in [24] as

$$\tilde{B}_D = \frac{1}{M\log_2 M} \sum_{u=1}^{\log_2 M} \sum_{v=0}^{(1-2^{-u})M-1} X_v^u \operatorname{erfc}(\sqrt{Y_v \gamma}) \tag{13}$$

where $X_v^u = (-1)^{\lfloor \frac{v.2^{u-1}}{M} \rfloor}(2^{u-1} - \lfloor \frac{v.2^{u-1}}{M} + \frac{1}{2} \rfloor)$, $Y_v = (2v+1)^2 3\log_2 M/(M^2-1)$, γ denotes the SNR per bit, $\lfloor x \rfloor$ denotes the largest integer to x, and erfc$(\cdot)$ is the complimentary error function. In order to obtain B_D, from (10), (11), and (13), we have [see Appendix IV]:

$$\begin{aligned} B_D &= \frac{1}{M\log_2 M}\sum_{u=1}^{\log_2 M}\sum_{v=0}^{(1-2^{-u})M-1} X_v^u \sum_{j=1}^{K}\left[\begin{array}{l}(-1)^{j-1} \sum\limits_{\substack{i_1,i_2,\cdots,i_j=1 \\ i_1<i_2<\cdots<i_j}}^{K} \\ \times \int_0^\infty \left[\begin{array}{l}\operatorname{erfc}\left(\sqrt{Y_v\gamma}\right) \\ \times \Gamma_j e^{-\Gamma_j\gamma}\end{array}\right] d\gamma\end{array}\right] \\ &= \frac{1}{M\log_2 M}\sum_{u=1}^{\log_2 M}\sum_{v=0}^{(1-2^{-u})M-1} X_v^u \sum_{j=1}^{K}\left[(-1)^{j-1}\sum_{\substack{i_1,i_2,\cdots,i_j=1 \\ i_1<i_2<\cdots<i_j}}^{K} I_3(Y_v, \Gamma_j)\right] \\ &= \frac{1}{M\log_2 M}\sum_{u=1}^{\log_2 M}\sum_{v=0}^{(1-2^{-u})M-1} X_v^u \sum_{j=1}^{K}\left[\begin{array}{l}(-1)^{j-1} \sum\limits_{\substack{i_1,i_2,\cdots,i_j=1 \\ i_1<i_2<\cdots<i_j}}^{K} \\ \times \left(1 - \sqrt{\frac{Y_v\Gamma_j^{-1}}{1+Y_v\Gamma_j^{-1}}}\right)\end{array}\right]. \end{aligned} \tag{14}$$

A.2 M-ary QAM

For M-ary QAM in which $M = 2^m$ with $m = 1, 2, \cdots$, the SER from the source to the ith relay in the Rayleigh fading channel is given as [22]

$$\begin{aligned} S_i &= \mathbb{E}\left[2pQ\left(\sqrt{q\rho_S\alpha_{1,i}}\right) - p^2 Q^2\left(\sqrt{q\rho_S\alpha_{1,i}}\right)\right] \\ &= 2pI_1\left(q, \frac{1}{\rho_S\lambda_{1,i}}\right) - p^2 I_2\left(q, \frac{1}{\rho_S\lambda_{1,i}}\right) \\ &= p\left[\begin{array}{l} 1 - \frac{p}{4} - \sqrt{\frac{q\rho_S\lambda_{1,i}}{2+q\rho_S\lambda_{1,i}}} \\ \times\left(1 - \frac{p}{\pi}\tan^{-1}\sqrt{\frac{2+q\rho_S\lambda_{1,i}}{q\rho_S\lambda_{1,i}}}\right)\end{array}\right] \end{aligned} \tag{15}$$

where $p = 2(1 - 1/\sqrt{M})$, $q = (3\log_2 M)/(M-1)$, and $Q^2(x) = \frac{1}{\pi}\int_0^{\pi/4} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta$ is defined in [23]. I_2 is further defined in Appendix III.

It is straightforward to find BER of a rectangular or square QAM if we treat it as two independent PAM constellations [22], [24]. Consider two independent PAM constellations: I-ary PAM for the in-phase component and J-ary PAM for the quadrate component, where $M = 2^m = I \times J$. The exact average BER of M-QAM in an AGWN channel is given by [24]

$$\tilde{B}_D = \frac{1}{\log_2(I\cdot J)}\left[\begin{array}{l} \frac{1}{I}\sum\limits_{u=1}^{\log_2 I}\left[\sum\limits_{l_1=0}^{(1-2^{-u})I-1} G_I(u,l_1)\mathrm{erfc}\left(\sqrt{V_u\gamma}\right)\right] + \\ \frac{1}{J}\sum\limits_{v=1}^{\log_2 J}\left[\sum\limits_{l_2=0}^{(1-2^{-v})J-1} G_J(v,l_2)\mathrm{erfc}\left(\sqrt{V_v\gamma}\right)\right]\end{array}\right] \quad (16)$$

where $G_H(x,y) = (-1)^{\left\lfloor \frac{y.2^{x-1}}{H}\right\rfloor}\left(2^{x-1} - \left\lfloor \frac{y.2^{x-1}}{H} + \frac{1}{2}\right\rfloor\right)$, $V_x = \frac{(2x+1)^2 3\log_2(I\cdot J)}{I^2+J^2-2}$ with $H \in \{I, J\}$, $x \in \{u, v\}$, and $y \in \{l_1, l_2\}$.

For this case, it is similar to the case of M-PAM, from (10), (11) and (16), the average conditional BER (B_D) for M-QAM can be obtained as (17) shown at the top of the next page where I_3 is derived in Appendix IV.

A.3 M-ary PSK

For the case of coherently detected M-PSK, to evaluate the average SER from the source to the ith relay on Rayleigh fading, we merely replace γ_S with $\log_2(M)\rho_S\alpha_{1,i}$ in [25, (5)] and use the MGF-based approach, namely

$$S_i = \frac{1}{\pi}\int_0^{(M-1)\pi/M} M_{\gamma_i}\left(-\frac{g_{\text{MPSK}}}{\sin^2\theta}\right)d\theta \quad (18)$$

where $g_{\text{MPSK}} = \sin^2(\pi/M)$ and $M_{\gamma_i}(s) = 1/(1 - s\log_2(M)\rho_S\lambda_{1,i})$ for Rayleigh fading [23]. Finally, calculating (18) gives the desired result as follows:

$$\begin{aligned} S_i &= \frac{1}{\pi}\int_0^{(M-1)\pi/M} \frac{\sin^2\theta}{\sin^2\theta + a_i}d\theta \\ &= \frac{1}{\pi}I_4\left(a_i, (M-1)\pi/M\right) \\ &= \left(\frac{M-1}{M}\right)\left[\begin{array}{l} 1 - \sqrt{\frac{a_i}{1+a_i}}\left(\frac{M}{(M-1)\pi}\right)\times \\ \left[\frac{\pi}{2} + \tan^{-1}\left(\sqrt{\frac{a_i}{1+a_i}}\cot\frac{\pi}{M}\right)\right]\end{array}\right] \end{aligned} \quad (19)$$

where $a_i = \log_2(M)g_{\text{MPSK}}\rho_S\lambda_{1,i}$ and I_4 is derived in Appendix V.

To obtain the average conditional BER B_D for M-PSK with SC on i.n.d. Rayleigh fading channels, we proceed analogous to [26].

$$B_D = \frac{1}{\log_2 M}\sum_{u=1}^{M} e_u \Pr\{\theta \in \Theta_u\} \quad (20)$$

where $\Theta_u = [\theta_L^u, \theta_U^u] = [(2u-3)\pi/M, (2u-1)\pi/M]$ for $u = 1, \cdots, M$ and e_u is the number of bit errors in the decision region Θ_u. With no loss of generality, it is assumed that $\phi = 0$, the probability $\Pr\{\theta \in \Theta_u\}$ is

$$\begin{aligned} \Pr\{\theta \in \Theta_u\} &= \int_{\theta_L^u}^{\theta_U^u}\int_0^{\infty} f_\theta(\theta\,|\phi\,,\gamma)f_\gamma(\gamma)d\gamma d\theta \\ &= \int_{\theta_L^u}^{\theta_U^u}\int_0^{\infty} f_\theta(\theta\,|\phi\,,\gamma)\sum_{j=1}^{K}\left[(-1)^{j-1}\sum_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K}\Gamma_j e^{-\Gamma_j\gamma}\right]d\gamma d\theta \\ &= \sum_{j=1}^{K}\left[(-1)^{j-1}\sum_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K}\int_{\theta_L^u}^{\theta_U^u}\int_0^{\infty}\left[\begin{array}{l} f_\theta(\theta\,|\phi\,,\gamma)\\ \times\Gamma_j e^{-\Gamma_j\gamma}\end{array}\right]d\gamma d\theta\right] \end{aligned} \quad (21)$$

where $f_\theta(\theta\,|\phi\,,\gamma)$ is defined by [26, (9b)]. By using [26, (10)], (18), and (10), we have

$$\Pr\{\theta \in \Theta_u\} = \sum_{j=1}^{K}\left[(-1)^{j-1}\sum_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K} I\left(\theta_U^u, \theta_L^u, \frac{1}{\Gamma_j}\right)\right] \quad (22)$$

where

$$I\left(\theta_U^u, \theta_L^u, \frac{1}{\Gamma_j}\right) = \frac{\theta_U^u - \theta_L^u}{2\pi} + \frac{1}{2}\left[\begin{array}{l} \beta_U^u\left(\frac{1}{2} + \frac{\tan^{-1}(\alpha_U^u)}{\pi}\right) - \\ \beta_L^u\left(\frac{1}{2} + \frac{\tan^{-1}(\alpha_L^u)}{\pi}\right)\end{array}\right] \quad (23)$$

and

$$\mu_U^u = \sqrt{\log_2 M\Gamma_j^{-1}}\sin(\theta_U^u), \quad (24a)$$

$$\mu_L^u = \sqrt{\log_2 M\Gamma_j^{-1}}\cos(\theta_L^u), \quad (24b)$$

$$\alpha_U^u = \frac{\sqrt{\log_2 M\Gamma_j^{-1}}\cos(\theta_U^u)}{\sqrt{(\mu_U^u)^2 + 1}}, \quad (24c)$$

$$\alpha_L^u = \frac{\sqrt{\log_2 M\Gamma_j^{-1}}\cos(\theta_L^u)}{\sqrt{(\mu_L^u)^2 + 1}}, \quad (24d)$$

$$\beta_U^u = \mu_U^u/\sqrt{(\mu_U^u)^2 + 1}, \quad (24e)$$

$$\beta_L^u = \mu_L^u/\sqrt{(\mu_L^u)^2 + 1}. \quad (24f)$$

B. *i.i.d. Fading Channels*

When the channels are assumed to be i.i.d, the BER of the system is obtained by simplifying (1) which can be expressed under binomial distribution. Letting $\lambda_0 = \lambda_{1,i} = \lambda_{2,i} = \lambda$, $\rho_S = \rho_{R_i} = \rho$ for $i = 1, \cdots, N$, hence $\bar{\gamma}_j = \lambda\rho = \bar{\gamma}$ for all j, it is straightforward to arrive at

$$\Pr(C_D) = (1 - S)^n (S)^{N-n} \quad (25)$$

where $S_i = S$ for $i = 1, \cdots, N$.

Furthermore, we can rewrite B_D under simplified forms for each case of M-PAM, M-QAM, and M-PSK as follows:

$$
\begin{aligned}
B_D &= \int_0^\infty \left\{ \frac{1}{\log_2(I\cdot J)} \left[\begin{array}{l} \frac{1}{I}\sum\limits_{u=1}^{\log_2 I}\left[\sum\limits_{l_1=0}^{(1-2^{-u})I-1} G_I(u,l_1)\,\mathrm{erfc}\left(\sqrt{V_u\gamma}\right)\right] + \\ \frac{1}{J}\sum\limits_{v=1}^{\log_2 J}\left[\sum\limits_{l_2=0}^{(1-2^{-v})J-1} G_J(v,l_2)\,\mathrm{erfc}\left(\sqrt{V_v\gamma}\right)\right] \end{array}\right] \times \sum_{j=1}^{K}\left[(-1)^{j-1}\sum_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K} \Gamma_j e^{-\Gamma_j\gamma}\right]\right\} d\gamma \\
&= \frac{1}{\log_2(I\cdot J)} \left\{ \begin{array}{l} \frac{1}{I}\sum\limits_{u=1}^{\log_2 I}\left[\sum\limits_{l_1=0}^{(1-2^{-u})I-1}\left[G_I(u,l_1)\sum\limits_{j=1}^{K}\left[(-1)^{j-1}\sum\limits_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K}\int_0^\infty \mathrm{erfc}(\sqrt{V_u\gamma})\Gamma_j e^{-\Gamma_j\gamma}d\gamma\right]\right]\right] + \\ \frac{1}{J}\sum\limits_{v=1}^{\log_2 J}\left[\sum\limits_{l_2=0}^{(1-2^{-v})J-1}\left[G_J(v,l_2)\sum\limits_{j=1}^{K}\left[(-1)^{j-1}\sum\limits_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K}\int_0^\infty \mathrm{erfc}(\sqrt{V_v\gamma})\Gamma_j e^{-\Gamma_j\gamma}d\gamma\right]\right]\right] \end{array}\right\} \\
&= \frac{1}{\log_2(I\cdot J)} \left\{ \begin{array}{l} \frac{1}{I}\sum\limits_{u=1}^{\log_2 I}\left[\sum\limits_{l_1=0}^{(1-2^{-u})I-1}\left[G_I(u,l_1)\sum\limits_{j=1}^{K}\left[(-1)^{j-1}\sum\limits_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K} I_3(V_u,\Gamma_j)\right]\right]\right] + \\ \frac{1}{J}\sum\limits_{v=1}^{\log_2 J}\left[\sum\limits_{l_2=0}^{(1-2^{-v})J-1}\left[G_J(v,l_2)\sum\limits_{j=1}^{K}\left[(-1)^{j-1}\sum\limits_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K} I_3(V_v,\Gamma_j)\right]\right]\right] \end{array}\right\} \\
&= \frac{1}{\log_2(I\cdot J)} \left\{ \begin{array}{l} \frac{1}{I}\sum\limits_{u=1}^{\log_2 I}\left[\sum\limits_{l_1=0}^{(1-2^{-u})I-1}\left[G_I(u,l_1)\sum\limits_{j=1}^{K}\left[(-1)^{j-1}\sum\limits_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K}\left(1-\sqrt{\frac{V_u\Gamma_j^{-1}}{1+V_u\Gamma_j^{-1}}}\right)\right]\right]\right] + \\ \frac{1}{J}\sum\limits_{v=1}^{\log_2 J}\left[\sum\limits_{l_2=0}^{(1-2^{-v})J-1}\left[G_J(v,l_2)\sum\limits_{j=1}^{K}\left[(-1)^{j-1}\sum\limits_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K}\left(1-\sqrt{\frac{V_v\Gamma_j^{-1}}{1+V_v\Gamma_j^{-1}}}\right)\right]\right]\right] \end{array}\right\}. \qquad (17)
\end{aligned}
$$

B.1 M-PAM

From (14), we have

$$B_D = \frac{1}{M\log_2 M}\sum_{u=1}^{\log_2 M}\sum_{v=0}^{(1-2^{-u})M-1} X_v^u \sum_{j=1}^{K}\left[\begin{array}{l}(-1)^{j-1}\binom{K}{j}\times \\ \left(1-\sqrt{\frac{Y_v\bar{\gamma}}{j+Y_v\bar{\gamma}}}\right)\end{array}\right]. \quad (26)$$

B.2 M-QAM

From (17), B_D of M-QAM for i.i.d channels can be reduced as (27) shown at the top of the next page.

B.3 M-PSK

From (20) and (22), B_D can be simplified as

$$B_D = \frac{1}{\log_2 M}\sum_{u=1}^{M}\left[e_u\sum_{j=1}^{K}(-1)^{j-1}\binom{K}{j} I\left(\theta_U^u,\theta_L^u,\frac{\bar{\gamma}}{j}\right)\right]. \quad (28)$$

Finally, substituting (25)-(28) into (1), we can obtain the end-to-end average bit error rate for DF relaying systems for an arbitrary number of relays with SC at the destination over i.i.d. Rayleigh channels.

IV. NUMERICAL RESULTS AND DISCUSSION

Using the analysis presented in Section III, various number of performance evaluation will be presented and compared with simulation results. For a fair of comparison, we assumed that the total transmit power is fixed as $\rho_S + \sum_{i=1}^{N}\rho_{R_i} = \rho_{DT}$ where ρ_{DT} is the average transmit signal to noise ratio of the source in case of direct transmission. In addition, for simplicity, it is assumed that the average transmit SNRs for all transmit nodes

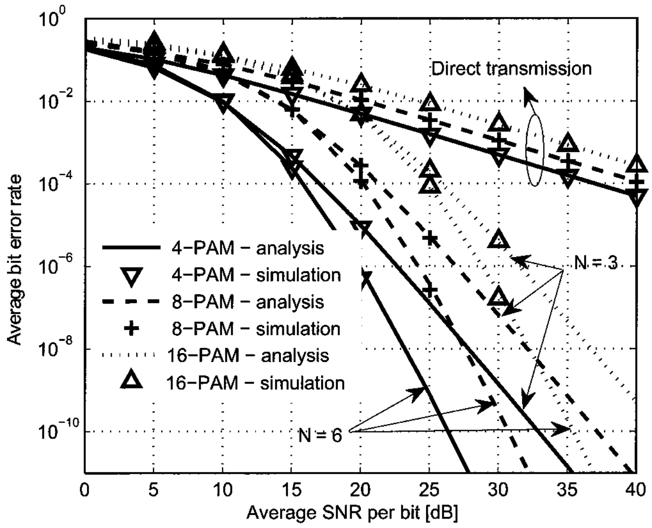

Fig. 2. Average BER for M-PAM. Channel setup: $\lambda_0 = 1$, $\{\lambda_{1,i}\}_{i=1}^{N} = 2$, and $\{\lambda_{2,i}\}_{i=1}^{N} = 3$.

$$B_D = \frac{1}{\log_2(I \cdot J)} \left\{ \begin{array}{l} \frac{1}{I} \sum\limits_{u=1}^{\log_2 I} \left[\sum\limits_{l_1=0}^{(1-2^{-u})I-1} \left[G_I(u,l_1) \sum\limits_{j=1}^{K} \left[(-1)^{j-1} \binom{K}{j} \left(1 - \sqrt{\frac{V_u \bar{\gamma}}{j+V_u \bar{\gamma}}} \right) \right] \right] \right] + \\ \frac{1}{J} \sum\limits_{v=1}^{\log_2 J} \left[\sum\limits_{l_2=0}^{(1-2^{-v})J-1} \left[G_J(v,l_2) \sum\limits_{j=1}^{K} \left[(-1)^{j-1} \binom{K}{j} \left(1 - \sqrt{\frac{V_v \bar{\gamma}}{j+V_v \bar{\gamma}}} \right) \right] \right] \right] \end{array} \right\} \tag{27}$$

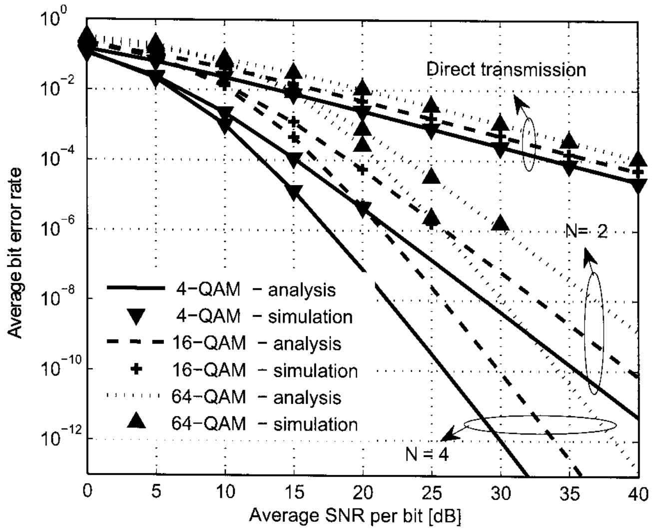

Fig. 3. Average BER for M-QAM. Channel setup: $\lambda_0 = 1$, $\{\lambda_{1,i}\}_{i=1}^{N} = 2$, and $\{\lambda_{2,i}\}_{i=1}^{N} = 3$.

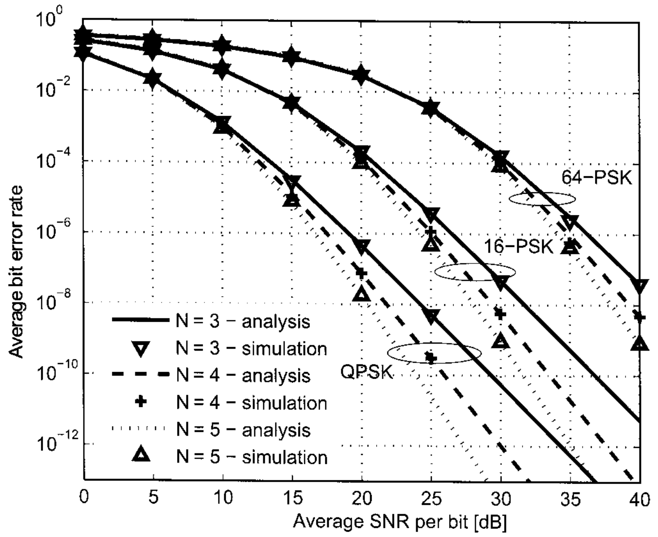

Fig. 4. Average BER for M-PSK. Channel setup: $\lambda_0 = 1$, $\{\lambda_{1,i}\}_{i=1}^{N} = 2$, and $\{\lambda_{2,i}\}_{i=1}^{N} = 3$.

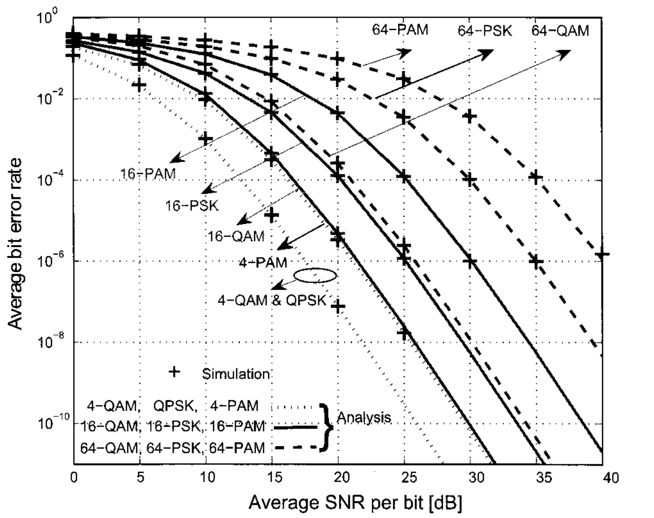

Fig. 5. Comparing the system with different modulation schemes. Number of relays: $N = 4$, channel setup: $\lambda_0 = 1$, $\{\lambda_{1,i}\}_{i=1}^{N} = 2$, and $\{\lambda_{2,i}\}_{i=1}^{N} = 3$.

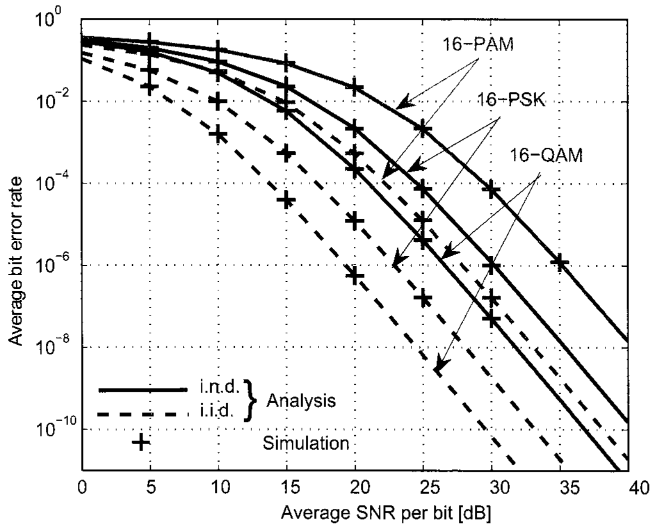

Fig. 6. BER of DF relaying with SC over i.i.d. channels ($\lambda_0 = \{\lambda_{1,i}\}_{i=1}^{N} = \{\lambda_{2,i}\}_{i=1}^{N} = 5$) and i.n.d. channels ($\lambda_0 = 1$, $\{\lambda_{1,i}\}_{i=1}^{N} = \{1.5\ 0.7\ 0.8\}$, $\{\lambda_{2,i}\}_{i=1}^{N} = \{0.5\ 1.3\ 1.2\}$), number of relays: $N = 3$.

are equal, i.e., $\rho_S = \rho_{R_1} = \cdots = \rho_{R_N} = \rho = \rho_{DT}/(N+1)$.

Figs. 2–4 show the average BER of the DF relaying system with different numbers of cooperative nodes and different modulation schemes. As shown in the figures, in high SNR regime, the improvement of the average BER is proportional to the number of relays. However with low SNR regime, using more relays could make the system performance worsen. For example, in Fig. 2, we can see that the system with 3 relays for 4-PAM outperforms that with 6 relays for 8-PAM when SNR is lower than 27 dB. It reveals that the optimal number of relays in DF relaying systems depends not only on SNRs but also on modulation schemes.

In Fig. 5, the performance of the proposed system of M-PAM, M-QAM, and M-PSK are illustrated. Among them, M-PAM

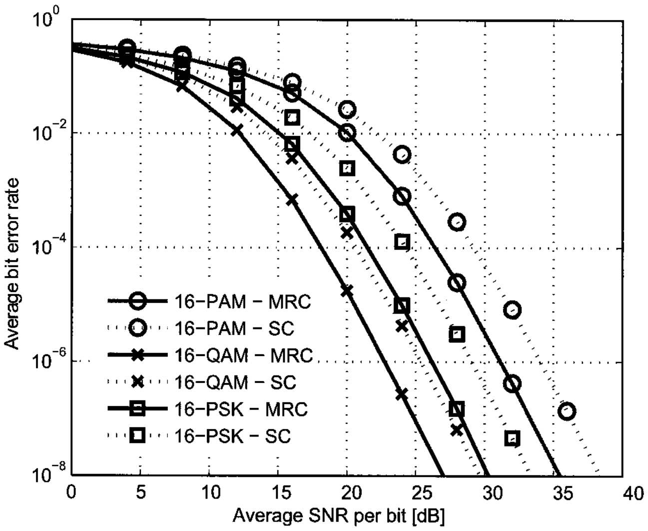

Fig. 7. BER of DF relaying with SC and MRC. Number of relays: $N = 4$, channel setup: λ_0, $\lambda_{1,i}$, and $\lambda_{2,i}$ are uniformly distributed between 0 and 1.

gives the most inferior performance, M-QAM gives the best performance and M-PSK has a performance quality in between the others. Note that with Gray code used for bit-symbol mapping, average BER for QPSK is same with that for 4-QAM.

In Fig. 6, the BER of DF relaying with 3 relays in both i.i.d and i.n.d. channels was examined. The results are based on the assumption that $\lambda_0 = \{\lambda_{1,i}\}_{i=1}^3 = \{\lambda_{2,i}\}_{i=1}^3 = 5$ for i.i.d. channels and $\lambda_0 = 1$, $\{\lambda_{1,i}\}_{i=1}^3 = \{1.5\ 0.7\ 0.8\}$ and $\{\lambda_{2,i}\}_{i=1}^3 = \{0.5\ 1.3\ 1.2\}$ for i.n.d. channels. It is seen that the performance of DF relaying systems under i.i.d. channels is better than that under i.n.d. channels. In addition, our analytical results and the simulation results are in excellent agreement.

In Fig. 7, the BER curses confirm that, under same channel conditions, the performance of systems employing MRC receiver [17] is always better as compared to equivalent systems using SC by around 1–3 dB.

V. CONCLUSION

The performance of DF relaying systems with SC diversity receiver operating over i.n.d. and i.i.d. Rayleigh fading channels has been analyzed. The exact closed expression for bit error rate has been derived. Various performance evaluation results have been also presented for verifying the analysis. Simulation results are in excellent agreement with the derived expression. The derived BER expression is general and offers a convenient way to evaluate the DF relaying systems which employs SC technique at the destination with three kinds of modulations: M-PAM, M-QAM, and M-PSK. In addition, the results also show that the loss in performance of DF systems employed SC technique is not much when compared to DF systems that use more complex MRC technique. The proposed protocol not only allows us to exploit all diversity gain offered by the channels but also reduces the complexity and power consumption of all nodes in the network. Moreover, our analysis reveals an interesting result for this relaying protocol: the optimal number of cooperative relays under average BER viewpoint is a complex function of two variables: Operating SNR and modulation scheme.

APPENDICES

The purpose of these appendices is to prove and to evaluate some equations and the integrals used in this paper.

I. PDF of SNR for the Case i.n.d. Diversity Branches in Rayleigh Fading Channel

$$\begin{aligned} f_\gamma(\gamma) &= \sum_{j=1}^{K} \frac{1}{\bar{\gamma}_j} e^{-\gamma/\bar{\gamma}_j} \prod_{\substack{l=1\\ l\neq j}}^{K} (1 - e^{-\gamma/\bar{\gamma}_l}) \\ &= \sum_{i_1=1}^{K} \bar{\gamma}_{i_1}^{-1} e^{-\gamma\bar{\gamma}_{i_1}^{-1}} \\ &+ (-1)^{2-1} \sum_{\substack{i_1,i_2=1\\ i_1<i_2}}^{K} \left[\left(\sum_{l=1}^{2} \bar{\gamma}_{i_l}^{-1}\right) e^{-\gamma\left(\sum_{l=1}^{2}\bar{\gamma}_{i_l}^{-1}\right)}\right] + \cdots \\ &+ (-1)^{j-1} \sum_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K} \left[\left(\sum_{l=1}^{j} \bar{\gamma}_{i_l}^{-1}\right) e^{-\gamma\left(\sum_{l=1}^{j}\bar{\gamma}_{i_l}^{-1}\right)}\right] + \cdots \\ &+ (-1)^{K-1} \sum_{\substack{i_1,i_2,\cdots,i_K=1\\ i_1<i_2<\cdots<i_K}}^{K} \left[\left(\sum_{l=1}^{K} \bar{\gamma}_{i_l}^{-1}\right) e^{-\gamma\left(\sum_{l=1}^{K}\bar{\gamma}_{i_l}^{-1}\right)}\right]. \end{aligned} \quad \text{(A.1)}$$

Let us define $\Gamma_j = \sum_{l=1}^{j} \bar{\gamma}_{i_l}^{-1}$, we can rewrite (A.1) as the following form:

$$f_\gamma(\gamma) = \sum_{j=1}^{K} \left[(-1)^{j-1} \sum_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{K} \Gamma_j e^{-\Gamma_j\gamma}\right]. \quad \text{(A.2)}$$

For example with $K = 2$:

$$\begin{aligned} f_\gamma(\gamma) &= \frac{1}{\bar{\gamma}_1} e^{-\frac{\gamma}{\bar{\gamma}_1}} \left(1 - e^{-\frac{\gamma}{\bar{\gamma}_2}}\right) + \frac{1}{\bar{\gamma}_2} e^{-\frac{\gamma}{\bar{\gamma}_2}} \left(1 - e^{-\frac{\gamma}{\bar{\gamma}_1}}\right) \\ &= \frac{1}{\bar{\gamma}_1} e^{-\frac{\gamma}{\bar{\gamma}_1}} + \frac{1}{\bar{\gamma}_2} e^{-\frac{\gamma}{\bar{\gamma}_2}} - \left(\frac{1}{\bar{\gamma}_1} + \frac{1}{\bar{\gamma}_2}\right) e^{-\gamma\left(\frac{1}{\bar{\gamma}_1}+\frac{1}{\bar{\gamma}_2}\right)} \\ &= (-1)^0 \sum_{i_1=1}^{2} \Gamma_1 e^{-\gamma\Gamma_1} + (-1)^1 \sum_{\substack{i_1,i_2=1\\ i_1<i_2}}^{2} \Gamma_2 e^{-\gamma\Gamma_2} \\ &= \sum_{j=1}^{2} \left[(-1)^{j-1} \sum_{\substack{i_1,i_2,\cdots,i_j=1\\ i_1<i_2<\cdots<i_j}}^{2} \Gamma_j e^{-\Gamma_j\gamma}\right] \end{aligned}$$

where $\Gamma_1 |_{i_1=1} = \bar{\gamma}_1^{-1}$, $\Gamma_1 |_{i_1=2} = \bar{\gamma}_2^{-1}$, and $\Gamma_2 \Big|_{\substack{i_1=1\\ i_2=2}} = \bar{\gamma}_1^{-1} + \bar{\gamma}_2^{-1}$.

II. Calculating $I_1(a,b)$

$$I_1(a,b) = \int_0^\infty Q\left(\sqrt{a\gamma}\right) be^{-b\gamma} d\gamma = \int_0^\infty \left(\frac{1}{\pi} \int_0^{\pi/2} e^{-\frac{a\gamma}{2\sin^2\theta}} d\theta \right) be^{-b\gamma} d\gamma \tag{B.1}$$

where $Q(x) = \frac{1}{\pi}\int_0^{\pi/2} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta$ is defined in [23]. Interchanging the order of integration and applying the result in Appendix V yields

$$\begin{aligned} I_1(a,b) &= \frac{b}{\pi} \int_0^{\pi/2} \left(\int_0^\infty e^{-\gamma\left(b+\frac{a}{2\sin^2\theta}\right)} d\gamma \right) d\theta \\ &= \frac{1}{\pi}\int_0^{\pi/2} \frac{\sin^2\theta}{\sin^2\theta + ab^{-1}/2} d\theta = \frac{1}{\pi} I_4(ab^{-1}/2, \frac{\pi}{2}) \\ &= \frac{1}{2}\left(1 - \sqrt{\frac{ab^{-1}}{2+ab^{-1}}}\right). \end{aligned} \tag{B.2}$$

III. Calculating $I_2(a,b)$

$$I_2(a,b) = \int_0^\infty Q^2\left(\sqrt{a\gamma}\right) be^{-b\gamma} d\gamma = \int_0^\infty \left(\frac{1}{\pi} \int_0^{\pi/4} e^{-\frac{a\gamma}{2\sin^2\theta}} d\theta \right) be^{-b\gamma} d\gamma \tag{C.1}$$

where $Q^2(x) = \frac{1}{\pi}\int_0^{\pi/4} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta$ is defined in [23]. Interchanging the order of integration and applying the result in Appendix V yields

$$\begin{aligned} I_2(a,b) &= \frac{b}{\pi} \int_0^{\pi/4} \left(\int_0^\infty e^{-\gamma\left(b+\frac{a}{2\sin^2\theta}\right)} d\gamma \right) d\theta \\ &= \frac{1}{\pi}\int_0^{\pi/4} \frac{\sin^2\theta}{\sin^2\theta + ab^{-1}/2} d\theta = \frac{1}{\pi} I_4(ab^{-1}/2, \frac{\pi}{4}) \\ &= \frac{1}{4}\left[1 - \sqrt{\frac{ab^{-1}}{2+ab^{-1}}} \left(\frac{4}{\pi}\tan^{-1}\sqrt{\frac{2+ab^{-1}}{ab^{-1}}}\right)\right]. \end{aligned} \tag{C.2}$$

IV. Calculating $I_3(a,b)$

$$I_3(a,b) = \int_0^\infty \operatorname{erfc}\left(\sqrt{a\gamma}\right) be^{-b\gamma} d\gamma = \int_0^\infty \left(\frac{2}{\pi} \int_0^{\pi/2} e^{-\frac{a\gamma}{\sin^2\theta}} d\theta \right) be^{-b\gamma} d\gamma \tag{D.1}$$

where $\operatorname{erfc}(x) = \frac{2}{\pi}\int_0^{\pi/2} \exp\left(-\frac{x^2}{\sin^2\theta}\right) d\theta$ is defined in [23]. Interchanging the order of integration and applying the result in Appendix V yields

$$\begin{aligned} I_3(a,b) &= \frac{2b}{\pi} \int_0^{\pi/2} \left(\int_0^\infty e^{-\gamma\left(b+\frac{a}{\sin^2\theta}\right)} d\gamma \right) d\theta \\ &= \frac{2}{\pi}\int_0^{\pi/2} \frac{\sin^2\theta}{\sin^2\theta + ab^{-1}} d\theta = \frac{2}{\pi} I_4(ab^{-1}, \frac{\pi}{2}) \\ &= 1 - \sqrt{\frac{ab^{-1}}{1+ab^{-1}}}. \end{aligned} \tag{D.2}$$

V. Calculating $I_4(c,\varphi)$

$$\begin{aligned} I_4(c,\varphi) &= \int_0^\varphi \frac{\sin^2\theta}{\sin^2\theta + c} d\theta \\ &= \int_0^\varphi \left(1 - \frac{c}{\sin^2\theta + c}\right) d\theta \\ &= \varphi - \int_0^\varphi \frac{c}{\sin^2\theta + c} d\theta \\ &= \varphi - \int_0^\varphi \frac{c(1+\tan^2\theta)}{\tan^2\theta + c(1+\tan^2\theta)} d\theta, \quad 0 < \varphi \le \frac{\pi}{2}. \end{aligned} \tag{E.1}$$

Letting change the variable $t = \tan\theta \Rightarrow d\theta = dt/(1+t^2)$

$$\begin{aligned} I_4(c,\varphi) &= \varphi - \int_0^{\tan\varphi} \frac{c}{t^2(1+c)+c} dt \\ &= \varphi - \sqrt{\frac{c}{1+c}} \tan^{-1}\left(\sqrt{\frac{1+c}{c}}\, t\right)\Bigg|_0^{\tan\varphi} \\ &= \varphi - \sqrt{\frac{c}{1+c}} \tan^{-1}\left(\sqrt{\frac{1+c}{c}} \tan\varphi\right) \\ &= \varphi - \sqrt{\frac{c}{1+c}} \left[\frac{\pi}{2} - \tan^{-1}\left(\sqrt{\frac{c}{1+c}} \cot\varphi\right)\right]. \end{aligned} \tag{E.2}$$

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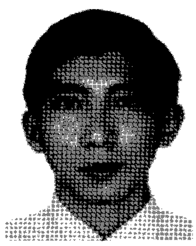

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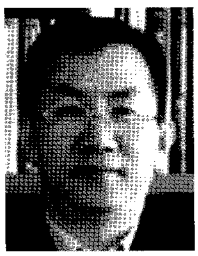

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