High-Performance Cooperative Demodulation with Decode-and-Forward Relays

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***Abstract*— Cooperative communication systems using various relay strategies can achieve spatial diversity gains, enhance coverage and potentially increase capacity. For the practically attractive decode-and-forward (DF) relay strategy, we derive a high-performance low-complexity coherent demodulator at the destination in the form of a weighted combiner. The weights are selected adaptively to account for the quality of both source-relay-destination and source-destination links. Analysis proves that the novel coherent demodulator can achieve the maximum possible diversity, regardless of the underlying constellation. Its error performance tightly bounds that of maximum-likelihood (ML) demodulation which provably quantifies the diversity gain of ML detection with DF relaying. Simulations corroborate the analysis and compare performance of the novel decoder with existing diversity-achieving strategies including analog amplify-and-forward and selective-relaying.**

***Index Terms*— User cooperation, diversity gain, full diversity, relay channel, relaying protocol, decode-and-forward.**

I. Introduction

The proliferation of wireless terminals has naturally led to cooperative (a.k.a. relay) links whereby communicators benefit from their neighbors [4], [8], [9], [11], [13], [14]. As with multi-input multi-output (MIMO) systems, where multiple collocated antennas are deployed at the transmit- or receive-ends, a main objective with single-antenna cooperating terminals is also to enable spatial diversity. Beyond MIMO with collocated antennas, relay transmissions offer resilience against shadowing and enhanced coverage. In such cooperative links, the message sent by the source arrives at the destination through diverse paths: one directly from the source node and others through *relay* nodes. Performance will thus depend on the number of cooperating relay nodes as well as the processing operations at both relays and destination. If properly designed, cooperative networks can achieve diversity order up to the number of diverse paths (what we henceforth refer to as full diversity). To put our contribution in context, we next review existing relay strategies, detectors and their error performance in terms of diversity.

If relays can afford analog processing, they can amplify-and-forward (AF) the source waveform to the destination. Unfortunately, analog AF transceivers require expensive RF chains to mitigate the coupling effects present. This motivates digital processing at relay nodes to sample and store the source waveform digitally before retransmission. Because such relays forward the decoded message to the destination, they are known as decode-and-forward (DF) relays. A third option is to have relays forward only those correctly decoded messages, in which case we say that they implement selective-relaying (SR). Use of SR presumes incorporation of e.g., cyclic redundancy check (CRC) codes from a higher layer in order to detect errors.

When it comes to performance analysis, the symbol-error-probability (SEP) for general AF links has been reported in [12], where it is also proved that full diversity is achievable with AF; see also [10]. In SR-links, a weighted superposition of the source and relay signals arriving at the destination can be formed using maximum-ratio-combining (MRC), which also collects the maximum available diversity order. Unfortunately, with DF relays MRC does not offer a full diversity achieving receiver [2]. To the best of our knowledge, no efficient demodulation technique is available which at affordable complexity can provably collect full diversity, regardless of the constellation, when using the most practical relay strategy, namely DF. And this is the gap which this paper aspires to fill.

A. Assumptions and Related Work

Being the most practical relay strategy, DF has drawn a great deal of interest recently. Although it has been shown that relay transmissions equipped with error control codes can improve error performance considerably [6], for simplicity and due to space limitations, we will focus on uncoded DF-streams and symbol-by-symbol demodulators; i.e., in the remainder of this paper, DF will refer to uncoded-DF.

Paper approved by S. N. Batalama, the Editor for Spread Spectrum and Estimation of the IEEE Communications Society. Manuscript received August 5, 2005; revised March 12, 2006 and August 3, 2006. This work was supported through collaborative participation in the Communications and Networks Consortium sponsored by the U.S. Army Research Laboratory under the Collaborative Technology Alliance Program, Cooperative Agreement DAAD19-01-2-0011. The U.S. Government is authorized to reproduce and distribute reprints for Government purposes notwithstanding any copyright notation thereon. The work of the second author was supported by the Spanish government TEC2005-06766-C03-01/TCM. The work of the fourth author was supported by NSF Grant ECS03-29766. This paper was presented in part at the 39th Asilomar Conference on Signals, Systems, and Computers, Pacific Grove, CA, November 2005.

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Publisher Item Identifier

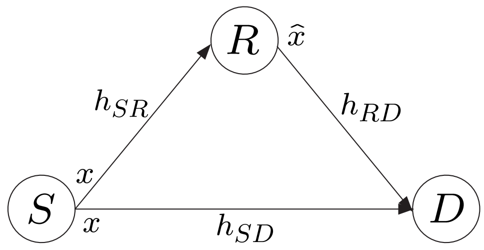

Fig. 1. Block model for a single-relay cooperative system.

Our goal is to show that, through a suitable demodulator, full diversity can be achieved with DF links. This result was also alluded to in [13], where a maximum-likelihood (ML) optimal detector has been presented only for binary phase shift keying (BPSK). As recognized in [13], performance analysis of such a detector is quite complicated, which prevents one from any quantitative diversity assessment, especially for general constellations. For this reason, a suboptimum combiner termed λ-MRC was derived in [13]. Such a combiner facilitates performance analysis and leads to closed-form expressions for the probability of error. Simulations illustrate that λ-MRC performs close to ML, but since the combiner parameter λ has not been specified analytically, full-diversity claims can not be proved. In [4], a piece-wise linear (PL) near-ML decoder has been derived for coherent and non-coherent demodulation of binary modulations only. Exploiting the *average* bit-error probability (BEP) of the source-relay link that can be made available to the destination, this PL approximation leads to closed-form bounds on error performance, concluding that with $M-1$ parallel relays, the diversity order in coherent operation is at most $(M/2)+1$ for M even, and $(M+1)/2$ for M odd.

Our idea in this paper is to exploit knowledge of the *instantaneous* (per fading realization) BEP of the source-relay link at the destination, and derive a novel combiner capable of collecting full diversity with DF for any coherent modulation and in any general (possibly multi-branch and multi-hop) relay links. Such a combiner that we term *cooperative* MRC (C-MRC), will be shown to provide a tight lower bound on the performance of ML detection; meaning that full diversity claims we will establish for this combiner carry over to the ML case too. Unlike ML and PL demodulators, it will turn out that C-MRC offers a high-performance demodulator with low-complexity *regardless* of the underlying constellation.

The rest of this paper is organized as follows. In Section II, the relay model will be stated for a single cooperating terminal, the C-MRC detector will be derived and the diversity gain of its asymptotic error probability will be analyzed; Section III will extend these results to general cooperation setups with multiple cooperating branches and multiple cooperating hops per branch; in Section IV, simulations for several settings will corroborate performance claims; and finally, conclusions will be drawn in Section V.

Notation: $(\cdot)^*$ denotes conjugation; $\mathcal{CN}(0,\sigma^2)$ denotes the circular symmetric complex Gaussian distribution with zero mean and variance σ^2; $Re\{z\}$ denotes the real part of a complex number z; for a random variable γ, $p(\gamma)$ denotes its probability density function and $\bar{\gamma}=E\{\gamma\}$ denotes its mean.

II. Single Relay Cooperation

A. System Model

With reference to Fig. 1, let us consider first the simplest model in which only one relay (R) helps the source (S) to communicate with the destination (D). The source S broadcasts symbols to D that are also received by R, which forwards the decoded symbols to D; because signals from S and R arrive through two different paths, if fading is present, one can (at least in principle) design a detector capable of collecting diversity up to order two.

In practice, terminals cannot transmit and receive at the same time and over the same frequency band; however, S and R can transmit over orthogonal channels. In this paper, we suppose a time division duplex (TDD) mode, where data transmission consists of two slots. In *Slot I*, S broadcasts modulated symbol x with average power P_x. The received symbols at R and D are

$$y_{SR} = h_{SR}x + z_{SR}, \tag{1}$$
$$y_{SD} = h_{SD}x + z_{SD}, \tag{2}$$

where h_{SR} and h_{SD} denote the fading coefficients from S to R and D, modelled as $h_{SR}\sim\mathcal{CN}(0,\sigma_{SR}^2)$, $h_{SD}\sim\mathcal{CN}(0,\sigma_{SD}^2)$, with $\sigma_{SR}^2 := E\{|h_{SR}|^2\}$ and $\sigma_{SD}^2 := E\{|h_{SD}|^2\}$, respectively. Without loss of generality, we assume that the noise terms z_{SR} and z_{SD} have equal variances N_0 and are modelled as: $z_{SR}\sim\mathcal{CN}(0,N_0)$, $z_{SD}\sim\mathcal{CN}(0,N_0)$. In the uncoded DF protocol, R performs coherent ML demodulation

$$\hat{x}_R = \arg\min_{x\in\mathcal{A}_x} |y_{SR}-h_{SR}x\,|^2,$$

where $|\mathcal{A}_x| = \Theta$ denotes the cardinality (size) of the Θ-ary constellation. Then the detected symbol $\hat{x}_R$ is re-modulated and subsequently transmitted during *Slot II* with the same average power P_x. The received symbol at D is

$$y_{RD} = h_{RD}\hat{x}_R + z_{RD}, \tag{3}$$

where $\hat{x}_R$ is the re-modulated symbol at R, h_{RD} denotes the channel coefficient from R to D, $h_{RD}\sim\mathcal{CN}(0,\sigma_{RD}^2)$ with $\sigma_{RD}^2 := E\{|h_{RD}|^2\}$, and $z_{RD}\sim\mathcal{CN}(0,N_0)$ denotes the noise at D.

B. Relay Link Analysis

Define the instantaneous signal-to-noise-ratio (SNR) at links $S-R$, $R-D$ and $S-D$ as $\gamma_{SR} := |h_{SR}|^2\bar{\gamma}$, $\gamma_{RD} := |h_{RD}|^2\bar{\gamma}$, and $\gamma_{SD} := |h_{SD}|^2\bar{\gamma}$, respectively, with $\bar{\gamma}=P_x/N_0$ denoting average SNR. Notice also that errors at the destination occur either when the $S-R$ transmission is received correctly and the $R-D$ transmission is received in error; or, when the $S-R$ transmission is received in error and the $R-D$ transmission is received correctly. Hence, for any given modulation the two-hop $S-R-D$ channel has end-to-end BEP given by

$$\begin{aligned} P^b_{\text{eq}}(\gamma_{SR},\gamma_{RD}) &= [1-P^b_{SR}(\gamma_{SR})]P^b_{RD}(\gamma_{RD}) \\ &+ [1-P^b_{RD}(\gamma_{RD})]P^b_{SR}(\gamma_{SR}), \end{aligned} \tag{4}$$

where $P^b_{SR}(\gamma_{SR})$ and $P^b_{RD}(\gamma_{RD})$ are the conditional BEPs at both hops which we assume available at D [5]. Due to possible errors at the relay, the $S-R-D$ channel is clearly nonlinear and non-Gaussian. However, one can think of the BEP in (4) as the error probability at the receiver of an *equivalent* one-hop AWGN link whose output SNR γ_{eq} is

$$\gamma_{\text{eq}} := \frac{1}{\alpha}\left\{Q^{-1}[P^b_{\text{eq}}(\gamma_{SR},\gamma_{RD})]\right\}^2, \tag{5}$$

where $Q(x) := (1/\sqrt{2\pi})\int_x^\infty \exp(-t^2/2)dt$, and α is a constant which depends on the underlying constellation; e.g., $\alpha = 2$ for BPSK. This equivalent one-hop SNR γ_{eq} jointly accounts for the quality of both $S-R$ and $R-D$ links and will guide the design of our novel demodulator. To this end, we establish the following property for γ_{eq} (see Appendix A for the proof):

Property 1: *Upon defining $\gamma_{\min} := \min\{\gamma_{RD},\gamma_{SR}\}$, it holds that γ_{eq} in (5) is bounded by:*

$$\gamma_{\min} - \frac{3.24}{\alpha} < \gamma_{\text{eq}} \le \gamma_{\min}. \tag{6}$$

Property 1 upper-bounds the end-to-end equivalent SNR by the minimum of its single-hop SNRs. This is intuitively expected, because the BEP over the aggregate $S-R-D$ link cannot exceed that of $R-D$ or $S-R$. On the other hand, the lower bound in (6) implies that for a relatively large $\gamma_{\min}$, the constant $3.24/\alpha$ can be negligible, showing that indeed $\gamma_{\min}$ can offer a tight approximation to γ_{eq}. Property 1 will come handy in our subsequent asymptotic analyses.

As recognized by [12] and [17] in the context of AF, knowledge of the $S-R$ link quality (γ_{SR}) at D is possible by sending pilot symbols through R. In regenerative schemes such as DF, one can acquire the $S-R$ link quality by sending a pilot from S for the relay to estimate the $S-R$ channel, and forward it to the destination via a second pilot whose power is scaled according to the estimated channel coefficient. At the receiver, as in AF, one again recovers the product of the two $S-R$ and $R-D$ fading coefficients. Without knowledge of the $S-R$ link at D, SR requires bandwidth consuming CRC codes to ensure perfect error detection at the relay.

C. Cooperative MRC

Consider combining the received y_{SD} and y_{RD} at the destination to obtain:

$$\hat{x}_D = \arg\min_{x\in\mathcal{A}_x} |w_{SD}y_{SD}+w_{RD}y_{RD}-(w_{SD}h_{SD}+w_{RD}h_{RD})x|^2, \tag{7}$$

where weights w_{SD} and w_{RD} are functions of h_{SD}, h_{SR}, and h_{RD} to be specified later. In a collocated multi-antenna setup, MRC employs weights $w_{SD} = h^*_{SD}$ and $w_{RD} = h^*_{RD}$, and is known to maximize the SNR at the combiner output. This would also be the optimal choice in our context if $\hat{x}_R = x$. However, since the fading link $S-R$ causes detection errors at the relay, performance of the standard MRC is far from being optimal.

Motivated by this, we fix $w_{SD} = h^*_{SD}$ to maximize γ_{SD}, and seek a weight w_{RD} to maximize the equivalent SNR γ_{eq} in the link $S-R-D$, instead of $R-D$ alone. These considerations lead to the choice

$$w_{RD}(h_{SR},h_{RD}) = \frac{\gamma_{\text{eq}}}{\gamma_{RD}}h^*_{RD}. \tag{8}$$

Combiner (7) with weights $w_{SD} = h^*_{SD}$ and w_{RD} as in (8) constitutes what we term cooperative MRC (C-MRC). Consider the product $w_{RD}h_{RD}$, and define $|h_{\text{eq}}|^2 := \gamma_{\text{eq}}/\bar{\gamma}$. If one opts for MRC, $w_{RD} = h^*_{RD}$ and thus $w_{RD}h_{RD} = |h_{RD}|^2$. However, the weight (8) modifies the last product to $w_{RD}h_{RD} = |h_{\text{eq}}|^2$, which now jointly considers the $S-R$ and $R-D$ links. Notice also from Property 1 that $\gamma_{\text{eq}} \approx \gamma_{\min}$ at sufficiently high SNR. From this approximation, w_{RD} can be seen as either part of a conventional (one-hop) MRC (when $\gamma_{SR} > \gamma_{RD}$) or as a *two-hop* weighted combiner (when $\gamma_{SR} < \gamma_{RD}$). In the first case, because the link $S-R$ is better than $R-D$, the combiner places full confidence to the arriving symbols from R. In the second case, $S-R$ is a weak link and thus the confidence is placed on the link $S-R-D$ by $|h_{\text{eq}}|^2$, instead of the link $R-D$ alone by $|h_{RD}|^2$.

Our C-MRC is reminiscent of the λ-MRC in [14], if we select $w_{RD} = \lambda h^*_{RD}$ with $0 \le \lambda \le 1$. Recall though that in lieu of a closed-form, the optimum λ in [14] is found through numerical search. Our closed-form in (8) will allow for analytical BEP evaluation. Interestingly, our simulations will show that (8) is intimately close to the optimum λ obtained by [14].

Relative to ML, the C-MRC defined by (7) and (8) is sub-optimum. To confirm the latter, it suffices to write down the ML coherent detector which for BPSK takes a relatively simple form

$$\hat{x}^{ML}_D = \arg\max_{x\in\mathcal{A}_x} \left\{\frac{1-P_{SR}(\gamma_{SR})}{2\pi N_0}\exp[-\frac{|y_{SD}-h_{SD}x|^2+|y_{RD}-h_{RD}x|^2}{2N_0}] + \frac{P_{SR}(\gamma_{SR})}{2\pi N_0}\exp[-\frac{|y_{SD}-h_{SD}x|^2+|y_{RD}+h_{RD}x|^2}{2N_0}]\right\}, \tag{9}$$

where $|\mathcal{A}_x| = 2$. An implementation of the BPSK demodulator in (9) can be found in [14]. Clearly, implementing (9) for higher-order constellations becomes prohibitively complex. Moreover, asymptotic analysis of (9) is very complicated, which prevents one from assessing ML performance [4], [14]. Assuming that the average SNR of the $S-R$ link is available at the destination, the PL-ML approximation was advocated in [4] to overcome this problem.

D. BEP of DF using BPSK

Although the C-MRC in (7) is suitable for any Θ-ary constellation, for clarity in exposition, we will first analyze its performance for BPSK. Extensions to any general constellation will be discussed in Subsection II-F. With BPSK, $\hat{x}_R$ at the relay can only take one of two values: $\hat{x}_R = x$, or,

$$\begin{aligned} P^b(\gamma_{SR}, h_{SD}, h_{RD}) \quad &= \quad [1 - P_{SR}^b(\gamma_{SR})] \int_0^\infty \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left\{ -\frac{[y + (w_{SD}h_{SD} + w_{RD}h_{RD})\sqrt{P_x}]^2}{2\sigma^2} \right\} dy \\ &+ P_{SR}^b(\gamma_{SR}) \int_0^\infty \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left\{ -\frac{[y + (w_{SD}h_{SD} - w_{RD}h_{RD})\sqrt{P_x}]^2}{2\sigma^2} \right\} dy. \end{aligned} \tag{10}$$

$\hat{x}_R = -x$. For each value, the C-MRC output is:

$$y_D = w_{RD}y_{RD} + w_{SD}y_{SD} = \begin{cases} (w_{SD}h_{SD} + w_{RD}h_{RD})x + w_{RD}z_{RD} + w_{SD}z_{SD}, & \text{if } \hat{x}_R = x, \\ (w_{SD}h_{SD} - w_{RD}h_{RD})x + w_{RD}z_{RD} + w_{SD}z_{SD}, & \text{if } \hat{x}_R = -x. \end{cases}$$

Since BPSK is real-valued, it suffices to consider only the real part $y := Re\{y_D\}$, which is a real Gaussian random variable with zero mean and variance $\sigma^2 = (|w_{RD}|^2 + |w_{SD}|^2)N_0/2$. The BEP can be expressed in terms of γ_{SR}, h_{SD}, h_{RD} as in (10) at the top of this page.

The sub-optimality of C-MRC relative to ML shows up in the second integral of (10), in which the integration region may be different from that of ML (with probability $P_{SR}^b(\gamma_{SR})$), whenever $(w_{SD}h_{SD} - w_{RD}h_{RD}) < 0$. The effect of this will be pronounced when both $w_{RD}h_{RD}$ and $P_{SR}^b(\gamma_{SR})$ are large, which happens if the link $R-D$ is strong compared to $S-R$ and $S-D$. Notice though that this sub-optimality becomes negligible as error probability in the link $S-R$ decreases. We will see that by judiciously selecting w_{RD} it will become possible to reduce error probability in this second integral. Plugging $w_{SD} = h_{SD}^*$ and (8) into (10), we can re-write compactly (10) in terms of the Q-function as

$$\begin{aligned} P^b(\gamma_{SR}, \gamma_{SD}, \gamma_{RD}) &= [1 - P_{SR}^b(\gamma_{SR})] Q\left[\frac{\sqrt{2}(\gamma_{SD} + \gamma_{\text{eq}})}{\sqrt{\gamma_{SD} + \gamma_{\text{eq}}^2/\gamma_{RD}}} \right] \\ &+ P_{SR}^b(\gamma_{SR}) Q\left[\frac{\sqrt{2}(\gamma_{SD} - \gamma_{\text{eq}})}{\sqrt{\gamma_{SD} + \gamma_{\text{eq}}^2/\gamma_{RD}}} \right]. \end{aligned} \tag{11}$$

Taking the expectation over the instantaneous SNRs, (11) yields the average BEP which we will analyze in the ensuing subsection for sufficiently large SNR values.

E. Diversity Analysis of DF Relaying

Diversity gain (diversity order) G_d is defined as the negative exponent of the average BEP plotted in a log-log scale when the average SNR tends to infinity, that is

$$P^b \overset{\bar{\gamma}\to\infty}{\approx} (G_c\bar{\gamma})^{-G_d}, \tag{12}$$

where G_c denotes the coding gain. As mentioned in the Introduction, we assume uncoded DF-streams and symbol-by-symbol demodulation; so in this case G_c depend solely on the constellation distances and the weights in (7). However, our goal in this subsection is to assess the diversity order of C-MRC by establishing bounds on the average BEP obtained from (11). To this end, we first notice that $\gamma_{\text{eq}} \leq \gamma_{RD}$, which upper-bounds (11) as

$$\begin{aligned} &P^b(\gamma_{SR}, \gamma_{SD}, \gamma_{RD}) \\ &\leq Q\left[\sqrt{2(\gamma_{SD} + \gamma_{\text{eq}})}\right] + Q[\sqrt{2\gamma_{SR}}] Q\left[\frac{\sqrt{2}(\gamma_{SD} - \gamma_{\text{eq}})}{\sqrt{\gamma_{SD} + \gamma_{\text{eq}}}} \right]. \end{aligned} \tag{13}$$

Because a sum is dominated by the term with the lowest diversity exponent, we need to prove that both terms in the right-hand side of (13) decay with the same exponent (diversity order), which here equals two. Because these two terms affect diversity through distinct means, we will analyze them separately. Let us start by defining $P_1^b(\gamma_{SR}, \gamma_{SD}, \gamma_{RD}) := Q\left[\sqrt{2(\gamma_{SD} + \gamma_{\text{eq}})}\right]$ and $(\bar{\gamma}_{SR}, \bar{\gamma}_{SD}, \bar{\gamma}_{RD}) := (\sigma_{SR}^2\bar{\gamma}, \sigma_{SD}^2\bar{\gamma}, \sigma_{RD}^2\bar{\gamma})$. Next, we invoke Property 1, use the Chernoff bound and take expectation over the three instantaneous SNRs to bound P_1^b as

$$P_1^b \leq \frac{\exp(1.62)(\bar{\gamma}_{SR} + \bar{\gamma}_{RD})}{2(\bar{\gamma}_{SD} + 1)(\bar{\gamma}_{SR}\bar{\gamma}_{RD} + \bar{\gamma}_{SR} + \bar{\gamma}_{RD})} \overset{\bar{\gamma}\to\infty}{\approx} (k_1\bar{\gamma})^{-2}, \tag{14}$$

where k_1 is a constant which depends on $\sigma_{SR}^2, \sigma_{SD}^2, \sigma_{RD}^2$. To appreciate (14), recall that $(k_1\bar{\gamma})^{-2}$ is present whenever the relay is forwarding the correct symbol ($\hat{x}_R = x$), which corresponds to the co-located multi-antenna scenario, that is capable of collecting full diversity.

Turning back our attention to (13), let us define the second summand in the bound as

$$P_2^b(\gamma_{SR}, \gamma_{SD}, \gamma_{RD}) := Q[\sqrt{2\gamma_{SR}}] Q\left[\frac{\sqrt{2}(\gamma_{SD} - \gamma_{\text{eq}})}{\sqrt{\gamma_{SD} + \gamma_{\text{eq}}}} \right]. \tag{15}$$

The next proposition upper bounds P_2^b (see Appendix B for the proof).

Proposition 1: *The expectation* $E\{P_2^b(\gamma_{SR}, \gamma_{SD}, \gamma_{RD})\} := P_2^b$ *can be bounded by a term* $\tilde{P}_2^b$*, which decays with exponent equal to two; i.e., with* k_2 *denoting a constant, we have*

$$P_2^b \leq \tilde{P}_2^b \overset{\bar{\gamma}\to\infty}{\approx} (k_2\bar{\gamma})^{-2}, \tag{16}$$

Equation (16) *together with* (14) *establish that C-MRC achieves full diversity with DF, when BPSK is used.*

One can gain insight about Proposition 1 by inspecting (15), where γ_{eq} is always smaller than γ_{SR}. An increase in γ_{eq} implies an increase of the second Q-function factor, which is mitigated by a decrease in the first Q-function factor in (15). This trade-off suggests that the use of γ_{eq} may be optimal in C-MRC to jointly account for the quality of both $S-R$ and $R-D$ links.

F. Performance Bounds for General Constellations

Following steps similar to Section II-E, we analyze here the SEP of C-MRC for higher-order constellations. Using the same notational conventions, the SEP can be expressed as the superposition of two terms, now denoted as P_1^s and P_2^s. Defining the SEP from S to R as $P_{SR}^s(\gamma_{SR})$, P_1^s corresponds to the case where, with probability $(1 - P_{SR}^s(\gamma_{SR}))$, we are

$$P_2^s(\gamma_{SR}, h_{SD}, h_{RD}) \leq 2P_{SR}^s(\gamma_{SR}) \left[\int_0^\infty \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left\{ -\frac{[y + (w_{SD}h_{SD}\tilde{d}_{\min} - w_{RD}h_{RD}\tilde{d}_{\max})]^2}{2\sigma^2} \right\} dy + \int_{-\infty}^{-2w_{SD}h_{SD}\tilde{d}_{min}} \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left\{ -\frac{[y + (w_{SD}h_{SD}\tilde{d}_{\min} - w_{RD}h_{RD}\tilde{d}_{\max})]^2}{2\sigma^2} \right\} dy \right], \quad (17)$$

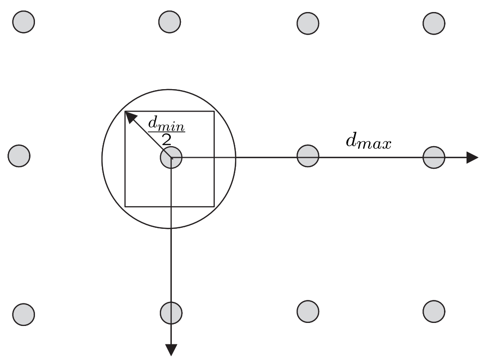

Fig. 2. Worst case in general modulation.

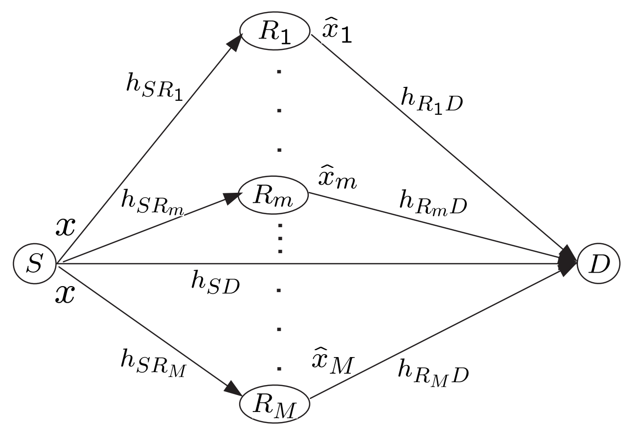

Fig. 3. Block model for a multi-branch cooperative system.

combining identical symbols arriving from S and R. Error performance of the latter is the same as in a co-located antenna system and the associated error probability is known to decay with an exponent of order two. To ensure full diversity, we also need to show that P_2^s decays with the same exponent when R is sending an erroneously decoded symbol. To prove this, we again rely on bounds of P_2^s, which happens with probability $P_{SR}^s(\gamma_{SR})$. With $d_{\min}$ ($d_{\max}$) denoting the minimum (maximum) Euclidean distance between points in the constellation, P_2^s can be bounded by the worst case which corresponds to the decoded symbol at R being at distance $d_{\max}$ from the actual symbol sent from S. In such a case, integration of the associated probability density functions (pdfs) can be simplified by reducing the decision region to a square inscribed within the circle of radius $d_{\min}$, as shown in Fig. 2. Finally, invoking the union bound, we can write $P_2^s \leq 2P_{2,1D}^s$, where $P_{2,1D}^s$ is the error probability across one-dimension, and arrive at (17) at the top of this page, where $\tilde{d}_{\min} := d_{\min}/(2\sqrt{2})$, $\tilde{d}_{\max} := d_{\max} - \tilde{d}_{\min}$, and $\sigma^2 = (|w_{RD}|^2 + |w_{SD}|^2)N_0/2$.

Knowing the BER conditioned on the instantaneous γ_{SR} for a general Θ-ary modulation, call it P_{SR}^b, we can readily upper bound the corresponding SEP as: $P_{SR}^s(\gamma_{SR}) \leq (\log_2 \Theta) P_{SR}^b(\gamma_{SR}) = (\log_2 \Theta) Q[\sqrt{\alpha\gamma_{SR}}]$. Using the latter, we can re-write (17) compactly as

$$P_2^s(\gamma_{SR}, \gamma_{SD}, \gamma_{RD}) \leq 4(\log_2 \Theta) Q[\sqrt{\alpha\gamma_{SR}}] Q\left[\frac{\sqrt{2}(\gamma_{SD} - \beta\gamma_{\text{eq}})}{\sqrt{(\gamma_{SD} + \beta\gamma_{\text{eq}})}} \right], \quad (18)$$

where $\beta := \tilde{d}_{\max}/\tilde{d}_{\min} \geq 1$ and the average transmit-SNR has also been bounded by $\bar{\gamma} \geq \tilde{d}_{\min}^2/N_0$.

We now observe that (18) has form identical (within a scale) to $P_2^b(\gamma_{SR}, \gamma_{SD}, \gamma_{RD})$ in Proposition 1. Because high SNR behavior is not affected by the constant β, our performance claims in Proposition 1 are thus still valid for (18), which establishes that our C-MRC demodulator enjoys full diversity in DF-based cooperation, regardless of the underlying constellation.

III. General Cooperation Scenarios

In this section, we generalize the results of Section II in several directions. For simplicity in exposition, we confine ourselves to BPSK, keeping in mind that as shown in Section II-F, any general constellation can be bounded using worst-case distances among points in the constellation and diversity claims can thus be mapped to those stated here for BPSK.

A. Multi-branch Cooperative Diversity

First, we generalize our model to the multi-branch cooperation setup depicted in Fig. 3. We have M relays $R_1, \ldots, R_M$, forming M branches besides the direct link $S - D$. Each R_m, $m = 1, \ldots, M$, processes symbols as R does in the single-relay case. We assume that all relays transmit over mutually orthogonal channels ($M+1$ time slots, for example). Letting $\hat{x}_m$ denote the transmitted symbol from R_m in *Slot* $m+1$, the corresponding received symbol at D is $y_{R_mD} = h_{R_mD}\hat{x}_m + z_{R_mD}$. Once again, we use C-MRC to obtain

$$\hat{x}_D = \arg\min_x |w_{SD}y_{SD} + \sum_{m=1}^{M} w_{R_mD}y_{R_mD} - (w_{SD}h_{SD} + \sum_{m=1}^{M} w_{R_mD}h_{R_mD})x|^2, \quad (19)$$

where $y_{SD} = h_{SD}x + z_{SD}$ is the symbol received directly from S. We choose $w_{SD} = h_{SD}^*$ and $w_{R_mD} = \gamma_{\text{eq}_m} h_{R_mD}^* / \gamma_{R_mD}$, with $\gamma_{\text{eq}_m} := (1/\alpha)\left\{Q^{-1}[P_{\text{eq}}(\gamma_{SR_m}, \gamma_{R_mD})]\right\}^2$, and

$$P^b(\underline{\gamma}) = \sum_{\epsilon=0}^{M}\sum_{j=1}^{\binom{M}{\epsilon}}\left\{\prod_{k=1}^{\epsilon} P^b_{SR_{E^j_k}}(\gamma_{SR_{E^j_k}})\prod_{l=1}^{M-\epsilon}[1-P^b_{SR_{C^j_l}}(\gamma_{SR_{C^j_l}})]Q\left[\frac{\sqrt{2}(\gamma_{SD}+\sum_{l=1}^{M-\epsilon}\gamma_{\text{eq}_{C^j_l}}-\sum_{k=1}^{\epsilon}\gamma_{\text{eq}_{E^j_k}})}{\sqrt{\gamma_{SD}+\sum_{m=1}^{M}(\gamma_{\text{eq}_m})^2/\gamma_{R_mD}}}\right]\right\}, \quad (20)$$

$P_{\text{eq}}(\gamma_{SR_m},\gamma_{R_mD}) = [1-P_{SR_m}(\gamma_{SR_m})]P_{R_mD}(\gamma_{R_mD}) + [1-P_{R_mD}(\gamma_{R_mD})]P_{SR_m}(\gamma_{SR_m})$.

As with (13), the conditional BEP can be expressed in terms of γ_{SR_m}, γ_{SD}, γ_{R_mD} as in (20) at the top of this page, where ϵ is used to index the nodes forwarding erroneously detected symbols $\hat{x}_R$ to D; E^j (C^j) is the set of ϵ ($M-\epsilon$) distinct elements from the set $\{1, 2, \ldots, M\}$; E^j_k (C^j_l) is the kth (lth) element of E^j (C^j), $E^j \cup C^j$=$\{1, 2, \ldots, M\}$, $E^j \cap C^j = \phi$ and for any $j \neq j'$, $E^j \neq E^{j'}$, $C^j \neq C^{j'}$; finally, $\underline{\gamma} := [\gamma_{SD}, \gamma_{SR_1}, \ldots, \gamma_{SR_M}, \gamma_{R_1D}, \ldots, \gamma_{R_MD}]$.

Notice that $\epsilon = 0$ corresponds to all relays having decoded correctly the symbol x, which is again tantamount to a MIMO system with $(M+1)$-collocated antennas where full diversity of order $M+1$ is ensured. For this reason, we only need to prove our full diversity claims for $\epsilon \geq 1$. Taking expectation over $\underline{\gamma}$ on both sides of (20), we obtain the average BEP

$$P^b = \int P^b(\underline{\gamma})p(\underline{\gamma})d\underline{\gamma} \leq \sum_{\epsilon=0}^{M}\sum_{j=1}^{\binom{M}{\epsilon}} I(\epsilon, j).$$

Considering for simplicity but without loss of generality (wlog) $\gamma_{\text{eq}_m} \overset{\bar{\gamma}\to\infty}{\approx} \min(\gamma_{SR_m}, \gamma_{R_mD}) := \gamma_{\min_m}$, we can express $I(\epsilon, j)$ as

$$I(\epsilon, j) = \int \frac{1}{2^\epsilon}\exp(-\gamma^\epsilon_{SR})Q\left[\frac{\sqrt{2}(\gamma_{SD}+\gamma^c_{\min}-\gamma^\epsilon_{\min})}{\sqrt{\gamma_{SD}+\gamma^c_{\min}+\gamma^\epsilon_{\min}}}\right]p(\underline{\gamma})d\underline{\gamma}, \quad (21)$$

where $\gamma^\epsilon_{SR} := \sum_{k=1}^{\epsilon}\gamma_{SR_{E^j_k}}$, $\gamma^c_{\min} := \sum_{l=1}^{M-\epsilon}\gamma_{\min_{C^j_l}}$ and $\gamma^\epsilon_{\min} := \sum_{k=1}^{\epsilon}\gamma_{\min_{E^j_k}}$. At this point we should clarify that at high SNR the variance $\sigma^2_{SR_m}$ does not affect the slope of $I(\epsilon, j)$ we are looking for. Let us now assume wlog that $\bar{\gamma}_{SR_m} = \bar{\gamma}_{R_mD} = \bar{\gamma}\ \forall\ m$, $\bar{\gamma}_{SD} = \bar{\gamma}/2$, and define $\gamma^\epsilon_{RD} := \sum_{k=1}^{\epsilon}\gamma_{R_{E^j_k}D}$. One can easily verify that $\gamma^\epsilon_{\min} \leq \min(\gamma^\epsilon_{SR}, \gamma^\epsilon_{RD}) := \gamma_m$, which implies that $I(\epsilon, j) \leq I(\epsilon)$, with

$$I(\epsilon) := \int_0^\infty\int_0^\infty\int_0^\infty \frac{1}{2^\epsilon}\exp(-\gamma^\epsilon_{SR})Q\left[\frac{\sqrt{2}(\gamma_s-\gamma_m)}{\sqrt{\gamma_s+\gamma_m}}\right] p(\gamma_s)p(\gamma^\epsilon_{SR})p(\gamma^\epsilon_{RD})d\gamma_s d\gamma^\epsilon_{SR} d\gamma^\epsilon_{RD}, \quad (22)$$

where we further simplified our notation by using $\gamma_s := \gamma_{SD} + \gamma^c_{\min}$. The inner most integral in (22) has the same structure encountered in Proposition 1 with the expectation taken over different pdfs. For the given SNR forms, the corresponding

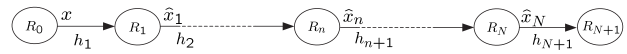

Fig. 4. Block model for a multi-hop cooperative system.

pdfs are gamma distributed; i.e.,

$$p(\gamma_s) = \frac{2^{M-\epsilon+1}(\gamma_s)^{M-\epsilon}}{(M-\epsilon)!\bar{\gamma}^{M-\epsilon+1}}\exp(-\frac{2\gamma_s}{\bar{\gamma}}), \quad (23)$$

$$p(\gamma^\epsilon_{SR}) = \frac{(\gamma^\epsilon_{SR})^{\epsilon-1}}{(\epsilon-1)!\bar{\gamma}^\epsilon}\exp(-\frac{\gamma^\epsilon_{SR}}{\bar{\gamma}}), \quad (24)$$

$$p(\gamma^\epsilon_{RD}) = \frac{(\gamma^\epsilon_{RD})^{\epsilon-1}}{(\epsilon-1)!\bar{\gamma}^\epsilon}\exp(-\frac{\gamma^\epsilon_{RD}}{\bar{\gamma}}). \quad (25)$$

Based on (23)-(25), we prove in Appendix C the following result which bounds the performance of $I(\epsilon)$ in (22).

Proposition 2: *With the pdfs in (23)-(25), $I(\epsilon)$ with $\epsilon \geq 1$ can be bounded by a term $\tilde{I}(\epsilon)$, which decays with exponent equal to $M+1$; i.e., with $k(\epsilon)$ denoting a constant, we have*

$$I(\epsilon) \leq \tilde{I}(\epsilon) \overset{\bar{\gamma}\to\infty}{\approx} [k(\epsilon)\bar{\gamma}]^{-M-1}, \quad (26)$$

Together with the case $\epsilon = 0$, eq. (26) proves that C-MRC achieves full diversity $(M+1)$ in the multi-branch cooperation scenario, where all relays utilize the DF strategy.

B. Multi-hop Cooperative Diversity

Before tackling the most general cooperative scenario with one source and one destination, we will analyze the case of N relays $R_1, \ldots, R_N$ in cascade, as depicted in Fig. 4; see also [3], [5] and [12]. We will also view this system from our end-to-end equivalent SNR perspective.

For uniformity, in notation, we rename S as R_0 and D as R_{N+1}. Each R_n, $n = 1, \ldots, N$, transmits its decoded symbol $\hat{x}_n$, while R_0 transmits $\hat{x}_0 = x$. The received symbol at each R_n, $n = 1, \ldots, N+1$ is now renamed as $y_n = h_n\hat{x}_{n-1} + z_n$, where h_n (γ_n) denotes the fading coefficient (output SNR) of the link between R_n and R_{n-1}. Again, we assume relays transmitting over mutually orthogonal channels ($N+1$ time slots, for example).

For each node R_n, we express the BEP from R_{n-1} to R_n as $P^b_{n-1,n} = Q[\sqrt{2\gamma_n}]$. The BEP P^b_n at node R_n is affected by all previous $n-1$ hops and can be iteratively calculated according to the recursion $P^b_n = (1-P^b_{n-1})P^b_{n-1,n} + (1-P^b_{n-1,n})P^b_{n-1}$, with $P^b_0 = 0$. Finally, the end-to-end BEP at the destination takes the form $P^b_{N+1} = (1-P^b_N)P^b_{N,N+1} + (1-P^b_{N,N+1})P^b_N$. Using this recursion, it is not necessary for the destination (R_{N+1}) to have centralized knowledge of all per-hop BEPs. Let us now define γ_{eq_n} to be the solution of the equation

$$P^b(\underline{\gamma})=\sum_{\epsilon=0}^{M}\sum_{j=1}^{\binom{M}{\epsilon}}\left\{\prod_{k=1}^{\epsilon}P^b_{E_k^j,N_{E_k^j}}(\gamma_{\mathrm{eq}_{E_k^j,N_{E_k^j}}})\prod_{l=1}^{M-\epsilon}[1-P^b_{C_l^j,N_{C_l^j}}(\gamma_{\mathrm{eq}_{C_l^j,N_{C_l^j}}})]Q\left[\frac{\sqrt{2}(\gamma_0+\sum_{l=1}^{M-\epsilon}\gamma_{\mathrm{eq}_{C_l^j}}-\sum_{k=1}^{\epsilon}\gamma_{\mathrm{eq}_{E_k^j}})}{\sqrt{\gamma_0+\sum_{m=1}^{M}(\gamma_{\mathrm{eq}_m})^2/\gamma_{m,N_m+1}}}\right]\right\}, \quad (27)$$

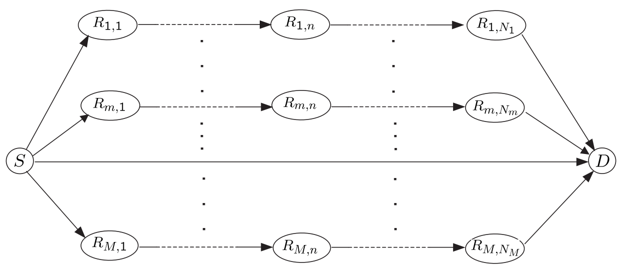

Fig. 5. Block model for multi-hop and multi-branch cooperation.

$P_n^b = Q\left[\sqrt{2\gamma_{\mathrm{eq}_n}}\right]$. We can again simplify notation by setting $\gamma_{\mathrm{eq}} := \gamma_{\mathrm{eq}_{N+1}}$. Then, the average BEP is just

$$\begin{aligned} P^b &= \int_0^\infty Q\left[\sqrt{2\gamma_{\mathrm{eq}}}\right]p(\gamma_{\mathrm{eq}})d\gamma_{\mathrm{eq}} \\ &\le \int_0^\infty \frac{1}{2}\exp(-\gamma_{\mathrm{eq}})p(\gamma_{\mathrm{eq}})d\gamma_{\mathrm{eq}}. \end{aligned} \quad (28)$$

Letting $\gamma_{\min_n} := \min(\gamma_1, \gamma_2, \ldots, \gamma_n)$, we can easily prove by induction that at high SNR $\gamma_{\mathrm{eq}_n} \overset{\bar{\gamma}\to\infty}{\approx} \gamma_{\min_n}$, since $\gamma_{\mathrm{eq}_n} \overset{\bar{\gamma}\to\infty}{\approx} \min(\gamma_{\mathrm{eq}_{n-1}}, \gamma_n)$, and the case $n = 2$ is already proven in Property 1. Clearly, we have $\gamma_{\mathrm{eq}} \overset{\bar{\gamma}\to\infty}{\approx} \gamma_{\min_{N+1}} := \gamma_{\min}$. If $\tilde{P}^b$ denotes the right hand side of (28), we can write

$$\tilde{P}^b \overset{\bar{\gamma}\to\infty}{\approx} \int_0^\infty \frac{1}{2}\exp(-\gamma_{\min})\frac{1}{\bar{\gamma}_{\min}}\exp(-\frac{\gamma_{\min}}{\bar{\gamma}_{\min}})d\gamma_{\min}.$$

Let us further consider wlog that all average SNRs are identical; i.e., we set $\bar{\gamma}_n = \bar{\gamma}$, $n = 1, \ldots, N+1$ and thus $\bar{\gamma}_{\min} = \bar{\gamma}/(N+1)$. Under these conventions, we obtain

$$P^b \le \tilde{P}^b \overset{\bar{\gamma}\to\infty}{\approx} \frac{N}{2(\bar{\gamma}+N)}. \quad (29)$$

Equation (29) shows that the diversity order of P^b is 1. We therefore confirm that a multi-hop cooperative system can only achieve diversity of order one.

C. Multi-hop, Multi-branch Cooperative Diversity

Combining results from Sections III-A and III-B, we are ready to derive the diversity gain in the BEP of multi-hop and multi-branch systems using C-MRC with DF relays. Let us consider a cooperative system with $M+1$ branches, $B_0, B_1, \ldots, B_M$, as depicted in Fig. 5. Without loss of generality, let B_0 denote the direct link. Branch B_m consists of N_m relays and $R_{m,n}$ denotes the nth node in branch B_m $\forall$ $n = 1, \ldots, N_m$ and $N_0 = 0$. As before, $R_{m,0}$ denotes S, and R_{m,N_m+1} denotes D. The received symbol at node $R_{m,n}$ is $y_{m,n} = h_{m,n}\hat{x}_{m,n-1} + z_{m,n}$, where $h_{m,n}$ ($\gamma_{m,n}$) denotes the fading coefficient (output SNR) of the link between $R_{m,n}$ and $R_{m,n-1}$. Relaxing the notation for the direct link, we write $y_0 = h_0x_0 + z_0$, where $h_0 = h_{0,1}$ and $\gamma_0 = \gamma_{0,1}$. The C-MRC detector is now

$$\begin{aligned} \hat{x}_D = \arg\min_{x\in\mathcal{A}_x} &|w_0y_0 + \sum_{m=1}^{M} w_m y_{m,N_m+1} \\ &-(w_0h_0 + \sum_{m=1}^{M} w_m h_{m,N_m+1})x\,|^2, \end{aligned} \quad (30)$$

where w_m weighs the symbol arriving from the mth branch. Letting $P^b_{m,n}$ denote the BEP at node $R_{m,n}$, one can focus on branch m and calculate the equivalent end-to-end fading coefficient $\gamma_{\mathrm{eq}_{m,n}}$ from $P^b_{m,n} = Q[\sqrt{2\gamma_{\mathrm{eq}_{m,n}}}]$, and set again $\gamma_{\mathrm{eq}_m} := \gamma_{\mathrm{eq}_{m,N_m+1}}$. The C-MRC weights here are chosen as $w_0 = h_0^*$ and $w_m = \gamma_{\mathrm{eq}_m}h^*_{m,N_m+1}/\gamma_{m,N_m+1}$.

For given $\gamma_{\mathrm{eq}_{m,N_m}}$, γ_0 and γ_{m,N_m+1}, the conditional BEP is given by (27) at the top of this page, where $\underline{\gamma} := [\gamma_0, \gamma_{1,1}, \ldots, \gamma_{1,N_1+1}, \gamma_{2,1}, \ldots, \gamma_{m,n}, \ldots, \gamma_{M,N_M+1}]$.

For the same reason as in Subsection III-A, full diversity of order $M+1$ will be collected when $\epsilon = 0$. For $\epsilon \ge 1$, taking expectations on both sides of (27), we obtain

$$P^b = \int P^b(\underline{\gamma})p(\underline{\gamma})d\underline{\gamma} \le \sum_{\epsilon=0}^{M}\sum_{j=1}^{\binom{M}{\epsilon}} J(\epsilon, j). \quad (31)$$

Given m, and since $\gamma_{\mathrm{eq}_{m,n}} \overset{\bar{\gamma}\to\infty}{\approx} \min(\gamma_{m,1}, \ldots, \gamma_{m,n}) := \gamma_{\min_{m,n}}$ $\forall$ $n = 2, \ldots, N_m+1$, we find that $J(\epsilon, j)$ for $\bar{\gamma} \to \infty$ is given by

$$J(\epsilon, j) = \int \frac{1}{2^\epsilon}\exp(-\gamma^\epsilon_{SR})Q\left[\frac{\sqrt{2}(\gamma_0 + \gamma^c_{\min} - \gamma^\epsilon_{\min})}{\sqrt{\gamma_0 + \gamma^c_{\min} + \gamma^\epsilon_{\min}}}\right]p(\underline{\gamma})d\underline{\gamma}, \quad (32)$$

where $\gamma^\epsilon_{SR} := \sum_{k=1}^{\epsilon}\gamma_{\min_{E_k^j,N_{E_k^j}}}$, $\gamma^\epsilon_{RD} := \sum_{k=1}^{\epsilon}\gamma_{E_k^j,N_{E_k^j}}$, $\gamma^\epsilon_{\min} := \sum_{k=1}^{\epsilon}\gamma_{\min_{E_k^j,N_{E_k^j}+1}}$, $\gamma^c_{\min} := \sum_{l=1}^{M-\epsilon}\gamma_{\min_{C_l^j,N_{C_l^j}+1}}$ and $\gamma^\epsilon_{\min} \le \min(\gamma^\epsilon_{SR}, \gamma^\epsilon_{RD})$.

Once arrived to (32), the remaining steps to establish the full diversity of order $M+1$ mimic those after (21) and will be thus omitted. In a nutshell, we have proved that with DF-based relays, C-MRC is full diversity achieving in general cooperation scenarios.

IV. Numerical Results and Simulations

This section presents simulated BEP tests in single-relay and multi-hop/branch scenarios for practical SNR values. Unless otherwise stated, the adopted modulation is BPSK; γ_{eq} is defined as in (5) and nodes transmit with the same power P_x, resulting in an average input SNR $\bar{\gamma} = P_x/N_0$.

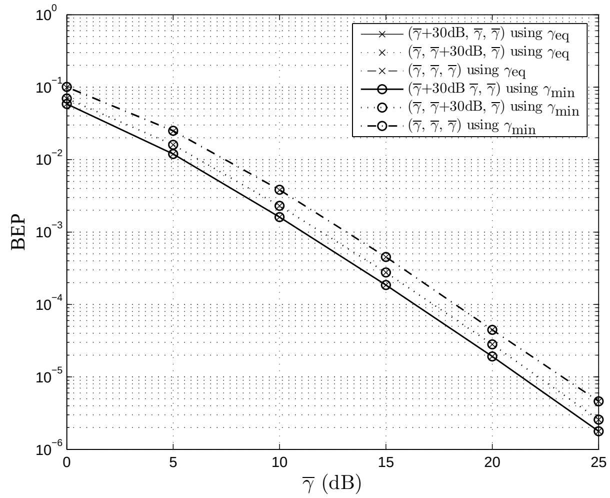

Fig. 6. Performance comparison for a single-relay system using C-MRC with BPSK: $\gamma_{\rm eq}$ vs. $\gamma_{\rm min}$.

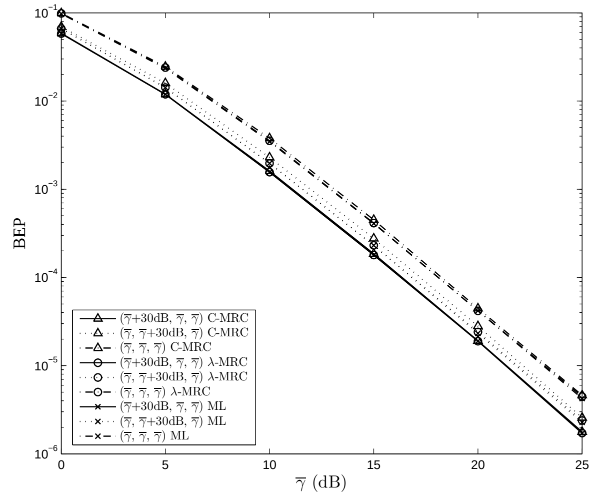

Fig. 7. Performance comparison for a single-relay system using C-MRC, λ-MRC or ML demodulation with BPSK.

A. Single Relay Cooperation

With reference to Fig. 1, we consider representative attenuation levels that correspond to those in which R is located either close to the source, close to the destination, or, equi-distant from both; the corresponding average output SNRs ($\bar\gamma_{SR}$, $\bar\gamma_{RD}$, $\bar\gamma_{SD}$) in logarithmic-scale are ($\bar\gamma$+30dB, $\bar\gamma$, $\bar\gamma$), ($\bar\gamma$, $\bar\gamma$+30dB, $\bar\gamma$), and ($\bar\gamma$, $\bar\gamma$, $\bar\gamma$), respectively.

The validity of Property 1 is tested in Fig. 6, where the $\gamma_{\rm eq}$ in (5) is substituted by $\gamma_{\rm min}$ defined in Property 1. Performance is seen to be virtually identical, which implies that the bounds we derived are tight. The SNR setup of ($\bar\gamma$+30dB, $\bar\gamma$, $\bar\gamma$) performs better than ($\bar\gamma$, $\bar\gamma$+30dB, $\bar\gamma$) because with the $S-R$ link having higher SNR, error propagation to D is mitigated. Same explanation applies also to the case ($\bar\gamma$, $\bar\gamma$, $\bar\gamma$), for which error performance is even worse because both $S-R$ and $R-D$ links have low SNR at the same time. Nevertheless, full diversity is achieved in all three settings. Fig. 7 compares bounds associated with C-MRC, λ-MRC (with numerically optimized λ as in [14]) and the ML detector in (9). As expected, C-MRC tightly bounds error performance of the rest, which as a by-product demonstrates the performance claims not proven in [14].

We next compare DF with alternative diversity-achieving strategies, namely AF and SR for block length $=100$ and CRC bits $=16$, which is assumed to be perfect. We re-iterate that the focus here is on diversity gain, which is not affected by the block length; i.e., the diversity gain of different schemes will not change when the block length changes. Based on analog processing, AF slightly outperforms DF and SR, as shown in Fig. 8. When it comes to regenerative protocols, the proposed DF decoder outperforms SR. Increasing the block length in SR, will increase this gap. However, if one decreases the block length, the spectral efficiency of SR decreases. Moreover, this can also be explained if one considers that although both DF based on C-MRC and SR are adaptive protocols achieving full diversity, the SR one is based on *hard* decisions at the relay, while DF with C-MRC exploits *soft* reliability of the $S-R-D$ link, measured by $\gamma_{\rm eq}$ at the destination.

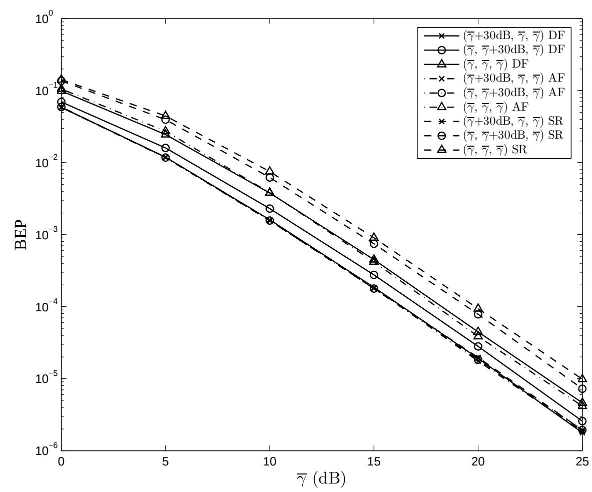

Fig. 8. Performance comparison for a single-relay system using DF-based C-MRC vs. AF vs. SR (block length=100, 16bit-CRC is assumed to be perfect), with BPSK.

To check modulations higher than BPSK, we also tested QPSK and compared C-MRC with traditional MRC. As shown in Fig. 9, diversity loss is particularly evident for low-quality $S-R$ links, which justifies the importance of link quality information which C-MRC exploits to collect full diversity.

B. Multi-branch, Multi-hop Cooperation

Here, we validate full-diversity claims for an arbitrary number of branches and relaying nodes per branch. Diversity claims are simulated when considering multiple hops per path. A fair comparison requires hops $R_1, \ldots, R_N$ to be equally spaced between S and D, which may be realistic when D is far from S [7], [12]. Assuming path-loss exponent equal to 3,

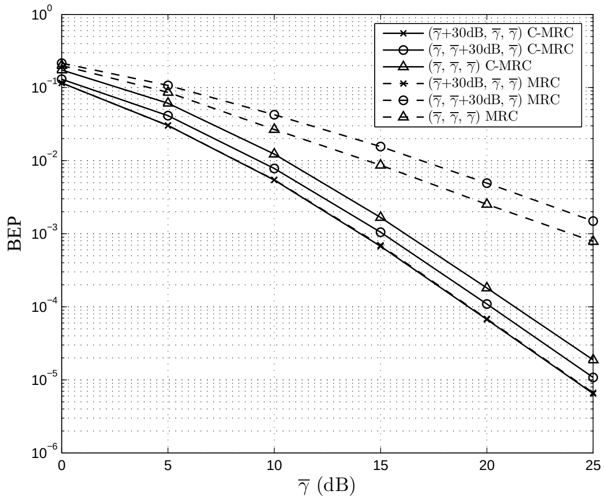

Fig. 9. Performance comparison for a single-relay system using C-MRC vs. MRC (QPSK modulation).

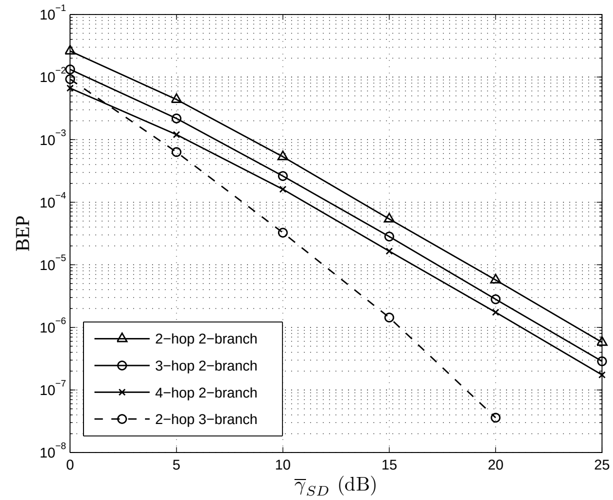

Fig. 10. Performance comparison for multi-branch and multi-hop cooperation using C-MRC with BPSK.

the average output SNR per hop is $\bar{\gamma}_n = \bar{\gamma}_{SD}(d_0/d_{n,n-1})^3$, where d_0 stands for the distance between S and D, and $d_{n,n-1}$ denotes the distance between nodes R_n and R_{n-1}. Fig. 10 shows that the full diversity is achieved regardless of the number of hops; i.e., with the same number of branches, the BEP curves have identical slopes. On the other hand, more cooperating nodes per branch improve performance in parallel BEP shifts, which is the well-known coverage enhancement effect achieved by multi-hop transmissions [5].

Nevertheless, we may compare performance if we allow these same terminals to directly communicate with the destination. The resulting BEP curve is also depicted in the same figure for comparison. We see that in this scenario, the number of arriving paths (diversity order) increases to 3, with no loss in coding gain. In view of this result, we infer that when fading is present, having all relay nodes communicate with the destination offers a desirable tradeoff between diversity and coding gains. By inspecting the slope of the BEP curves, we can also verify that the achieved full diversity order increases as M increases.

V. Conclusions

We developed a high-performance C-MRC demodulator when cooperating relays utilize the practical DF strategy. We proved that full diversity gains can be achieved with C-MRC in general cooperative links entailing any number of hops and branches, and regardless of the underlying constellation. Simulations illustrated that our C-MRC performs surprisingly close to ML while its computational simplicity relative to ML is irrespective of the constellation used. C-MRC adapts its structure depending on knowledge of the $S-R$ link quality, which is feasible to acquire through training. Relative to competing alternatives, C-MRC with DF relays outperforms existing link-adaptive regenerative strategies and comes very close to AF, which is certainly more expensive to implement in practice. [1]

Appendix

A. Proof of Property 1

We first derive an upper bound on γ_{eq}. Since $Q[x]$ is monotonically decreasing, we have $P^b_{RD}(\gamma_{RD}) = Q\left[\sqrt{\alpha\gamma_{RD}}\right] \le Q[0] = 1/2$, and $P^b_{SR}(\gamma_{SR}) = Q\left[\sqrt{\alpha\gamma_{SR}}\right] \le 1/2$. It thus follows readily from (4) that $P^b_{\text{eq}}(\gamma_{SR},\gamma_{RD}) \ge P^b_{RD}(\gamma_{RD}) = Q\left[\sqrt{\alpha\gamma_{RD}}\right]$ and $P^b_{\text{eq}}(\gamma_{SR},\gamma_{RD}) \ge P^b_{SR}(\gamma_{SR}) = Q\left[\sqrt{\alpha\gamma_{SR}}\right]$. Because $Q^{-1}[y]$ is also monotonically decreasing, we deduce that

$$\begin{aligned}\gamma_{\text{eq}} &\le \frac{1}{\alpha}\left\{\min\{\sqrt{\alpha\gamma_{RD}},\sqrt{\alpha\gamma_{SR}}\}\right\}^2\\ &= \min\{\gamma_{RD},\gamma_{SR}\} := \gamma_{\min}.\end{aligned} \tag{33}$$

We will next lower-bound γ_{eq} by $(1/\alpha)\left\{Q^{-1}\left\{Q(\sqrt{\alpha\gamma_{RD}}) + Q(\sqrt{\alpha\gamma_{SR}})\right\}\right\}^2$ using (4) and (5). As $\frac{1}{\mu\sqrt{2\pi}x}\exp(-\frac{x^2}{2}) \le Q[x] \le \frac{1}{\sqrt{2\pi}x}\exp(-\frac{x^2}{2})$, when $x \ge \sqrt{\frac{\mu}{\mu-1}}$, we infer that $Q^{-1}[y] \ge \sqrt{W_0(\frac{1}{2\mu^2y^2\pi^2})}$, when $Q^{-1}[y] \ge \sqrt{\frac{\mu}{\mu-1}}$, where $W_0(x)$ is the principal branch of the Lambert W-function [16]. It thus follows that $\gamma_{\text{eq}} \ge \frac{1}{\alpha}W_0\left[\frac{1}{2\mu^2y^2\pi^2}\right]$, where $y = Q(\sqrt{\alpha\gamma_{RD}}) + Q(\sqrt{\alpha\gamma_{SR}}) \le \frac{\exp(-\alpha\gamma_{SR}/2)}{\sqrt{2\pi}\sqrt{\alpha\gamma_{SR}}} + \frac{\exp(-\alpha\gamma_{RD}/2)}{\sqrt{2\pi}\sqrt{\alpha\gamma_{RD}}}$, when $y \le Q(\sqrt{\frac{\mu}{\mu-1}})$. Using the property of $W_0[z]$, we have $W_0[z] \ge \ln[z] - \ln[\ln[z]]$, when $z \ge \exp(1)$ [16]. Then, we

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$$\tilde{C} := \sum_{k=0}^{2M-2\epsilon+1} \Bigg\{ 2^{M-2\epsilon}(\bar{\gamma})^{-\epsilon}\left(\frac{\sqrt{\bar{\gamma}}}{2+\bar{\gamma}}\right)^{k}(2+\bar{\gamma})^{-1+\epsilon+k/2-M}\binom{2M-2\epsilon+1}{k}\frac{\Gamma[1-\epsilon-k/2+M]}{\Gamma[\epsilon]\Gamma[M-\epsilon+1]}$$
$$\left[\Gamma[\epsilon+k/2]\left(1+\frac{1}{\bar{\gamma}}+\frac{2}{2+\bar{\gamma}}\right)^{-\epsilon-k/2}+(\bar{\gamma})^{-\epsilon}\sum_{i=0}^{\epsilon-1}\left(1+\frac{1}{\bar{\gamma}}\right)^{i-\epsilon}\left(1+\frac{2}{\bar{\gamma}}+\frac{2}{2+\bar{\gamma}}\right)^{-\epsilon-k/2-i}\frac{\Gamma[\epsilon+k/2+i]}{\Gamma[i+1]}\right]\Bigg\}, \quad (34)$$

have $\gamma_{\text{eq}} \geq \frac{1}{\alpha}\{\ln[\gamma_z]-\ln[\ln[\gamma_z]]\}$, when $\gamma_{\min} \geq d/\alpha$, where

$$\gamma_z = \frac{\alpha^2\gamma_{SR}\gamma_{RD}}{\mu^2\left[\sqrt{\alpha\gamma_{SR}}\exp(-\alpha\gamma_{RD}/2)+\sqrt{\alpha\gamma_{RD}}\exp(-\alpha\gamma_{SR}/2)\right]^2},$$
$$d := \left\{Q^{-1}[\frac{1}{2}\min\{Q(\sqrt{\frac{\mu}{\mu-1}}),\sqrt{\frac{1}{2\mu^2\pi^2\exp(1)}}\}]\right\}^2.$$

With some additional manipulations, we find that $\gamma_{\text{eq}} \geq \gamma_{\min}-(1/\alpha)\left\{2\ln(2\mu)+\ln\left(\frac{d+\nu\ln 2}{d}\right)\right\}$, for any $\mu>1$ and $\nu>1$. Using numerical search to find μ and ν, we obtain a tight lower bound when $\mu=1.89, \nu=3.6$, $d=3.23157$, from which we deduce that $\gamma_{\text{eq}} \geq \gamma_{\min}-3.23165/\alpha > \gamma_{\min}-3.24/\alpha$, if $\gamma_{\min} \geq d/\alpha$. Combining this lower bound with the upper bound in (33), it follows that for any $\gamma_{\text{eq}} \geq 0$

$$\gamma_{\min}-\frac{3.24}{\alpha} < \gamma_{\text{eq}} \leq \gamma_{\min}. \quad (35)$$

B. Proof of Proposition 1

From (15) and Property 1, we obtain $P_2^b \leq A+B$, where

$$A := \int_0^\infty\int_0^\infty\int_{\gamma_{\min}}^\infty \frac{1}{2}\exp(-\gamma_{SR})\frac{1}{2}\exp\left[-\frac{(\gamma_{SD}-\gamma_{\min})^2}{\gamma_{SD}+\gamma_{\min}}\right] p(\gamma_{SD})p(\gamma_{SR})p(\gamma_{RD})d\gamma_{SD}d\gamma_{SR}d\gamma_{RD},$$
$$B := \int_0^\infty\int_0^\infty\int_0^{\gamma_{\min}}\frac{1}{2}\exp(-\gamma_{SR}) p(\gamma_{SD})p(\gamma_{SR})p(\gamma_{RD})d\gamma_{SD}d\gamma_{SR}d\gamma_{RD}.$$

For A, we have

$$A = \int_0^\infty\int_0^\infty\frac{1}{4}\exp(-\gamma_{SR})\frac{1}{\bar{\gamma}_{SR}}\exp(-\frac{\gamma_{SR}}{\bar{\gamma}_{SR}}) \frac{1}{\bar{\gamma}_{RD}}\exp(-\frac{\gamma_{RD}}{\bar{\gamma}_{RD}})\frac{1}{\bar{\gamma}_{SD}}f_A(\gamma_{\min})d\gamma_{SR}d\gamma_{RD},$$

where

$$\begin{aligned} &f_A(\gamma_{\min}) \\ =&\int_{\gamma_{\min}}^\infty \exp\left[-(\gamma_{SD}+\gamma_{\min}-\frac{4\gamma_{SD}\gamma_{\min}}{\gamma_{SD}+\gamma_{\min}})\right]\exp(-\frac{\gamma_{SD}}{\bar{\gamma}_{SD}})d\gamma_{SD} \\ \leq&\int_{\gamma_{\min}}^\infty \exp\left[-(\gamma_{SD}+\gamma_{\min}-2\sqrt{\gamma_{SD}\gamma_{\min}})\right]\exp(-\frac{\gamma_{SD}}{\bar{\gamma}_{SD}})d\gamma_{SD} \\ \leq&\frac{\bar{\gamma}_{SD}}{(1+\bar{\gamma}_{SD})^2}\exp(-\frac{\gamma_{\min}}{\bar{\gamma}_{SD}})\Big[1+\bar{\gamma}_{SD}+\sqrt{(1+\bar{\gamma}_{SD})\pi\bar{\gamma}_{SD}\gamma_{\min}}\Big]. \end{aligned}$$

After integrating over different regions and recalling that $\gamma_{\min}=\min\{\gamma_{SR},\gamma_{RD}\}$, we arrive at

$$\begin{aligned} A \leq \tilde{A} :=& \frac{\bar{\gamma}_{SD}\gamma_1}{4(1+\bar{\gamma}_{SD})(1+\bar{\gamma}_{SR})(\bar{\gamma}_{SR}\ \bar{\gamma}_{RD}+\bar{\gamma}_{SD}\gamma_1)} \\ &+\frac{\pi\bar{\gamma}_{SD}(\frac{\bar{\gamma}_{SD}}{1+\bar{\gamma}_{SD}})^{3/2}\bar{\gamma}_{SR}^3\bar{\gamma}_{RD}\sqrt{1+\frac{1}{\bar{\gamma}_{SD}}+\frac{1}{\bar{\gamma}_{SR}}+\frac{1}{\bar{\gamma}_{RD}}}}{8(1+\bar{\gamma}_{SR})\gamma_2(\bar{\gamma}_{SR}\ \bar{\gamma}_{RD}+\bar{\gamma}_{SD}\gamma_1)^2} \\ &+\frac{\bar{\gamma}_{SD}^2\bar{\gamma}_{RD}\sqrt{\bar{\gamma}_{SR}}\sqrt{1+\frac{\bar{\gamma}_{SD}\ \bar{\gamma}_{SR}}{\gamma_2\bar{\gamma}_{RD}}}\pi}{8(1+\bar{\gamma}_{SD})^{3/2}\sqrt{\gamma_2}(\bar{\gamma}_{SR}\ \bar{\gamma}_{RD}+\bar{\gamma}_{SD}\gamma_1)}, \end{aligned} \quad (36)$$

where $\gamma_1 := \bar{\gamma}_{SR}+\bar{\gamma}_{RD}+\bar{\gamma}_{SR}\ \bar{\gamma}_{RD}$ and $\gamma_2 := \bar{\gamma}_{SD}+\bar{\gamma}_{SR}+\bar{\gamma}_{SD}\ \bar{\gamma}_{SR}$.

Likewise for B, we have

$$\begin{aligned} B =& \int_0^\infty\int_0^\infty\frac{1}{2}\exp(-\gamma_{SR})\frac{1}{\bar{\gamma}_{SR}}\exp(-\frac{\gamma_{SR}}{\bar{\gamma}_{SR}}) \\ &\frac{1}{\bar{\gamma}_{RD}}\exp(-\frac{\gamma_{RD}}{\bar{\gamma}_{RD}})[1-\exp(-\frac{\gamma_{\min}}{\bar{\gamma}_{SD}})]d\gamma_{SR}d\gamma_{RD} \\ =& \frac{\bar{\gamma}_{SR}\ \bar{\gamma}_{RD}}{2(1+\bar{\gamma}_{SR})[\bar{\gamma}_{SR}\bar{\gamma}_{RD}+\bar{\gamma}_{SD}(\bar{\gamma}_{SR}+\bar{\gamma}_{RD}+\bar{\gamma}_{SR}\bar{\gamma}_{RD})]}. \end{aligned}$$

Upon defining $(\bar{\gamma}_{SR},\bar{\gamma}_{SD},\bar{\gamma}_{RD}) := (\sigma_{SR}^2\bar{\gamma},\sigma_{SD}^2\bar{\gamma},\sigma_{RD}^2\bar{\gamma})$, where σ_{SR}^2, σ_{SD}^2, and σ_{RD}^2 are three finite constants depending on the practical design, one can easily verify that $P_2^b \leq \tilde{A}+B \overset{\bar{\gamma}\to\infty}{\approx} (k_2\bar{\gamma})^{-2}$, where k_2 is a constant which depends on σ_{SR}^2, σ_{SD}^2, and σ_{RD}^2.

C. Proof of Proposition 2

Expanding (22), we obtain $I(\epsilon)=C+D$ where

$$C := \int_0^\infty\int_0^\infty\int_{\gamma_m}^\infty\frac{1}{2^\epsilon}\exp(-\gamma_{SR}^\epsilon)\frac{1}{2}\exp\left[-\frac{(\gamma_s-\gamma_m)^2}{\gamma_s+\gamma_m}\right] p(\gamma_s)p(\gamma_{SR}^\epsilon)p(\gamma_{RD}^\epsilon)d\gamma_s d\gamma_{SR}^\epsilon d\gamma_{RD}^\epsilon,$$
$$D := \int_0^\infty\int_0^\infty\int_0^{\gamma_m}\frac{1}{2^\epsilon}\exp(-\gamma_{SR}^\epsilon) p(\gamma_s)p(\gamma_{SR}^\epsilon)p(\gamma_{RD}^\epsilon)d\gamma_s d\gamma_{SR}^\epsilon d\gamma_{RD}^\epsilon.$$

Following steps similar to those in Appendix B, we find $C \leq \tilde{C}$, with $\tilde{C}$ defined in (34) at the top of this page, where $\Gamma[a,x] := \int_x^\infty t^{a-1}\exp(-t)dt$ is the upper incomplete gamma function, and $\Gamma[a] := \Gamma[a,0]$ [15].

Using $\Gamma[n,x] := (n-1)!\exp(-x)\sum_{k=0}^{n-1}\frac{x^k}{k!}$ for any integer n, we obtain (after integrating and manipulating) that $D \leq$

$\tilde{D} := D_1 + D_2$, where

$$D_1 = 2^{M-2\epsilon+2}\left(\frac{1}{3+\bar{\gamma}}\right)^{M+1} \times {}_2F_1\left(1,1+M;2-\epsilon+M;\frac{2}{3+\bar{\gamma}}\right)\frac{\Gamma[M+1]}{\Gamma[M-\epsilon+2]\Gamma[\epsilon]},$$

$$D_2 = \frac{(\bar{\gamma})^{-\epsilon}}{\Gamma[\epsilon]\Gamma[M-\epsilon+2]}\left(\frac{2}{4+\bar{\gamma}}\right)^{M+1}\sum_{i=0}^{\epsilon-1}\left\{\left(1+\frac{4}{\bar{\gamma}}\right)^{-\epsilon} \times {}_2F_1\left(1,1+i+M;2-\epsilon+M;\frac{2}{4+\bar{\gamma}}\right)\frac{\Gamma[1+i+M]}{\Gamma[1+i]}\right\},$$

where ${}_2F_1(a,b;c;z)$ denotes the Gauss hypergeometric function defined in e.g., [1, eq. (15.1.1)]. Given $\epsilon \geq 1$, one can verify that $I(\epsilon) \leq \tilde{C} + \tilde{D} \overset{\bar{\gamma}\to\infty}{\approx} [k(\epsilon)\bar{\gamma}]^{-M-1}$, where $k(\epsilon)$ is a constant which does not depend on $\bar{\gamma}$.

REFERENCES

[1] M. Abramowitz and I. A. Stegun, *Handbook of mathematical functions with formulas, graphs, and mathematical tables*, 9th ed. New York: Dover, 1970.

[2] J. Adeane, M. R. D. Rodrigues, and I. J. Wassell, "Characterisation of the performance of cooperative networks in Ricean fading channels," *12th International Conference on Telecommunications (ICT 2005)*, Cape Town, South Africa, May 3-6, 2005.

[3] J. Boyer, D. D. Falconer, and H. Yanikomeroglu, "Multihop diversity in wireless relaying channels," *IEEE Trans. on Commun.*, vol. 52, no. 10, pp. 1820-1830, Oct. 2004.

[4] D. Chen and J. N. Laneman, "Modulation and demodulation for cooperative diversity in wireless systems," *IEEE Trans. on Wireless Commun.*, vol. 5, no. 7, pp. 1785-1794, Jul. 2006.

[5] M. O. Hasna and M.-S. Alouini, "End-to-end performance of transmission systems with relays over Rayleigh-fading channels," *IEEE Trans. on Wireless Commun.*, vol. 2, no. 6, pp. 1126-1131, Nov. 2003.

[6] M. Janani, A. Hedayat, T. E. Hunter, and A. Nosratinia, "Coded cooperation in wireless communications: Space-time transmission and iterative decoding," *IEEE Trans. on Signal Processing*, vol 52, no. 2, pp. 362 - 371, Feb. 2004.

[7] T. Kang and V. Rodoplu, "Algorithms for the spatial single relay channel," *Proc. of Intl. Conf. on Communications*, vol. 4, pp. 2667 - 2673, Seoul, Korea, May 16-20, 2005.

[8] J. N. Laneman, D. N. C. Tse, and G. W. Wornell, "Cooperative diversity in wireless networks: Efficient protocols and outage behavior," *IEEE Trans. on Inform. Theory*, vol. 50, no. 12, pp. 3062 - 3080, Dec. 2004.

[9] J. N. Laneman and G. W. Wornell, "Distributed space-time-coded protocols for exploiting cooperative diversity in wireless networks," *IEEE Trans. on Inform. Theory*, vol. 49, no. 10, pp. 2415-2425, Oct. 2003.

[10] W. Mo and Z. Wang, "Average symbol error probability and outage probability analysis for general cooperative diversity system at high signal to noise ratio," In *Proc. of Conf. on Info. Sciences and Systems*, Princeton, NJ, Mar. 17-19, 2004.

[11] R. Pabst, B. H. Walke, D. C. Schultz, P. Herhold, H. Yanikomeroglu, S. Mukherjee, H. Viswanathan, M. Lott, W.Zirwas, M. Dohler, H. Aghvami, D. D. Falconer, and G. P. Fettweis, "Relay-based deployment concepts for wireless and mobile broadband cellular radio," *IEEE Communications Magazine*, vol. 42, no. 9, pp. 80-89, Sep. 2004.

[12] A. Ribeiro, X. Cai, and G. B. Giannakis, "Symbol error probabilities for general cooperative links," *IEEE Trans. on Wireless Commun.*, vol. 4, no. 3, pp. 1264 - 1273, May 2005.

[13] A. Sendonaris, E. Erkip, and B. Aazhang, "User cooperation diversity-Part I: System description," *IEEE Trans. on Commun.*, vol. 51, no. 11, pp. 1927-1938, Nov. 2003.

[14] A. Sendonaris, E. Erkip, and B. Aazhang, "User cooperation diversity-Part II: Implementation aspects and performance analysis," *IEEE Trans. on Commun.*, vol. 51, no. 11, pp. 1939-1948, Nov. 2003.

[15] E. W. Weisstein. "Incomplete Gamma Function." From MathWorld – A Wolfram Web Resource. http://mathworld.wolfram.com/IncompleteGammaFunction.html

[16] E. W. Weisstein. "Lambert W-Function." From MathWorld – A Wolfram Web Resource. http://mathworld.wolfram.com/LambertW-Function.html

[17] E. Zimmermann, P. Herhold, and G. Fettweis, "On the performance of cooperative diversity protocols in practical wireless systems," *Proc. of 58th Vehicular Technology Conference*, vol. 4, pp. 2212 - 2216, Orlando, Florida, Oct. 6-9, 2003.

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