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Soft-Decision for Differential Amplify-and-Forward over Time-Varying Relaying Channel

Fengyue Gao[1,2], Lei Kong[3] and Feihong Dong[1]
[1] College of Communication Engineering, PLA University of Science and Technology
Nanjing, China 210007
[2] Institute of China Electronic System Engineering Corporation
Beijing, China 100141
[3] Xi'an Communications Institute
Xi'an, China 710100
[e-mail: fengyue80@gmail.com, dfh_sinlab@hotmail.com, kongleihappy@qq.com]
*Corresponding author: Fengyue Gao

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Abstract

Differential detection schemes do not require any channel estimation, which can be employed under user mobility with low computational complexity. In this work, a soft-input soft-output (SISO) differential detection algorithm is proposed for amplify-and-forward (AF) over time-varying relaying channels based cooperative communications system. Furthermore, maximum-likelihood (ML) detector for *M*-ary differential Phase-shift keying (DPSK) is derived to calculate a posteriori probabilities (APP) of information bits. In addition, when the SISO is exploited in conjunction with channel decoding, iterative detection and decoding approach by exchanging extrinsic information with outer code is obtained. Finally, simulation results show that the proposed non-coherent approach improves detection performance significantly. In particular, the system can obtain greater performance gain under fast-fading channels.

Keywords: Amplify-and-forward, differential phase-shift keying, soft-input soft-output, iterative detection, time-varying relaying channels

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1. Introduction

As a potential solution for coverage problem of cellular networks, cooperative communications is an attractive technology. In cooperative communications, a user in the network acts as a relay to receive signals from a source during its transmission phase, processes and rebroadcasts to a destination in another phase. In this way, additional links, other than the direct link from a source to a destination, can be constructed via relays and hence the overall spatial diversity and performance of the system would benefit from a virtual antenna array that is constructed cooperatively by multiple users. There has been some work investigating cooperative relay networks. [1] has proposed a relay selection method with a low implementation complexity to achieve full diversity in a multi-relay network. In order to develop a unified framework for performance analysis and optimal algorithm design, where efficiency and fairness are both taken into consideration, [2] has focused on the joint channel allocation and relay selection with different cooperation schemes for multi-flow cooperative networks. Due to the broadcast nature of propagation, the security in cooperative relay networks has attracted considerable interest in recent years [3, 4]. Depending on the signal processing strategy that a relay utilizes, relay networks are generally classified as decode-and-forward (DF)[5, 6] and amplify-and-forward (AF)[7-10]. In this paper, we focus our study on AF relaying communication systems due to its lower complexity and superior performance.

Coherent detection of transmitted symbols can be achieved by providing the instantaneous channel state information (CSI) of all transmission links at the destination such as [1, 2]. Although this requirement can be accomplished by sending pilot signals and using channel estimation techniques, all channel estimation techniques are subject to impairments that would directly translate to performance degradation and the computational complexity and overhead would increase with the number of relays. Differential AF (D-AF) is very attractive as it requires less computational burden at the relays and destination. In D-AF, data symbols are differentially encoded at the source. The relay's function is simply to multiply the received signal with a fixed amplification factor. At the destination, the received signals from multi-links are combined to achieve the diversity, and used for non-coherent detection of the transmitted signals without the need of instantaneous CSI.

Differential phase-shift-keying (DPSK) modulation has been employed in cooperative communication system [7, 8] to reduce the computational complexity. At the destination node, the diversity combining differential receiver takes the previous received signals as the references and computes the weighted sum of the phase differences between two adjacent received samples over different propagation to detect the modulated symbols. For mobile fading channels, if the normalized Doppler spread of the channel is small enough, the channel is called slow fading. For slow fading channels, [9] has proposed differential network coding schemes at the physical layer for both AF protocol and DF protocol. In [10], an exact single-integral-form maximum-likelihood (ML) detector and its closed-form approximate expression for AF networks have been derived. Selection combining (SC) for D-AF relay networks has been recently investigated and analyzed in [11]. This combiner can be seen as a counterpart of selection combining of DPSK in point-to-point communications with receive diversity studied in [12]-[15]. The soft-output probability formulas based high-order Differential Amplitude and Phase-Shift Keying (DAPSK) have been derived to approach the achievable channel capacity by iterative detection in [16]. On the other hand, if the normalized

Doppler spread of the channel is large due to high mobility and/or low data rate, the channel is called fast or time-selective fading [17]. The effect of time-selectivity on the performance of conventional DPSK modulation in generalized diversity Rayleigh was analyzed in [18] and [19]. The optimum and suboptimum diversity combining rule with differential BPSK (DBPSK) modulation have been derived in [20]. [21] considers the performance of D-AF relaying for multinode wireless communications and proposes a new set of combining weights for signal detection. The performance of SC for D-AF relay networks has been derived and analyzed in [22]. A practical suboptimal multiple-symbol detection (MSD) for differential dual-hop (D-DH) AF relaying is designed and theoretically analyzed in [23].

A soft-input soft-output (SISO) decoder module has been presented in [24], which implements a soft-output algorithm performing an update of the a posteriori probabilities (APP) of both information and coded symbols based on the code constraint. For many application scenarios, soft decision has been perceived as a promising approach to improve the system performance [25]. An soft information detection has been proposed in Rayleigh fading channel for differential space-time block codes (DSTBC) in [26] and non-coherent frequency-shift keying (NFSK) in [27].

In this paper, an effective SISO differential detection algorithm is proposed for AF over time-varying relaying channels based cooperative communications system. The proposed algorithm aims to evaluate a posteriori probabilities of the information bits. In addition, when the SISO is exploited in conjunction with channel decoding, iterative detection and decoding approach by exchanging extrinsic information with outer code is obtained. Simulation results show that the proposed non-coherent approach improves detection performance significantly, especially for fast fading channels. To be more specific, the main contributions of this paper are summarized as follows:

1. An SISO detection is proposed to calculate APP of information bits. In contrast to hard decision in time-varying relaying channels [22], the proposed algorithm provides more reliable detection due to using soft information. It is noticed that, the proposed algorithm is espectially attractive for the non-coherent relaying systems, since it is capable of exploiting accurate detection information without requiring channel estimation.
2. We consider the use of an error control coding for additional data protection, which little research has been done on with respect to D-AF. When the SISO is exploited in conjunction with channel decoding, iterative detection and decoding approach by exchanging extrinsic information with outer code is obtained. Therefore, it exhibits much better BER performance, especially for fast-fading channels. The error floor can be eliminated by channel coding.

The remainder of this paper is organized as follows. Section 2 describes the system model. Section 3 presents an SISO differential decision algorithm. Maximum a posteriori (MAP) detection rules and iterative decoding are also derived in this section. Simulation results and analysis are demonstrated in Section 4. Finally, conclusions and future work are drawn in Section 5.

Notations: $(\cdot)^*$ and $|\cdot|$ denote conjugate and absolute values of a complex number, respectively. $\mathcal{CN}(\mu,\sigma^2)$ stands for a circularly symmetric complex Gaussian distribution with mean μ and variance σ^2. $E[\cdot]$ denotes the statistical expectation. $\Re\{\cdot\}$ denotes the real part of a complex number.

2. System Model

As is shown in **Fig. 1**, we consider the AF relay system with one Source, one Relay and one Destination. There are a direct link and a cascaded link, via Relay from Source to Destination. Each node has a single antenna, and the communication between nodes is half-duplex, i.e., each node is able to only send or receive in any given time. The channel coefficients at time k, from Source to Destination (SD), from Source to Relay (SR) and from Relay to Destination (RD) are shown with $h_0[k]$, $h_1[k]$ and $h_2[k]$, respectively. A Rayleigh flat-fading model is assumed for each channel, i.e., $h_i \sim \mathcal{CN}(0, \sigma_i^2), i = 0, 1, 2$. The channels are spatially uncorrelated and changing continuously in time. Depending on the time variation in a channel due to mobile speed (characterized by the Doppler spread, which is inversely proportional to coherence time), fading can be classified as either fast fading or slow fading. The time correlation between two channel coefficients, n symbols apart, follows the Jakes' model [28]:

$$\varphi_i(n) = E\{h_i[k]h_i^*[k+n]\} = \sigma_i^2 J_0(2\pi f_i n) \tag{1}$$

where $J_0(\cdot)$ is the zeroth-order Bessel function of the first kind and f_i is the maximum normalized Doppler frequency of the i th channel.

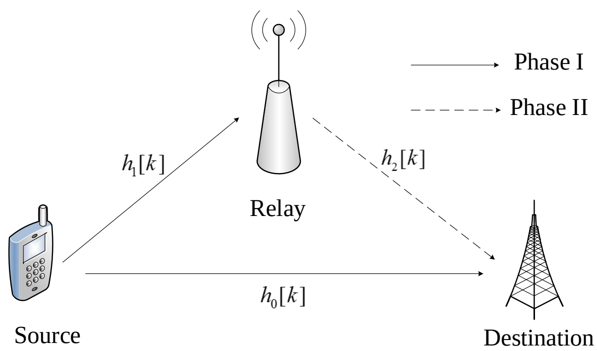

Fig. 1. Wireless relay model under consideration

The information bits $\mathbf{u} = [u_1, \ldots, u_K]$ with length K are passed by a rate r_s forward error correction encoder, e.g., convolutional encoder to get the binary encoding bits $\mathbf{b}' = [b_1', \ldots, b_L']$ with the length $L = K / r_s$. After encoded, the information bits are bitwise interleaved to form the code bits $\mathbf{b} = \mathbf{b}'\prod$. Let us now consider the classic differential phase shift keying (DPSK) scheme. At time k, a group of $q = \log_2 M$ code bits is mapped to an M-PSK symbol as $v[k] \in \mathcal{V}$, where $\mathcal{V} = \{e^{j2\pi m/M}, m = 0, \ldots, M-1\}$. Before transmission, the symbols are encoded differentially as

$$s[k] = v[k]s[k-1], \quad s[0] = 1 \tag{2}$$

The transmission process is divided into two phases. A symbol or a frame of symbols could be transmission in each phase. For both cases, the analysis is the same and only the channel auto-correlation value is different.

In the first phase, symbol $s[k]$ is transmitted by Source to Relay and Destination. Let P_0 be the average power of Source per symbol. The received signals at Destination and Relay are

$$y_0[k] = \sqrt{P_0} h_0[k]s[k] + w_0[k] \tag{3}$$

$$y_1[k] = \sqrt{P_0} h_1[k]s[k] + w_1[k] \tag{4}$$

where $w_0[k], w_1[k] \sim \mathcal{CN}(0, N_0)$ are noise components at Destination and Relay, respectively.

The received signal at Relay is then multiplied by an amplification factor, and re-transmitted to Destination. The amplification factor, based on the variance of SR channel, is commonly used in the literature as

$$A = \sqrt{\frac{P_1}{P_0\sigma_1^2 + N_0}} \tag{5}$$

where P_1 is the average transmitted power per symbol at Relay.

The corresponding received signal at Destination is

$$y_2[k] = Ah_2[k]y_1[k] + w_2[k] \tag{6}$$

where $w_2[k] \sim \mathcal{CN}(0, N_0)$ is the noise component at Destination in the second phase. Substituting (4) to (6) yields

$$y_2[k] = A\sqrt{P_0}h[k]s[k] + w[k] \tag{7}$$

where $h[k] = h_1[k]h_2[k]$ is the equivalent double-Rayleigh with zero mean and variance $\sigma_1^2\sigma_2^2$ [29] and $w[k] = Ah_2[k]w_1[k] + w_2[k]$ is the equivalent noise component. It should be noted that for a given $h_2[k]$, $w[k]$ is a complex Gaussian random variable with zero mean and variance $\sigma_w^2 = N_0(1 + A^2 |h_2[k]|^2)$. Thus $y_2[k]$, conditioned on $s[k]$ and $h_2[k]$, is a complex Gaussian random variable as well. In order to further improve the BER performance of the detection, it is necessary to investigate an efficient algorithm depending on soft decision. In the next section, we will attempt to analyze an SISO decision alogrithm.

3. SISO Decision Algorithm for D-AF

In this section, an SISO decision mathematical model will be derived for the D-AF relaying channels. Firstly, an optimum detection rule will be presented. Since the optimum detection rule involves the instantaneous RD channel gains which are usually hard to acquire, we devise a suboptimum detection rule which only requires the second-order statistics of the RD link CSI. Furthermore, an iterative detection algorithm will be shown with the aids of the decision metric.

3.1 Optimum Detection Rule

In the following, an optimum detection rule will be presented for D-AF relaying channels. At time k, the receiver at Destination detects the transmitted symbol $v[k]$ based on the received signals $y_0[k-1]$, $y_0[k]$, $y_2[k-1]$ and $y_2[k]$ in two consecutive symbol periods. Considering the independence of SD and SR links, the maximum a posteriori (MAP) detection rule minimizing the detection error probability is given by

$$\begin{aligned}
\hat{v}[k] &= \arg\max_{v[k]\in\mathcal{V}} p\{v[k] \mid y_0[k-1], y_0[k], y_2[k-1], y_2[k]\} \\
&= \arg\max_{v[k]\in\mathcal{V}} p\{y_0[k-1], y_0[k], y_2[k-1], y_2[k] \mid v[k]\} \\
&= \arg\max_{v[k]\in\mathcal{V}} p\{y_0[k-1], y_0[k] \mid v[k]\} p\{y_2[k-1], y_2[k] \mid v[k]\} \\
&= \arg\max_{v[k]\in\mathcal{V}} p\{y_0[k] \mid y_0[k-1], v[k]\} p\{y_2[k] \mid y_2[k-1], v[k]\}
\end{aligned} \tag{8}$$

where we assume that the a priori probabilities of $v[k]$ are equal and the statistics of different channels are independent. In addition, the expressions $p\{y_0[k] \mid y_0[k-1], v[k]\}$ and $p\{y_2[k] \mid y_2[k-1], v[k]\}$ are the conditional probability density functions (PDF) of $y_0[k]$ and

$y_2[k]$ conditioned on $y_0[k-1]$, $v[k]$ and $y_2[k-1]$ and $v[k]$, respectively.

First, we derive the explicit form of the conditional PDF expression $p\{y_0[k] \mid y_0[k-1], v[k]\}$. $y_0[k]$, conditioned on $y_0[k-1]$ and $v[k]$, is a complex Gaussian random variable with the mean $\mu_0[k]$ and variance η_0^2. Based on the results given in [30], the mean $\mu_0[k]$ can be expressed by

$$\begin{aligned}\mu_0[k] &= \frac{E\{y_0[k]y_0^*[k-1] \mid v[k]\}}{E\{y_0[k-1]y_0^*[k-1]\}} y_0[k-1] \\ &= \frac{P_0\varphi_0(1)}{P_0\sigma_0^2 + N_0} y_0[k-1]v[k]\end{aligned} \tag{9}$$

In addition, the variance can be expressed by

$$\begin{aligned}\eta_0^2 &= E\{y_0[k]y_0^*[k]\} - \frac{|E\{y_0[k]y_0^*[k-1] \mid v[k]\}|^2}{E\{y_0[k-1]y_0^*[k-1]\}} \\ &= P_0\sigma_0^2 + N_0 - \frac{P_0^2\varphi_0^2(1)}{P_0\sigma_0^2 + N_0}\end{aligned} \tag{10}$$

Therefore, the conditional PDF $p\{y_0[k] \mid y_0[k-1], v[k]\}$ is given by

$$p\{y_0[k] \mid y_0[k-1], v[k]\} = \frac{1}{\pi\eta_0^2} \exp\{-\frac{1}{\eta_0^2} |y_0[k] - \beta_0 y_0[k-1]v[k]|^2\} \tag{11}$$

where

$$\beta_0 = \frac{P_0\varphi_0(1)}{P_0\sigma_0^2 + N_0} \tag{12}$$

Next, we derive the explicit form of the conditional PDF expression $p\{y_2[k] \mid y_2[k-1], v[k]\}$ for the RD link. Conditioned on $y_2[k-1]$, $v[k]$, $h_2[k-1]$ and $h_2[k]$, $y_2[k]$ is a complex Gaussian random variable with mean $\mu_2[k]$ and variance $\eta_2^2[k]$. By using the formulas presented in [30], we get

$$\begin{aligned}\mu_2[k] &= \frac{E\{y_2[k]y_2^*[k-1] \mid v[k], h_2[k-1], h_2[k]\}}{E\{y_2[k-1]y_2^*[k-1] \mid h_2[k-1]\}} y_2[k-1] \\ &= \frac{A^2 P_0\varphi_1(1)h_2^*[k-1]h_2[k]}{A^2 |h_2[k-1]|^2 (P_0\sigma_1^2 + N_0) + N_0} y_2[k-1]v[k]\end{aligned} \tag{13}$$

Substituting the amplification gain formula (5) into (13), we have

$$\mu_2[k] = \frac{P_0 P_1\varphi_1(1)h_2^*[k-1]h_2[k]}{(P_1 |h_2[k-1]|^2 + N_0)(P_0\sigma_1^2 + N_0)} y_2[k-1]v[k] \tag{14}$$

Moreover, the variance $\eta_2^2[k]$ is given by

$$\begin{aligned}\eta_2^2[k] &= E\{y_2[k]y_2^*[k] \mid h_2[k]\} - \frac{|E\{y_2[k]y_2^*[k-1] \mid v[k], h_2[k-1], h_2[k]\}|^2}{E\{y_2[k-1]y_2^*[k-1] \mid h_2[k-1]\}} \\ &= A^2 |h_2[k]|^2 (P_0\sigma_1^2 + N_0) + N_0 - \frac{A^4 P_0^2\varphi_1^2(1) |h_2[k-1]|^2 |h_2[k]|^2}{A^2 |h_2[k-1]|^2 (P_0\sigma_1^2 + N_0) + N_0}\end{aligned} \tag{15}$$

Likewise, the conditional PDF $p\{y_2[k] \mid y_2[k-1], v[k], h_2[k-1], h_2[k]\}$ is

$$p\{y_2[k] \mid y_2[k-1], v[k], h_2[k-1], h_2[k]\} = \frac{1}{\pi\eta_2^2[k]} \exp\{-\frac{1}{\eta_2^2[k]} |y_2[k] - \beta_2[k]y_2[k-1]v[k]|^2\} \tag{16}$$

where

$$\beta_2[k] = \frac{P_0 P_1 \varphi_1(1) h_2^*[k-1] h_2[k]}{(P_1 |h_2[k-1]|^2 + N_0)(P_0 \sigma_1^2 + N_0)} \tag{17}$$

Substituting (11) and (16) into (8) and ignoring those terms independent of $v[k]$, we have

$$\begin{aligned}\hat{v}[k] &= \arg\max_{v[k]\in\mathcal{V}} \{-\frac{1}{\eta_0^2} | y_0[k] - \beta_0 y_0[k-1] v[k] |^2 - \frac{1}{\eta_2^2[k]} | y_2[k] - \beta_2[k] y_2[k-1] v[k] |^2 \} \\ &= \arg\max_{v[k]\in\mathcal{V}} \Re\left\{ \left[\frac{2\beta_0}{\eta_0^2} y_0[k-1] y_0^*[k] + \frac{2\beta_2[k]}{\eta_2^2[k]} y_2[k-1] y_2^*[k] \right] v[k] \right\}\end{aligned} \tag{18}$$

To implement the optimum detection rule (18), the receiver at Destination requires the knowledge of the instantaneous CSI $h_2[k-1]$ and $h_2[k]$, which are hardly available in the context of differential modulation. As an alternative, a suboptimum detection rule is needed.

3.2 Suboptimum Detection Rule

Since the instantaneous CSI $h_2[k-1]$ and $h_2[k]$ are not available at Destination. We can obtain the conditional PDF $p\{y_2[k] | y_2[k-1], v[k]\}$ from the conditional PDF $p\{y_2[k] | y_2[k-1], v[k], h_2[k-1], h_2[k]\}$ by taking expectation with respect to the joint PDF of $h_2[k-1]$ and $h_2[k]$, thus the MAP detection can be expressed as

$$\hat{v}[k] = \arg\max_{v[k]\in\mathcal{V}} \Re\left\{ \frac{2\beta_0}{\eta_0^2} y_0[k-1] y_0^*[k] v[k] + E_{\mathbf{h}_2}\left\{ \frac{2\beta_2[k]}{\eta_2^2[k]} y_2[k-1] y_2^*[k] v[k] \right\} \right\} \tag{19}$$

As can be seen, the MAP metric needs the expectation over the distribution of $\mathbf{h}_2 = (h_2[k-1], h_2[k])$, which does not yield a closed-form expression. As an alternative, it is proposed to use the second-order statistics to replace the instantaneous CSI. The modified detection metric can be expressed as

$$\hat{v}[k] = \arg\max_{v[k]\in\mathcal{V}} \Re\left\{ \left[\frac{2\beta_0}{\eta_0^2} y_0[k-1] y_0^*[k] + \frac{2\beta_2}{\eta_2^2} y_2[k-1] y_2^*[k] \right] v[k] \right\} \tag{20}$$

where

$$\begin{aligned}\beta_2 &= \frac{P_0 P_1 \varphi_1(1) E\{h_2^*[k-1] h_2[k]\}}{(P_1 E\{|h_2[k-1]|^2\} + N_0)(P_0 \sigma_1^2 + N_0)} \\ &= \frac{P_0 P_1 \varphi_1(1) \varphi_2(1)}{(P_1 \sigma_2^2 + N_0)(P_0 \sigma_1^2 + N_0)}\end{aligned} \tag{21}$$

$$\begin{aligned}\eta_2^2 &= A^2 E\{|h_2[k]|^2\}(P_0\sigma_1^2 + N_0) + N_0 - \frac{A^4 P_0^2 \varphi_1^2(1) E\{|h_2[k-1]|^2 |h_2[k]|^2\}}{A^2 E\{|h_2[k-1]|^2\}(P_0\sigma_1^2 + N_0) + N_0} \\ &= A^2\sigma_2^2 (P_0\sigma_1^2 + N_0) + N_0 - \frac{A^4 P_0^2 \varphi_1^2(1) \varphi_2^2(1)}{A^2\sigma_2^2 (P_0\sigma_1^2 + N_0) + N_0} \\ &= \frac{(P_1\sigma_2^2 + N_0)^2 (P_0\sigma_1^2 + N_0)^2 - P_0^2 P_1^2 \varphi_1^2(1) \varphi_2^2(1)}{(P_1\sigma_2^2 + N_0)(P_0\sigma_1^2 + N_0)^2}\end{aligned} \tag{22}$$

Substituting (10), (12), (21) and (22) into (20), we have

$$\hat{v}[k] = \arg\max_{v[k]\in\mathcal{V}} \Re\left\{ \left[\hat{\omega}_1 y_0[k-1] y_0^*[k] + \hat{\omega}_2 y_2[k-1] y_2^*[k] \right] v[k] \right\} \tag{23}$$

where

$$\hat{\omega}_1 = \frac{2 P_0 \varphi_0(1)}{(P_0\sigma_0^2 + N_0)^2 - P_0^2 \varphi_0^2(1)} \tag{24}$$

$$\hat{\omega}_2 = \frac{2(P_0\sigma_1^2 + N_0)P_0P_1\varphi_1(1)\varphi_2(1)}{(P_1\sigma_2^2 + N_0)^2(P_0\sigma_1^2 + N_0)^2 - P_0^2P_1^2\varphi_1^2(1)\varphi_2^2(1)} \tag{25}$$

3.3 Iterative Decoding Algorithm

To improve the BER performance, an iterative detection algorithm is formulated in this subsection. The main idea is that the detector demodulates the symbols, producing soft estimates of each bit as a priori or the extrinsic information of the input symbols, which will be delivered to channel decoder; the decoder computes the APP, and then sends back the extrinsic information to the detector as it's a priori information for the next iteration.

Now, we discuss iterative decoding at Destination in detail. The symbol probabilities $P(v;I)$ are fixed for all demodulation and decoding iterations. On the first and subsequent decoding iteration, the symbol probabilities are transformed to the set of q log-likelihood ratios (LLR) associated with each bit mapped to the symbol. The detector takes as input the symbol probabilities $P(v;I)$ and extrinsic information represented by bit probabilities associated with each coded bit $P(\mathbf{c};I)$ fed back the channel decoder, where $\mathbf{c}$ denotes the q coded bits mapped to v, $P(\mathbf{c};I) = \{P(c_i;I), 0 \le i \le q-1\}$, $c_i \in \{0,1\}$ denotes the i th bit mapped to symbol v. On the first demodulation iteration, no decoding has been performed, and the bit probabilities are assumed equally likely. The detector produces estimated probabilities of values taken by $\mathbf{c}: P(\mathbf{c};O) = \{P(c_i;O), 0 \le i \le q-1\}$. The input representing the LLR of the i th bit mapped to the symbol is related to the input distribution to the detector by

$$r_k = \log\frac{P(c_i = 1;I)}{P(c_i = 0;I)}, 0 \le i \le q-1 \tag{26}$$

The output representing the LLR of the i th bit mapped to the symbol is related to the output distribution to the detector by

$$z_k = \log\frac{P(c_i = 1;O)}{P(c_i = 0;O)}, 0 \le i \le q-1 \tag{27}$$

The detector output distribution is related to the input distribution by

$$P(c_i = l;O) = \sum_{v \in \mathcal{V}_i^l : c_i = l} p(\mathbf{y} \mid v) \prod_{\substack{j=0 \\ j \ne i}}^{q-1} P(c_j;I) \tag{28}$$

where $\mathcal{V}_i^l$ represents the set of transmitted symbols whose i th bit labeling c_i is $l \in \{0,1\}$. Substituting the specific values of the distribution (26) into the expression for output (28),

$$P(c_i = l;O) = \sum_{v \in \mathcal{V}_i^l : c_i = l} p(\mathbf{y} \mid v) \prod_{\substack{j=0 \\ j \ne i}}^{q-1} \frac{e^{c_j r_j}}{1 + e^{r_j}} \tag{29}$$

The output LLR of the detector may be found by combining (27) and (29):

$$z_k = \log\frac{\sum_{v \in \mathcal{V}_i^l : c_i = 1} p(\mathbf{y} \mid v) \prod_{\substack{j=0 \\ j \ne i}}^{q-1} e^{c_j r_j}}{\sum_{v \in \mathcal{V}_i^l : c_i = 0} p(\mathbf{y} \mid v) \prod_{\substack{j=0 \\ j \ne i}}^{q-1} e^{c_j r_j}} \tag{30}$$

where the term $(1 + e^{r_j})$ cancels in the ratio. For the purpose of numeric implementation, it is useful to simplify this expression using the *max-star* operator

$$\max_i *\{x_i\} = \log\left\{\sum_i e^{x_i}\right\} \tag{31}$$

where the binary max-star is $\max*(x,y) = \max(x,y) + \log(1+e^{-|x-y|})$ and multiple arguments imply a recursive relationship. Applying the max-star operator to (30)

$$\begin{aligned} z_k &= \max_{v\in\mathcal{V}_i^l:c_i=1} * \left[\log p(\mathbf{y}\,|\,v) + \sum_{\substack{j=0\\ j\neq i}}^{q-1} c_j r_j\right] - \max_{v\in\mathcal{V}_i^l:c_i=0} * \left[\log p(\mathbf{y}\,|\,v) + \sum_{\substack{j=0\\ j\neq i}}^{q-1} c_j r_j\right] \\ &= \max_{v\in\mathcal{V}_i^l:c_i=1} * \left[\Re\left\{\left[\hat{\omega}_1 y_0[k-1]y_0^*[k] + \hat{\omega}_2 y_2[k-1]y_2^*[k]\right]v[k]\right\} + \sum_{\substack{j=0\\ j\neq i}}^{q-1} c_j r_j\right] \\ &\quad - \max_{v\in\mathcal{V}_i^l:c_i=0} * \left[\Re\left\{\left[\hat{\omega}_1 y_0[k-1]y_0^*[k] + \hat{\omega}_2 y_2[k-1]y_2^*[k]\right]v[k]\right\} + \sum_{\substack{j=0\\ j\neq i}}^{q-1} c_j r_j\right] \end{aligned} \tag{32}$$

A non-iterative detector does not utilize feedback from the decoder, so it is implemented using (32) setting all $r_j = 0$.

4. Simulations

In this section, simulations are conducted to validate the BER performance of the proposed SISO detection algorithm. In our Monte Carlo experiments, the Rayleigh fading channel are simulated based on the modified Jakes' model. The total transmitted power by the Source and Relay nodes is denoted by $P = P_0 + P_1$. The D-AF relaying network under consideration is simulated for various channel qualities using different combinations (Selective Combination [22], semi-Maximum Ratio Combination [22], suboptimum detection). Based on suboptimum detection, we consider the convolutional encoder with generator (7, 5) using bitwise interleaving or not.

We consider both BPSK ($M = 2$) and QPSK ($M = 4$) constellations. Based on the fading power of the channels and the normalized Doppler frequency of the three channels, two cases are considered. In Case I, all nodes are fixed or slowly moving under symmetric channels with $\sigma_0^2 = 1, \sigma_1^2 = 1, \sigma_2^2 = 1, f_0 = 0.001, f_1 = 0.001, f_2 = 0.001$. In Case II, all nodes are fast moving under asymmetric channels with $\sigma_0^2 = 1, \sigma_1^2 = 10, \sigma_2^2 = 1, f_0 = 0.05, f_1 = 0.01, f_2 = 0.05$.

The BER results are plotted versus P/N_0 in Fig. 2-5. Compared Fig. 2 with Fig.3, the error floor increases with the fade rate. For fade rate equal 0.001, the error floor is very small (less than 10^{-6}). For fade rate more than 0.01, the error floor increases to 10^{-3}. Compared Fig. 4 with Fig. 5, we can draw the same conclusion. For high SNR region, BER performance mainly depends on the interference due to mobility, which leads to a large Doppler and little time-correlation.

As can be seen, the SISO algorithm can improve the error performance in both cases. In Case I, there is about 2dB performance gain compared with no channel coding. In Case II, about 5dB performance gain appears and the error floor disappears after using channel coding. Channel coding can resist the interference due to mobility. This observation verifies channel coding can bring about more performance gains in fast fading channels than slow fading channels.

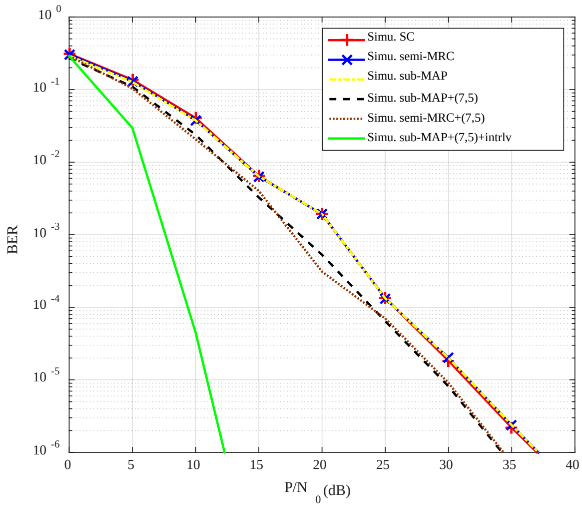

Fig. 2. Simulation BER of a D-AF relaying in Case I using different combinations for DBPSK

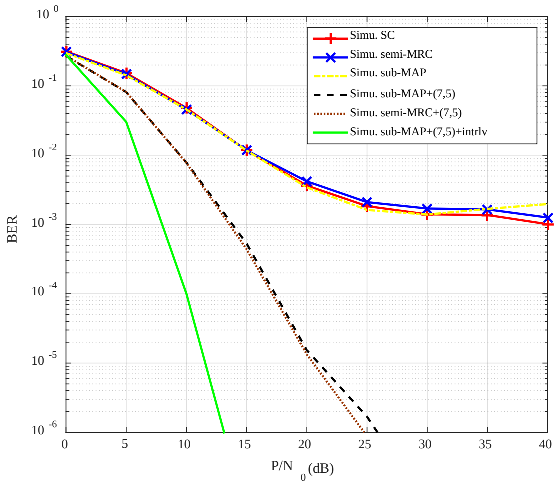

Fig. 3. Simulation BER of a D-AF relaying in Case II using different combinations for DBPSK

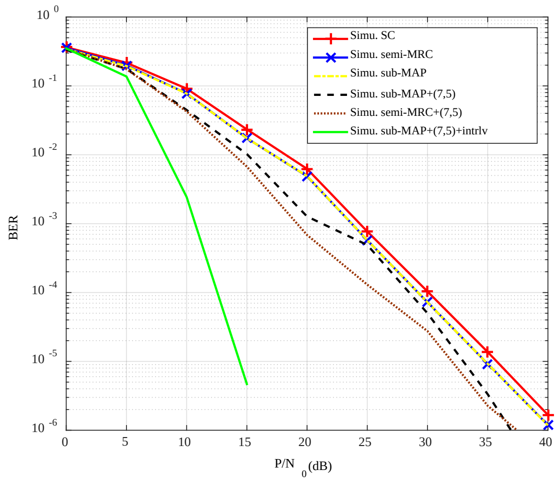

Fig. 4. Simulation BER of a D-AF relaying in Case I using different combinations for DQPSK

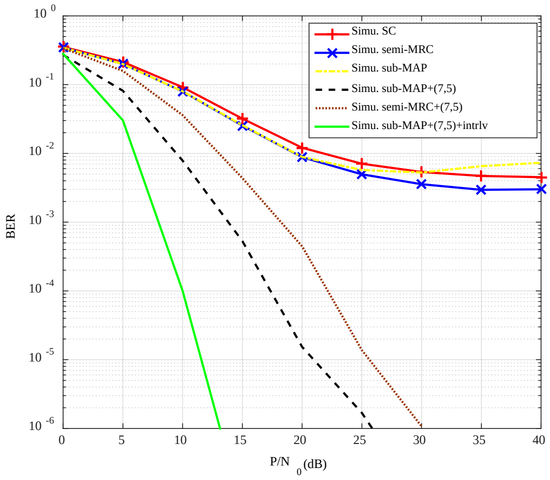

Fig. 5. Simulation BER of a D-AF relaying in Case II using different combinations for DQPSK

At the same time, bitwise interleaving can improve the performance much more effectively. More than 5dB performance gain is due to bitwise interleaving. The fading power of channels and moving speed of nodes can hardly affect the interleaving gain. The interleaving can change the continuous approximate estimation error into stochastic error, which can be corrected by channel coding.

5. Conclusion

In this work, an SISO detection algorithm has been proposed for D-AF over time-varying relaying channels based cooperative communications system. The proposed algorithm aims to evaluate a posteriori probabilities of the information bits. In addition, when the SISO is exploited in conjunction with channel decoding, iterative detection and decoding approach by exchanging extrinsic information with outer code is obtained. Simulation results show that the proposed non-coherent approach improves detection performance significantly.

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Fengyue Gao received the B.S. degree in communication engineering from PLA University of Science and Technology (PLAUST), Nanjing, China in2011, and M.S. degree in communication and information systems from PLAUST in 2014. He is currently working toward the PhD. Degree with the College of Communications Engineering, PLAUST. His research interests include broadband wireless communications, signal processing for communications, advanced channel coding and modulation techniques, network coding and wireless indoor localization.

Lei Kong received the M.S. degree in Electronic Science and Technology from PLA University of Science and Technology, Nanjing, China in 2014. He is currently working in Xi'an Communications Institute of PLA, Xi'an, China. His research interests include high altitude platform communication networks, satellite communication networks and image communication technology.

Feihong Dong received the B.E. degree in information system engineering from PLA University of Science and Technology, Nanjing, China in 2010. He is currently working toward the Ph.D. degree with the College of Communications Engineering, PLA University of Science and Technology. His research interests include high altitude platform communication networks, satellite communication networks and space information networks.