Two-hop AF MIMO Relay Systems with Direct Link – Transceiver Design Based on New Protocol

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***Abstract*—In this paper, we study the transceiver design for two-hop amplify-and-forward (AF) multiple-input multiple-output (MIMO) relay systems using a new protocol, where the source node transmits signals during both time slots. We develop a novel iterative algorithm to optimize the source, relay, and receiver matrices based on the minimum mean-squared error (MMSE) criterion. Simulation results show that compared with conventional AF MIMO relay systems, the proposed system provides a better bit-error-rate performance.**

I. Introduction

Cooperative relay communication has attracted much interest in recent years from both academia and industry due to its potential in increasing the coverage, throughput, and capacity of wireless communication systems [1]-[4]. Relay precoding matrix design for two-hop amplify-and-forward (AF) multiple-input multiple-output (MIMO) relay systems has been investigated in [5]-[10].

It is notable that in the conventional AF relay protocol on half-duplex two-hop MIMO relay systems adopted in [5]-[10], the source node transmits signal only at the first time slot. While making the source node silent at the second time slot simplifies the system design, it is strictly suboptimal [11]. Theoretically, additional diversity gains can be achieved by exploiting the signals transmitted by the source node through the direct link at the second time slot.

In this paper, we consider two-hop AF MIMO relay systems with the direct source-destination link. Different from the conventional AF relay protocol [5]-[10], the source node transmits signals during both time slots. Thus, valuable time diversity provided by the direct link can be utilized to improve the system performance. We investigate the joint source, relay, and receiver matrices design of this system to minimize the mean-squared error (MSE) of the signal waveform estimation at the destination node. Compared with existing works [5]-[10], the transceiver optimization problem in this paper is more challenging as we need to optimize the source precoding matrices at two time slots, rather than only the first time slot.

Since the joint source, relay, and receiver optimization problem is non-convex with matrix variables, the globally optimal solution is intractable to obtain. To solve this problem, we present an iterative algorithm, where the source, relay, and receiver matrices are optimized in an alternating fashion till convergence. In particular, we show that with given source and receiver matrices, the optimal relay precoding matrix has a closed-form solution. While with fixed relay and receiver matrices, the two source precoding matrices can be optimized through solving a quadratically constrained quadratic programming (QCQP) problem. Simulation results show that compared with conventional AF MIMO relay systems, the proposed system provides a better bit-error-rate (BER) performance.

II. System Model

We consider a three-node two-hop MIMO relay communication system as shown in Fig. 1, where the source node (node 1) transmits information to the destination node (node 3) with the aid of a relay node (node 2). We assume that the three nodes are equipped with $N_i, i = 1, 2, 3$ antennas, respectively. With a practical half-duplex relay node, the data transmission from the source node to the destination node is completed in two time slots. At the first time slot, the information-carrying source symbol vector $\mathbf{s} = [s_1, s_2, \cdots, s_{N_b}]^T$ is linearly precoded by a matrix $\mathbf{F}_1 \in \mathbb{C}^{N_1 \times N_b}$ and then transmitted to both the relay node and the destination node, where $(\cdot)^T$ stands for the matrix transpose and N_b denotes the number of independent data streams transmitted simultaneously at the source node.

The signal vectors received at the relay and destination nodes can be written respectively as

$$\mathbf{y}_r = \mathbf{H}_1\mathbf{F}_1\mathbf{s} + \mathbf{n}_r \tag{1}$$

$$\mathbf{y}_{d1} = \mathbf{H}_{31}\mathbf{F}_1\mathbf{s} + \mathbf{n}_{d1} \tag{2}$$

where $\mathbf{H}_1 \in \mathbb{C}^{N_2 \times N_1}$ and $\mathbf{H}_{31} \in \mathbb{C}^{N_3 \times N_1}$ are the MIMO channel matrices of the source-relay link and source-destination link at the first time slot, respectively, $\mathbf{n}_r \in \mathbb{C}^{N_2 \times 1}$ and $\mathbf{n}_{d1} \in \mathbb{C}^{N_3 \times 1}$ are the additive noise vectors at the relay node and the destination node at the first time slot, respectively.

During the second time slot, the relay node linearly precodes $\mathbf{y}_r$ with a matrix $\mathbf{F}_2 \in \mathbb{C}^{N_2 \times N_2}$ and then forwards the precoded signal vector to the destination node. Meanwhile, the source node transmits $\mathbf{F}_3\mathbf{s}$ to the destination node, where $\mathbf{F}_3 \in \mathbb{C}^{N_1 \times N_b}$ is the source precoding matrix at the second time slot. Note that this new AF relay protocol is different to that used in two-hop AF MIMO relay communication systems [5]-[10], where the source node is silent at the second time slot. The signal vector $\mathbf{y}_{d2}$ received at the destination node at the second time slot is given by

$$\mathbf{y}_{d2} = \mathbf{H}_2\mathbf{F}_2\mathbf{y}_r + \mathbf{H}_{32}\mathbf{F}_3\mathbf{s} + \mathbf{n}_{d2} \tag{3}$$

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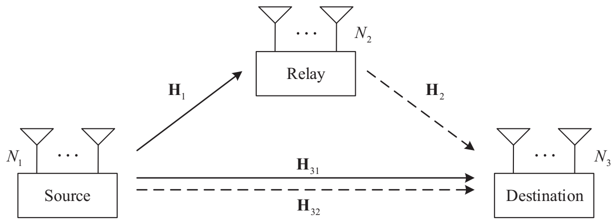

Fig. 1. A two-hop MIMO relay communication system with the direct source-destination link.

where $\mathbf{H}_2 \in \mathbb{C}^{N_3 \times N_2}$ and $\mathbf{H}_{32} \in \mathbb{C}^{N_3 \times N_1}$ are the MIMO channel matrices of the relay-destination link and the source-destination link at the second time slot, respectively, and $\mathbf{n}_{d2} \in \mathbb{C}^{N_3 \times 1}$ is the additive noise vector at the destination node at the second time slot. We assume that $\mathbf{n}_r$, $\mathbf{n}_{d1}$, and $\mathbf{n}_{d2}$ are independent and identically distributed (i.i.d.) Gaussian noise vectors with zero-mean and unit variance entries.

From (1)-(3), the signal vector received at the destination node over two time slots can be written as

$$\begin{aligned}\mathbf{y} &= \begin{bmatrix}\mathbf{y}_{d2}\\ \mathbf{y}_{d1}\end{bmatrix}\\ &= \begin{bmatrix}\mathbf{H}_2\mathbf{F}_2\mathbf{H}_1\mathbf{F}_1 + \mathbf{H}_{32}\mathbf{F}_3\\ \mathbf{H}_{31}\mathbf{F}_1\end{bmatrix}\mathbf{s} + \begin{bmatrix}\mathbf{H}_2\mathbf{F}_2\mathbf{n}_r + \mathbf{n}_{d2}\\ \mathbf{n}_{d1}\end{bmatrix}\\ &= \mathbf{G}\mathbf{s} + \mathbf{v}\end{aligned} \tag{4}$$

where $\mathbf{G} = \begin{bmatrix}\mathbf{H}_2\mathbf{F}_2\mathbf{H}_1\mathbf{F}_1 + \mathbf{H}_{32}\mathbf{F}_3\\ \mathbf{H}_{31}\mathbf{F}_1\end{bmatrix}$ is the equivalent source-destination MIMO channel matrix and $\mathbf{v} = \begin{bmatrix}\mathbf{H}_2\mathbf{F}_2\mathbf{n}_r + \mathbf{n}_{d2}\\ \mathbf{n}_{d1}\end{bmatrix}$ is the equivalent noise vector at the destination node. It is worth noting that (4) represents the most general case for a half-duplex three-node two-hop AF MIMO relay system.

- If $\mathbf{F}_3 = \mathbf{0}$, we have a conventional AF MIMO relay system with the direct link [8]-[10], where the source node is silent at the second time slot.
- AF MIMO relay systems where the direct link is neglected [6]-[7] correspond to (4) with $\mathbf{H}_{31} = \mathbf{H}_{32} = \mathbf{0}$ and $\mathbf{n}_{d1} = \mathbf{0}$.
- For slow-fading MIMO relay channels, there is $\mathbf{H}_{31} = \mathbf{H}_{32}$, while for the fast-fading environment, there is $\mathbf{H}_{31} \neq \mathbf{H}_{32}$.

Due to its simplicity, a linear receiver is used at the destination node to retrieve the source signal vector. Thus, the estimated source signal vector can be written as

$$\hat{\mathbf{s}} = \mathbf{W}^H\mathbf{y} \tag{5}$$

where $\mathbf{W} \in \mathbb{C}^{2N_3 \times N_b}$ is the weight matrix and $(\cdot)^H$ stands for the matrix Hermitian transpose. From (5), the MSE matrix of the signal waveform estimation is given by

$$\begin{aligned}&\mathbf{E}(\mathbf{W}, \mathbf{F}_1, \mathbf{F}_2, \mathbf{F}_3)\\ &= E[(\mathbf{W}^H\mathbf{y} - \mathbf{s})(\mathbf{W}^H\mathbf{y} - \mathbf{s})^H]\\ &= (\mathbf{W}^H\mathbf{G} - \mathbf{I}_{N_b})(\mathbf{W}^H\mathbf{G} - \mathbf{I}_{N_b})^H + \mathbf{W}^H\mathbf{C}_v\mathbf{W}\end{aligned} \tag{6}$$

where $\mathbf{I}_n$ stands for an $n \times n$ identity matrix, $E[\cdot]$ is the statistical expectation with respect to signal and noise, and

$$\mathbf{C}_v = E[\mathbf{v}\mathbf{v}^H] = \begin{bmatrix}\mathbf{H}_2\mathbf{F}_2\mathbf{F}_2^H\mathbf{H}_2^H + \mathbf{I}_{N_3} & \mathbf{0}\\ \mathbf{0} & \mathbf{I}_{N_3}\end{bmatrix}$$

is the noise covariance matrix. Here we assume $E[\mathbf{s}\mathbf{s}^H] = \mathbf{I}_{N_b}$.

From (1), the power of the signal transmitted at the relay node is given by

$$tr(E[\mathbf{y}_r\mathbf{y}_r^H]) = tr(\mathbf{F}_2(\mathbf{H}_1\mathbf{F}_1\mathbf{F}_1^H\mathbf{H}_1^H + \mathbf{I}_{N_2})\mathbf{F}_2^H) \tag{7}$$

where $tr(\cdot)$ stands for the matrix trace. The total power consumption at the source node over two time slots is

$$tr(\mathbf{F}_1\mathbf{F}_1^H + \mathbf{F}_3\mathbf{F}_3^H). \tag{8}$$

The aim of the transceiver design is to minimize the MSE of the signal waveform estimation while satisfying the transmission power constraints at the source and the relay nodes. From (6)-(8), the source, relay, and receiver matrices optimization problem can be written as

$$\min_{\mathbf{W}, \mathbf{F}_1, \mathbf{F}_2, \mathbf{F}_3} \quad tr(\mathbf{E}(\mathbf{W}, \mathbf{F}_1, \mathbf{F}_2, \mathbf{F}_3)) \tag{9}$$

$$\text{s.t. } tr(\mathbf{F}_1\mathbf{F}_1^H + \mathbf{F}_3\mathbf{F}_3^H) \le P_s \tag{10}$$

$$tr(\mathbf{F}_2(\mathbf{H}_1\mathbf{F}_1\mathbf{F}_1^H\mathbf{H}_1^H + \mathbf{I}_{N_2})\mathbf{F}_2^H) \le P_r \tag{11}$$

where P_s and P_r denote the transmission power available at the source node and the relay node, respectively. Note that (10) guarantees that the source node in the new AF MIMO relay system consumes the same amount of power as that of the conventional system where the source node is silent at the second time slot. Compared with existing works [5]-[10], the introduction of the source precoding matrix $\mathbf{F}_3$ at the second time slot makes the problem (9)-(11) much more challenging to solve. Note that the problem (9)-(11) is nonconvex with matrix variables, and the globally optimal solution is intractable to obtain. In the next section, we develop an iterative algorithm to solve the problem (9)-(11).

III. Proposed Transceiver Design Algorithm

In this section, we develop a novel algorithm to optimize $\mathbf{W}$ and $\mathbf{F}_i$, $i = 1, 2, 3$ alternatingly till convergence as follows. As is well known, for any given precoding matrices $\mathbf{F}_i$, $i = 1, 2, 3$, the optimal $\mathbf{W}$ minimizing (9) is the Wiener filter [12] given by

$$\mathbf{W} = \left(\mathbf{G}\mathbf{G}^H + \mathbf{C}_v\right)^{-1}\mathbf{G} \tag{12}$$

where $(\cdot)^{-1}$ denotes matrix inversion.

To optimize the relay precoding matrix, let us first rewrite (9) as a function of $\mathbf{F}_2$ as

$$\begin{aligned}&tr(\mathbf{E}(\mathbf{F}_2)) =\\ &tr((\mathbf{W}_1^H\mathbf{H}_2\mathbf{F}_2\mathbf{H}_1\mathbf{F}_1 + \mathbf{W}_1^H\mathbf{H}_{32}\mathbf{F}_3 + \mathbf{W}_2^H\mathbf{H}_{31}\mathbf{F}_1 - \mathbf{I}_{N_b})\\ &\times(\mathbf{W}_1^H\mathbf{H}_2\mathbf{F}_2\mathbf{H}_1\mathbf{F}_1 + \mathbf{W}_1^H\mathbf{H}_{32}\mathbf{F}_3 + \mathbf{W}_2^H\mathbf{H}_{31}\mathbf{F}_1 - \mathbf{I}_{N_b})^H\\ &+\mathbf{W}_1^H\mathbf{H}_2\mathbf{F}_2\mathbf{F}_2^H\mathbf{H}_2^H\mathbf{W}_1 + \mathbf{W}^H\mathbf{W})\end{aligned} \tag{13}$$

where $\mathbf{W}_1$ and $\mathbf{W}_2$ contain the first and last N_3 rows of $\mathbf{W}$, respectively, i.e., $\mathbf{W} = \left[\mathbf{W}_1^T, \mathbf{W}_2^T\right]^T$. It can be seen from

(13) that with given $\mathbf{W}$, $\mathbf{F}_1$, and $\mathbf{F}_3$, the problem (9)-(11) is converted to the following problem of optimizing $\mathbf{F}_2$

$$\min_{\mathbf{F}_2} tr((\breve{\mathbf{H}}_2\mathbf{F}_2\breve{\mathbf{H}}_1-\mathbf{\Pi})(\breve{\mathbf{H}}_2\mathbf{F}_2\breve{\mathbf{H}}_1-\mathbf{\Pi})^H+\breve{\mathbf{H}}_2\mathbf{F}_2\mathbf{F}_2^H\breve{\mathbf{H}}_2^H) \quad (14)$$

$$\text{s.t. } tr(\mathbf{F}_2(\breve{\mathbf{H}}_1\breve{\mathbf{H}}_1^H+\mathbf{I}_{N_2})\mathbf{F}_2^H)\leq P_r \quad (15)$$

where $\breve{\mathbf{H}}_1=\mathbf{H}_1\mathbf{F}_1, \breve{\mathbf{H}}_2=\mathbf{W}_1^H\mathbf{H}_2, \mathbf{\Pi}=\mathbf{I}_{N_b}-\mathbf{W}_1^H\mathbf{H}_{32}\mathbf{F}_3-\mathbf{W}_2^H\mathbf{H}_{31}\mathbf{F}_1$. The problem (14)-(15) is convex and can be solved by the Lagrange multiplier method.

By introducing a Lagrangian multiplier $\lambda\geq 0$, the optimal $\mathbf{F}_2$ can be expressed as

$$\mathbf{F}_2=(\breve{\mathbf{H}}_2^H\breve{\mathbf{H}}_2+\lambda\mathbf{I}_{N_2})^{-1}\breve{\mathbf{H}}_2^H\mathbf{\Pi}\breve{\mathbf{H}}_1^H(\breve{\mathbf{H}}_1\breve{\mathbf{H}}_1^H+\mathbf{I}_{N_2})^{-1}. \quad (16)$$

It is clear that λ can be determined through the power constraint (15) and the following complementary slackness condition associated with the problem (14)-(15)

$$\lambda(tr(\mathbf{F}_2(\breve{\mathbf{H}}_1\breve{\mathbf{H}}_1^H+\mathbf{I}_{N_2})\mathbf{F}_2^H)-P_r)=0. \quad (17)$$

It can be seen from (17) that if $\mathbf{F}_2$ in (16) with $\lambda=0$ satisfies the constraint (15), then $\mathbf{F}_2=(\breve{\mathbf{H}}_2^H\breve{\mathbf{H}}_2)^{-1}\breve{\mathbf{H}}_2^H\mathbf{\Pi}\breve{\mathbf{H}}_1^H(\breve{\mathbf{H}}_1\breve{\mathbf{H}}_1^H+\mathbf{I}_{N_2})^{-1}$ is the optimal solution to the problem (14)-(15). Otherwise, there must exist $\lambda>0$ such that

$$tr(\mathbf{F}_2(\breve{\mathbf{H}}_1\breve{\mathbf{H}}_1^H+\mathbf{I}_{N_2})\mathbf{F}_2^H)=P_r. \quad (18)$$

Let us introduce $\breve{\mathbf{H}}_2=\mathbf{U}_2\mathbf{\Lambda}_2\mathbf{V}_2^H$ as the singular value decomposition (SVD) of $\breve{\mathbf{H}}_2$. By substituting (16) into (18), we obtain

$$tr(\mathbf{\Lambda}_2(\mathbf{\Lambda}_2^2+\lambda\mathbf{I}_{N_b})^{-1}\mathbf{\Gamma}(\mathbf{\Lambda}_2^2+\lambda\mathbf{I}_{N_b})^{-1}\mathbf{\Lambda}_2)=P_r \quad (19)$$

where $\mathbf{\Gamma}=\mathbf{U}_2^H\mathbf{\Pi}\breve{\mathbf{H}}_1^H(\breve{\mathbf{H}}_1\breve{\mathbf{H}}_1^H+\mathbf{I}_{N_2})^{-1}\breve{\mathbf{H}}_1\mathbf{\Pi}^H\mathbf{U}_2$. By denoting μ_i and γ_i as the ith diagonal elements of $\mathbf{\Lambda}_2$ and $\mathbf{\Gamma}$ respectively, (19) can be rewritten as

$$\sum_{i=1}^{N_b}\frac{\mu_i^2\gamma_i}{(\mu_i^2+\lambda)^2}=P_r. \quad (20)$$

As the left-hand side (LHS) of (20) is a monotonically decreasing function of λ, the bisection method [13] can be applied to solve (20) to obtain λ.

With given $\mathbf{W}$ and $\mathbf{F}_2$, the MSE function (13) can be rewritten as the following function of $\mathbf{F}_1$ and $\mathbf{F}_3$

$$tr(\mathbf{E}(\mathbf{F}_1,\mathbf{F}_3))=tr((\mathbf{D}\mathbf{B}-\mathbf{I}_{N_b})(\mathbf{D}\mathbf{B}-\mathbf{I}_{N_b})^H+\mathbf{W}^H\mathbf{C}_v\mathbf{W}) \quad (21)$$

where $\mathbf{D}=\left[\mathbf{W}_1^H\mathbf{H}_2\mathbf{F}_2\mathbf{H}_1+\mathbf{W}_2^H\mathbf{H}_{31},\mathbf{W}_1^H\mathbf{H}_{32}\right]$, $\mathbf{B}=[\mathbf{F}_1^T,\mathbf{F}_3^T]^T$. Moreover, the LHS of (10) becomes $tr(\mathbf{B}^H\mathbf{B})$, while the LHS of (11) becomes $tr(\mathbf{B}^H\mathbf{A}_0\mathbf{B}+\mathbf{F}_2\mathbf{F}_2^H))$, where $\mathbf{A}_0=\begin{bmatrix}\mathbf{H}_1^H\mathbf{F}_2^H\mathbf{F}_2\mathbf{H}_1 & \mathbf{0}\\ \mathbf{0} & \mathbf{0}\end{bmatrix}$. Thus, the problem of optimizing $\mathbf{F}_1$ and $\mathbf{F}_3$ can be written as

$$\min_{\mathbf{B}} tr((\mathbf{D}\mathbf{B}-\mathbf{I}_{N_b})(\mathbf{D}\mathbf{B}-\mathbf{I}_{N_b})^H) \quad (22)$$

$$\text{s.t. } tr(\mathbf{B}^H\mathbf{B})\leq P_s \quad (23)$$

$$tr(\mathbf{B}^H\mathbf{A}_0\mathbf{B})\leq P_r-tr(\mathbf{F}_2\mathbf{F}_2^H). \quad (24)$$

TABLE I
PROCEDURE OF THE PROPOSED SOURCE, RELAY, AND RECEIVER MATRICES OPTIMIZATION ALGORITHM

1) Initialize the algorithm with $\mathbf{F}_1^{(0)}=\mathbf{F}_3^{(0)}=\sqrt{P_s/2N_b}[\mathbf{I}_{N_b},\mathbf{0}]^T$ and $\mathbf{F}_2^{(0)}=\sqrt{P_r/tr(\mathbf{H}_1\mathbf{F}_1^{(0)}(\mathbf{F}_1^{(0)})^H\mathbf{H}_1^H+\mathbf{I}_{N_2})}\mathbf{I}_{N_2}$, and set $n=0$.
2) Update $\mathbf{W}^{(n)}$ using $\mathbf{F}_1^{(n)}$, $\mathbf{F}_2^{(n)}$ and $\mathbf{F}_3^{(n)}$ as (12).
3) Update $\mathbf{F}_2^{(n+1)}$ using $\mathbf{W}^{(n)}$, $\mathbf{F}_1^{(n)}$, and $\mathbf{F}_3^{(n)}$ as (16).
4) Update $\mathbf{F}_1^{(n+1)}$ and $\mathbf{F}_3^{(n+1)}$ using $\mathbf{W}^{(n)}$ and $\mathbf{F}_2^{(n+1)}$ by solving the problem (22)-(24).
5) If $(\mathrm{mse}_1^{(n)}-\mathrm{mse}_1^{(n+1)})/\mathrm{mse}_1^{(n)}\leq\varepsilon$, iteration ends; otherwise go to step (2).

The problem (22)-(24) is a QCQP problem, which can be efficiently solved by the disciplined convex programming toolbox CVX [14]. We would like to note that the conditional updates of $\mathbf{W}$ and $\mathbf{F}_i$, $i=1,2,3$, may either decrease or maintain but cannot increase the objective function (9). Monotonic convergence of $\mathbf{W}$ and $\mathbf{F}_i$, $i=1,2,3$ towards (at least) a stationary point follows directly from this observation.

The procedure of the proposed iterative algorithm is summarized in Table I, where the superscript (n) denotes variables at the nth iteration, ε is a small positive number for which convergence is acceptable, and mse_1 is the MSE calculated from (9). With the decrease of ε, the system MSE performance improves, whereas the computational complexity increases.

IV. NUMERICAL EXAMPLES

In this section, we study the performance of the proposed source, relay, and receiver matrices optimization algorithm through numerical simulations. In the simulations, a two-hop AF MIMO relay system with 4 antennas at each node (i.e., $N_1=N_2=N_3=4$) is considered. The channel matrices $\mathbf{H}_1$, $\mathbf{H}_2$, $\mathbf{H}_{31}$, and $\mathbf{H}_{32}$ have i.i.d. complex Gaussian entries with zero mean and variances of σ_1^2, σ_2^2, σ_3^2, and σ_3^2, respectively. Note that we assume the same statistics for $\mathbf{H}_{31}$ and $\mathbf{H}_{32}$, as usually it remains unchanged over two consecutive time slots.

Based on the assumption that the noise has unit power, we define $\mathrm{SNR}_1=\sigma_1^2P_s/N_2$, $\mathrm{SNR}_2=\sigma_2^2P_r/N_3$, and $\mathrm{SNR}_3=\sigma_3^2P_s/N_3$ as the signal-to-noise ratio (SNR) for the source-relay, relay-destination, and source-destination links respectively. Following [10], we choose $\mathrm{SNR}_3=\mathrm{SNR}_1-\Delta_{\mathrm{SNR}}$, where Δ_{SNR} stands for the attenuation of the direct link relative to the first-hop channel. We set $\mathrm{SNR}_1=\mathrm{SNR}_2=\mathrm{SNR}$. Quadrature phase-shift keying (QPSK) constellations are used to modulate the source symbols. All simulation results are averaged over 1000 independent channel realizations.

We compare the proposed algorithm with the Tri-step and Bi-step algorithms developed in [10] using the exact CSI. We set $\varepsilon=0.001$ for the proposed scheme and both algorithms in [10]. For the proposed AF MIMO relay system, we consider two cases. In Case 1, $\mathbf{H}_{32}=\mathbf{H}_{31}$, i.e., the exact CSI of the direct link remains unchanged during two transmission slots, corresponding to a slow-fading environment. While in Case 2, we set $\mathbf{H}_{32}\neq\mathbf{H}_{31}$, which simulates a fast-fading environment.

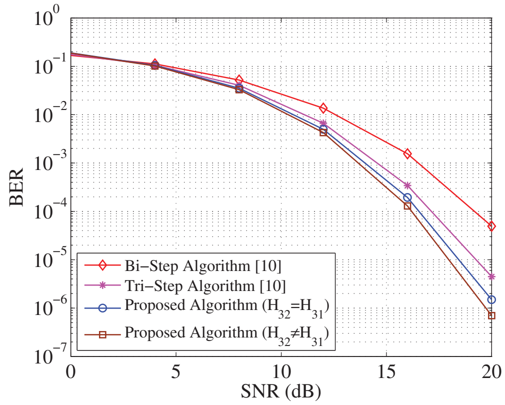

Fig. 2. Example 1. BER versus SNR, $N_b = 4$, $\Delta_{\mathrm{SNR}} = 20$dB.

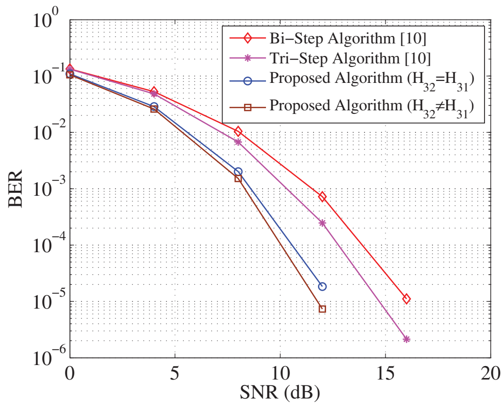

Fig. 3. Example 2. BER versus SNR, $N_b = 4$, $\Delta_{\mathrm{SNR}} = 10$dB.

In the first numerical example, we consider the case where $N_b = 4$ independent data streams are transmitted and set $\Delta_{\mathrm{SNR}} = 20$dB. The procedure in Table I is carried out for the proposed algorithm. Fig. 2 shows the system BER versus SNR of the three algorithms tested. It can be seen from Fig. 2 that the proposed algorithm has a better BER performance than those in [10], which confirms that additional gain can be achieved by making the source node transmit signals at the second time slot. We would like to note that for the algorithms in [10], as the source node is silent at the second time slot, the BER of these two algorithms does not depend on $\mathbf{H}_{32}$.

In the second example, we simulate a MIMO relay system with $N_b = 4$ and $\Delta_{\mathrm{SNR}} = 10$dB. The system BER yielded by three algorithms tested is shown in Fig. 3. Similar to Fig. 2, it can be observed from Fig. 3 that the proposed algorithm yields a lower BER than the algorithms in [10]. Moreover, it can be seen from Figs. 2 and 3 that in a fast-fading channel environment, the BER performance of the proposed AF protocol is further improved due to the valuable time diversity as the proposed algorithm jointly optimizes $\mathbf{F}_1$ and $\mathbf{F}_3$ considering both $\mathbf{H}_{31}$ and $\mathbf{H}_{32}$. It is worth noting that the gap between the BER of the proposed algorithm and that of the algorithms in [10] increases with the SNR, and such performance gap increases as the gain of the direct link is increased from $\Delta_{\mathrm{SNR}} = 20$dB to $\Delta_{\mathrm{SNR}} = 10$dB. This is expected as explained below. When SNR is low (and/or the direct link is weak), the benefit of distributing the transmission power at the source node over two time slots is less noticeable as the effective power for the direct link is limited. As the SNR increases (and/or the direct link becomes stronger), more effective power is available at the source node so that it can allocate the power more flexibly over two time slots to reduce the system BER.

V. Conclusion

We have studied a new AF relay protocol where the source node transmits signals at both time slots in half-duplex MIMO relay systems with the direct link. A novel iterative algorithm has been developed to optimize the source, relay, and receiver matrices in this new AF MIMO relay system. Simulation results show that the proposed algorithm has better BER performance compared with conventional AF MIMO relay systems, and more gains can be achieved with the improvement of the direct-link channel quality.

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