this modification, the scheme in [10] (originally dealing with only the FI imbalance) is applicable for the problem at hand. However, it compensates for the combined effect of I/Q imbalance and multipath channel. As a result, it has to recompute the $N/2$ compensate matrices (see [10, (31)]) for each data packets since channel changes for different packets. Therefore, the scheme in [10] bears high complexity in the long run. In contrast, we only need to run the proposed scheme for the first data packets. It is seen that for the proposed scheme, the curves of $I = 1$ and $I = 5$ almost merge together. This means that $I = 1$ iteration is sufficient. The curves of the scheme in [10] and the proposed scheme are close to the benchmark of the ideal case without TX I/Q imbalance.

V. Conclusion

TX I/Q imbalance is one of the major RF impairments in millimeter-wave SC-FDE systems. To handle this at baseband, this paper proposes a novel estimation and compensation method of TX I/Q imbalance in the presence of unknown multipath channel. By maximizing the likelihood function corresponding to the training sequence, we can iteratively refine the estimation of the TX I/Q imbalance and multipath channel. The obtained TX I/Q imbalance estimate and channel estimate facilitate the compensation of TX I/Q imbalance and FDE. Simulations verify that the proposed method is remarkably advantageous over the counterparts, and can significantly alleviate the adverse effect of TX I/Q imbalance.

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Transceiver Design for Interference MIMO Relay Systems With Direct Links

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***Abstract*—In this paper, we consider interference multiple-input multiple-output (MIMO) relay systems where multiple pairs of transmitters–receivers simultaneously communicate through a single relay node and where all nodes are equipped with multiple antennas. We propose a new iterative algorithm to jointly optimize the relay precoding matrix and the receiver matrices of such a system based on the minimum sum mean-squared error (MSE) criterion. The optimal structure of the relay precoding matrix is developed to decrease the computational complexity of transceiver optimization. Simulation results demonstrate that the proposed transceiver optimization algorithm has a better MSE and bit-error-rate (BER) performance and a faster convergence rate than existing works.**

***Index Terms*—Direct link, interference, minimum mean-squared error (MMSE), multiple-input multiple-output (MIMO) relay.**

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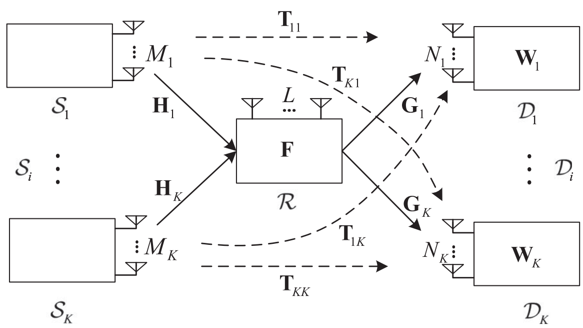

Fig. 1. Interference AF MIMO relay system with direct links.

I. Introduction

Multiple-input multiple-output (MIMO) relay communication systems have received considerable research interest in the past years, because of their potential to improve the link reliability and to extend the network coverage [1]. Compared with other relay strategies, the amplify-and-forward (AF) relay scheme has a lower computational complexity and shorter processing delay as the relay node only amplifies (including a possible linear transformation) and forwards the received signal [2].

Joint source and relay precoding matrices optimization for the two-hop AF MIMO relay systems without the direct source–destination link has been studied in [3]. For the single-user two-hop MIMO relay systems with the direct link, the optimization of the source and relay matrices has been investigated in [4] and [5] based on the maximum mutual information and the minimum mean-squared error (MMSE) criteria, respectively, and in [6] with a zero-forcing decision feedback equalization. Recently, statistically robust source and relay matrices design algorithms have been proposed in [7] for the single-user MIMO relay systems with the direct link when the channel state information (CSI) available is imperfect.

For interference MIMO relay systems, joint source and relay precoding matrices optimization has been studied in [8]. However, the direct links between the transmitters and receivers are not considered in [8]. For the single-user relay systems, it is well known that as the direct link contributes to the spatial diversity, it improves the system performance [4]–[7]. However, for interference relay systems, in addition to providing valuable spatial diversity, direct links also add interference from other transmitters to each receiver. Interestingly, it is shown in [9] that the net effect of direct links is positive in interference relay systems, i.e., they improve the system performance.

In this paper, we study the transceiver design for interference AF MIMO relay systems where multiple pairs of transmitter–receiver simultaneously communicate through a single relay node installed with multiple antennas. Different from [8], we consider the direct links in the design of the relay precoding matrix and the receiver matrices. We propose a new iterative algorithm to jointly optimize the relay precoding matrix and the receiver matrices of such system based on the minimum sum mean-squared error (MSE) criterion. Compared with [9], the optimal structure of the relay precoding matrix is used to decrease the computational complexity of transceiver design. Simulation results demonstrate that the proposed transceiver optimization algorithm has a better MSE and bit-error-rate (BER) performance and a faster convergence rate than the algorithms in [9].

II. System Model

We consider an interference AF MIMO relay system as shown in Fig. 1, where K transmitter–receiver pairs communicate simultaneously with the aid of a relay node. Different to [8], direct links between transmitters and receivers are considered. We assume that M_i and N_i antennas are installed at the ith transmitter and the ith receiver, respectively, and the relay node has L antennas.

Assuming that the relay node operates in the practical half-duplex mode, then the communication process between transmitter–receiver pairs can be completed using two time slots. During the first time slot, the ith transmitter sends an $M_i \times 1$ information-bearing source signal vector $\mathbf{x}_i$ to the relay node, and all receivers. The received signal vector at the relay node can be written as

$$\mathbf{y}_r = \sum_{i=1}^{K} \mathbf{H}_i \mathbf{x}_i + \mathbf{n}_r \tag{1}$$

where $\mathbf{H}_i$ is the $L \times M_i$ MIMO channel matrix from the ith transmitter to the relay node, and $\mathbf{n}_r$ denotes the $L \times 1$ noise vector. The received signal vector at the ith receiver is given by

$$\mathbf{y}_{d_{i1}} = \mathbf{T}_{ii}\mathbf{x}_i + \sum_{j=1, j\neq i}^{K} \mathbf{T}_{ij}\mathbf{x}_j + \mathbf{n}_{i,1}, \qquad i = 1, \ldots, K \tag{2}$$

where $\mathbf{T}_{ij}$ is the $N_i \times M_j$ MIMO channel matrix from the jth transmitter to the ith receiver, and $\mathbf{n}_{i,1}$ denotes the $N_i \times 1$ noise vector at the ith receiver at the first time slot.

During the second time slot, the relay node linearly precodes $\mathbf{y}_r$ as

$$\mathbf{x}_r = \mathbf{F}\mathbf{y}_r \tag{3}$$

and forwards $\mathbf{x}_r$ to all receivers, where $\mathbf{F}$ is the $L \times L$ relay precoding matrix. The signal vector received at the ith receiver can be written as

$$\mathbf{y}_{d_{i2}} = \mathbf{G}_i \mathbf{x}_r + \mathbf{n}_{i,2}, \qquad i = 1, \ldots, K \tag{4}$$

where $\mathbf{G}_i$ is the $N_i \times L$ MIMO channel matrix from the relay node to the ith receiver, and $\mathbf{n}_{i,2}$ denotes the $N_i \times 1$ noise vector at the ith receiver during the second time slot. By combining (1)–(4), we can obtain the received signal vector at the ith receiver during two successive time slots as

$$\begin{aligned}\mathbf{y}_i &= \begin{pmatrix} \mathbf{y}_{d_{i1}} \\ \mathbf{y}_{d_{i2}} \end{pmatrix} \\ &= \begin{pmatrix} \mathbf{T}_{ii} \\ \mathbf{G}_i\mathbf{F}\mathbf{H}_i \end{pmatrix} \mathbf{x}_i + \sum_{j=1, j\neq i}^{K} \begin{pmatrix} \mathbf{T}_{ij} \\ \mathbf{G}_i\mathbf{F}\mathbf{H}_j \end{pmatrix} \mathbf{x}_j \\ &\quad + \begin{pmatrix} \mathbf{n}_{i,1} \\ \mathbf{G}_i\mathbf{F}\mathbf{n}_r + \mathbf{n}_{i,2} \end{pmatrix}, \quad i = 1, \ldots, K.\end{aligned} \tag{5}$$

We assume that $E[\mathbf{x}_i\mathbf{x}_i^H] = \mathbf{I}_{M_i}$, where $E[\cdot]$ denotes the statistical expectation, $(\cdot)^H$ stands for the Hermitian transpose, and $\mathbf{I}_n$ represents the $n \times n$ identity matrix. We also assume that all noises are independent and identically distributed (i.i.d.) additive white Gaussian noise with zero mean and unit variance. For their simplicity, linear receivers are used to retrieve the source signals, and we have $M_i \leq L$ and $N_i \geq M_i$, $i = 1, \ldots, K$. Thus, the estimated source signal vector at the ith receiver can be written as

$$\hat{\mathbf{x}}_i = \mathbf{W}_i^H \mathbf{y}_i, \qquad i = 1, \ldots, K \tag{6}$$

where $\mathbf{W}_i$ is the $2N_i \times M_i$ receiver weight matrix.

From (5) and (6), the MSE of the signal waveform estimation at the ith receiver is given by

$$\begin{aligned}\mathrm{MSE}_i &= \mathrm{tr}(E[(\hat{\mathbf{x}}_i - \mathbf{x}_i)(\hat{\mathbf{x}}_i - \mathbf{x}_i)^H]) \\ &= \mathrm{tr}((\mathbf{W}_i^H\mathbf{B}_{ii} - \mathbf{I}_{M_i})(\mathbf{W}_i^H\mathbf{B}_{ii} - \mathbf{I}_{M_i})^H \\ &\quad + \mathbf{W}_i^H\mathbf{C}_i\mathbf{W}_i), \qquad i = 1, \ldots, K\end{aligned} \tag{7}$$

where $\mathrm{tr}(\cdot)$ denotes the matrix trace, $\mathbf{B}_{ij}$ is the equivalent MIMO channel matrix from the jth transmitter to the ith receiver, and $\mathbf{C}_i$ is the interference-plus-noise covariance matrix at the ith receiver given by

$$\mathbf{B}_{ij} = \begin{pmatrix} \mathbf{T}_{ij} \\ \mathbf{G}_i\mathbf{F}\mathbf{H}_j \end{pmatrix} \tag{8}$$

$$\mathbf{C}_i = \sum_{j=1, j\neq i}^{K} \mathbf{B}_{ij}\mathbf{B}_{ij}^H + \begin{pmatrix} \mathbf{I}_{N_i} & \mathbf{0} \\ \mathbf{0} & \mathbf{G}_i\mathbf{F}\mathbf{F}^H\mathbf{G}_i^H + \mathbf{I}_{N_i} \end{pmatrix}. \tag{9}$$

From (1) and (3), the transmission power of the relay node can be written as

$$E[\mathbf{x}_r\mathbf{x}_r^H] = \mathrm{tr}\left(\mathbf{F}\left(\sum_{i=1}^{K}\mathbf{H}_i\mathbf{H}_i^H + \mathbf{I}_L\right)\mathbf{F}^H\right). \tag{10}$$

Based on (7) and (10), the transceiver optimization problem that minimizes the sum MSE[1] of the signal waveform estimation of all receivers considering the transmission power constraint at the relay node can be formulated as

$$\min_{\{\mathbf{W}_i\},\mathbf{F}} \sum_{i=1}^{K} \mathrm{MSE}_i \tag{11}$$

$$\text{s.t.} \quad \mathrm{tr}\left(\mathbf{F}\left(\sum_{i=1}^{K}\mathbf{H}_i\mathbf{H}_i^H + \mathbf{I}_L\right)\mathbf{F}^H\right) \leq P_r \tag{12}$$

where P_r is the transmission power available at the relay node and $\{\mathbf{W}_i\} = \{\mathbf{W}_i, i = 1,\ldots,K\}$. Similar to [9], we assume that there is a controlling unit (CU), which can be any node in the system. However, to minimize the signaling overhead, particularly in the case of $L > \sum_{i=1}^K M_i$, the best CU is the relay node. The relay node can obtain the knowledge of $\mathbf{H}_i, i = 1,\ldots,K$, through channel training. The CSI of $\mathbf{T}_i = [\mathbf{T}_{i1},\ldots,\mathbf{T}_{iK}]$ and $\mathbf{G}_i$, which is obtained at the ith receiver by channel training, can be sent to the relay node. Therefore, by using all CSI obtained, the relay node can perform the transceiver optimization, and then sends the optimized $\mathbf{W}_i$ to the ith receiver. Although the transmission of CSI consumes system resources and the amount of CSI required increases with K, L, M_i, N_i, $i = 1,\ldots,K$, our algorithm is still valuable, as the algorithms in [9] also require the same amount of global CSI, and our algorithm has better performance and lower computational complexity than the algorithms in [9] as shown later. Compared with [8], the CSI of the direct links $\mathbf{T}_i$, $i = 1,\ldots,K$, is required at the relay node for the proposed transceiver optimization algorithm. Nevertheless, we show later that the direct links are indeed beneficial to the system performance.

Note that by including the direct links, the problems (11) and (12) are much more challenging to solve than that in [8]. We would like to mention that the source precoding matrices optimization is not included in the problems (11) and (12), as we observed that compared with the performance gain obtained from the relay and receiver matrices optimization, optimizing the source matrices brings only small performance improvement (particularly when the number of concurrent data streams at each transmitter is equal to the number of antennas at that transmitter [8]), but significantly increases the computational complexity as additional iterations between the optimization of source matrices and the relay and receiver matrices optimization are required [9].

[1]Similar to [9], in this paper, we focus on interference MIMO relay systems with a symmetric topology. Therefore, no transmitter–receiver pair suffers a fairness problem under the sum MSE criterion. Thus, minimizing the sum MSE improves the performance of signal detection and demodulation at receivers, leading to a better BER performance of every pair.

III. Proposed Transceiver Optimization Algorithm

The problems (11) and (12) have matrix variables and are nonconvex. Thus, the globally optimal solution is difficult to obtain.[2] In this section, we propose a novel iterative algorithm to optimize the relay precoding matrix $\mathbf{F}$ and the receiver matrices $\{\mathbf{W}_i\}$.

With given $\{\mathbf{W}_i\}$, the optimal structure of $\mathbf{F}$ can be obtained by solving the problems (11) and (12) using the Lagrange multiplier method. The Lagrangian function associated with the problems (11) and (12) is given by

$$\mathcal{L} = \sum_{i=1}^{K}\mathrm{MSE}_i + \mu\left[\mathrm{tr}\left(\mathbf{F}\left(\sum_{i=1}^{K}\mathbf{H}_i\mathbf{H}_i^H + \mathbf{I}_L\right)\mathbf{F}^H\right) - P_r\right]$$

where $\mu \geq 0$ is the Lagrangian multiplier. By solving $\frac{\partial\mathcal{L}}{\partial\mathbf{F}} = 0$, we can obtain the optimal structure of $\mathbf{F}$ as

$$\begin{aligned}\mathbf{F} &= (\mathbf{G}^H\mathbf{G} + \mu\mathbf{I}_L)^{-1}\mathbf{G}^H\mathbf{P}\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \mathbf{I}_L)^{-1} \\ &= \mathbf{F}_1\mathbf{P}\mathbf{F}_2 = \mathbf{L}\mathbf{F}_2\end{aligned} \tag{13}$$

where $(\cdot)^{-1}$ is matrix inversion, $\mathbf{F}_1 = \left(\mathbf{G}^H\mathbf{G} + \mu\mathbf{I}_L\right)^{-1}\mathbf{G}^H$, $\mathbf{F}_2 = \mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \mathbf{I}_L)^{-1}$, $\mathbf{L} = \mathbf{F}_1\mathbf{P}$, $\mathbf{H} = [\mathbf{H}_1,\ldots,\mathbf{H}_K]$, and

$$\mathbf{G} = \begin{pmatrix} \mathbf{W}_{1,2}^H\mathbf{G}_1 \\ \vdots \\ \mathbf{W}_{K,2}^H\mathbf{G}_K \end{pmatrix}$$

$$\mathbf{P} = \begin{pmatrix} \mathbf{I}_{M_1} - \mathbf{W}_{1,1}^H\mathbf{T}_{11} & \cdots & -\mathbf{W}_{1,1}^H\mathbf{T}_{1K} \\ \vdots & \ddots & \vdots \\ -\mathbf{W}_{K,1}^H\mathbf{T}_{K1} & \cdots & \mathbf{I}_{M_K} - \mathbf{W}_{K,1}^H\mathbf{T}_{KK} \end{pmatrix}. \tag{14}$$

Here, $\mathbf{W}_{i,1}$ and $\mathbf{W}_{i,2}$ contain the first and the last N_i rows of $\mathbf{W}_i$, respectively, i.e., $\mathbf{W}_i = [\mathbf{W}_{i,1}^T, \mathbf{W}_{i,2}^T]^T$, $i = 1,\ldots,K$, and $(\cdot)^T$ stands for matrix (vector) transpose. Interestingly, $\mathbf{F}_2$ can be viewed as the MMSE receiver for the first-hop multiple access MIMO channel $\mathbf{H}$, while $\mathbf{F}_1$ can be seen as the MMSE transmitter for the equivalent second-hop MIMO channel $\mathbf{G}$. The contribution of the direct links $\mathbf{T}_{ij}$, $i, j = 1,\ldots,K$, is considered by $\mathbf{P}$. When the direct links are ignored [8], there is $\mathbf{P} = \mathbf{I}_M$, where $M = \sum_{i=1}^K M_i$.

With given $\mathbf{F}$, the receiver matrices $\{\mathbf{W}_i\}$ which minimize (11) are the Wiener filter [10] given by

$$\mathbf{W}_i = (\mathbf{B}_{ii}\mathbf{B}_{ii}^H + \mathbf{C}_i)^{-1}\mathbf{B}_{ii}, \qquad i = 1,\ldots,K. \tag{15}$$

Substituting (15) back into (7), the MSE with the optimal $\mathbf{W}_i$ can be written as

$$\mathrm{MSE}_i = \mathrm{tr}(\mathbf{I}_{M_i} - \mathbf{B}_{ii}^H\mathbf{A}_i^{-1}\mathbf{B}_{ii}), \qquad i = 1,\ldots,K \tag{16}$$

where

$$\mathbf{A}_i = \mathbf{C}_i + \mathbf{B}_{ii}\mathbf{B}_{ii}^H, \qquad i = 1,\ldots,K. \tag{17}$$

[2]It is interesting to derive a tight upper bound of (11), which makes the optimization problem easier to solve. This can be a challenging future topic.

Substituting (13) to (8) and (17), we can rewrite

$$\mathbf{B}_{ii} = \begin{pmatrix} \mathbf{I}_{N_i} & \mathbf{0} \\ \mathbf{0} & \mathbf{G}_i\mathbf{L} \end{pmatrix} \begin{pmatrix} \mathbf{T}_{ii} \\ \mathbf{F}_2\mathbf{H}_i \end{pmatrix}$$

$$\begin{aligned} \mathbf{A}_i &= \begin{pmatrix} \mathbf{T}_i \\ \mathbf{G}_i\mathbf{F}\mathbf{H} \end{pmatrix} \begin{pmatrix} \mathbf{T}_i \\ \mathbf{G}_i\mathbf{F}\mathbf{H} \end{pmatrix}^H + \begin{pmatrix} \mathbf{I}_{N_i} & \mathbf{0} \\ \mathbf{0} & \mathbf{G}_i\mathbf{F}\mathbf{F}^H\mathbf{G}_i^H + \mathbf{I}_{N_i} \end{pmatrix} \\ &= \begin{pmatrix} \mathbf{I}_{N_i} & \mathbf{0} \\ \mathbf{0} & \mathbf{G}_i\mathbf{L} \end{pmatrix} \left[\begin{pmatrix} \mathbf{T}_i \\ \mathbf{F}_2\mathbf{H} \end{pmatrix} \begin{pmatrix} \mathbf{T}_i \\ \mathbf{F}_2\mathbf{H} \end{pmatrix}^H + \begin{pmatrix} \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{F}_2\mathbf{F}_2^H \end{pmatrix} \right] \\ &\quad \times \begin{pmatrix} \mathbf{I}_{N_i} & \mathbf{0} \\ \mathbf{0} & \mathbf{L}^H\mathbf{G}_i^H \end{pmatrix} + \mathbf{I}_{2N_i}. \end{aligned}$$

By introducing

$$\mathbf{D}_i = \begin{pmatrix} \mathbf{I}_{N_i} & \mathbf{0} \\ \mathbf{0} & \mathbf{G}_i\mathbf{L} \end{pmatrix}, \quad \mathbf{V}_i = \begin{pmatrix} \mathbf{T}_{ii} \\ \mathbf{F}_2\mathbf{H}_i \end{pmatrix}$$

$$\mathbf{M}_i = \begin{pmatrix} \mathbf{T}_i \\ \mathbf{F}_2\mathbf{H} \end{pmatrix} \begin{pmatrix} \mathbf{T}_i \\ \mathbf{F}_2\mathbf{H} \end{pmatrix}^H + \begin{pmatrix} \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{F}_2\mathbf{F}_2^H \end{pmatrix}$$

the MSE function (16) can be rewritten as

$$\mathrm{MSE}_i = \mathrm{tr}(\mathbf{I}_{M_i} - \mathbf{V}_i^H\mathbf{D}_i^H(\mathbf{D}_i\mathbf{M}_i\mathbf{D}_i^H + \mathbf{I}_{2N_i})^{-1}\mathbf{D}_i\mathbf{V}_i) \quad (18)$$

$$\begin{aligned} &= \mathrm{tr}(\mathbf{I}_{M_i} - \mathbf{V}_i^H\mathbf{M}_i^{-1}\mathbf{V}_i) \\ &\quad + \mathrm{tr}\big(\mathbf{V}_i^H\mathbf{M}_i^{-1}(\mathbf{D}_i^H\mathbf{D}_i + \mathbf{M}_i^{-1})^{-1}\mathbf{M}_i^{-1}\mathbf{V}_i\big) \end{aligned} \quad (19)$$

where the identity

$$\mathbf{B}^H(\mathbf{B}\mathbf{C}\mathbf{B}^H + \mathbf{I})^{-1}\mathbf{B} = \mathbf{C}^{-1} - (\mathbf{C}\mathbf{B}^H\mathbf{B}\mathbf{C} + \mathbf{C})^{-1}$$

is applied to obtain (19) from (18). It can be seen that only the second trace term in (19) depends on $\mathbf{L}$. Thus, based on (13) and (19), $\mathbf{F}$ can be designed through the problem of optimizing $\mathbf{L}$ given by

$$\min_{\mathbf{L}} \quad \sum_{i=1}^{K} \mathrm{tr}\left(\mathbf{V}_i^H\mathbf{M}_i^{-1}(\mathbf{D}_i^H\mathbf{D}_i + \mathbf{M}_i^{-1})^{-1}\mathbf{M}_i^{-1}\mathbf{V}_i\right) \quad (20)$$

$$\text{s.t.} \quad \mathrm{tr}(\mathbf{L}\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \mathbf{I}_L)^{-1}\mathbf{H}\mathbf{L}^H) \le P_r \quad (21)$$

where (21) is obtained by substituting (13) back into (12). As the dimension of $\mathbf{L}$ is $L \times M$, in the case of $L > M$, optimizing $\mathbf{L}$ has a lower computational complexity than directly optimizing $\mathbf{F}$ as in [9]. When $L \le M$, it is better to directly optimize $\mathbf{F}$, as in this case, $\mathbf{L}$ has a larger dimension than $\mathbf{F}$.

The problems (20) and (21) with a matrix variable are nonconvex and it is intractable to obtain the globally optimal solution. In the following, we propose an iterative algorithm to solve the problems (20) and (21).

First, we rewrite (20) as

$$\begin{aligned} &\mathrm{tr}\left(\mathbf{V}_i^H\mathbf{M}_i^{-1}(\mathbf{D}_i^H\mathbf{D}_i + \mathbf{M}_i^{-1})^{-1}\mathbf{M}_i^{-1}\mathbf{V}_i\right) \\ &= \mathrm{tr}\left(\mathbf{V}_i^H\mathbf{M}_i^{-\frac{H}{2}}\left(\mathbf{M}_i^{\frac{H}{2}}\mathbf{D}_i^H\mathbf{D}_i\mathbf{M}_i^{\frac{1}{2}} + \mathbf{I}_{M+N_i}\right)^{-1}\mathbf{M}_i^{-\frac{1}{2}}\mathbf{V}_i\right) \\ &= \mathrm{tr}\Big(\mathbf{V}_i^H\mathbf{M}_i^{-\frac{H}{2}}\left(\mathbf{J}_i\mathbf{J}_i^H + \mathbf{Q}_i\mathbf{L}^H\mathbf{G}_i^H\mathbf{G}_i\mathbf{L}\mathbf{Q}_i^H + \mathbf{I}_{M+N_i}\right)^{-1} \\ &\qquad \times \mathbf{M}_i^{-\frac{1}{2}}\mathbf{V}_i\Big) \\ &= \mathrm{tr}\Big(\mathbf{V}_i^H\mathbf{M}_i^{-\frac{H}{2}}\mathbf{E}_i^{-H}\big(\mathbf{I}_{M+N_i} + \mathbf{E}_i^{-1}\mathbf{Q}_i\mathbf{L}^H\mathbf{G}_i^H\mathbf{G}_i \\ &\qquad \times \mathbf{L}\mathbf{Q}_i^H\mathbf{E}_i^{-H}\big)^{-1}\mathbf{E}_i^{-1}\mathbf{M}_i^{-\frac{1}{2}}\mathbf{V}_i\Big) \\ &= \mathrm{tr}\left(\mathbf{Z}_i(\mathbf{I}_{M+N_i} + \mathbf{\Pi}_i^H\mathbf{\Pi}_i)^{-1}\mathbf{Z}_i^H\right) \end{aligned} \quad (22)$$

where $\mathbf{M}_i = \mathbf{M}_i^{\frac{1}{2}}\mathbf{M}_i^{\frac{H}{2}}$, $\mathbf{J}_i$ and $\mathbf{Q}_i$ contain the first N_i and the last M columns of $\mathbf{M}_i^{\frac{H}{2}}$, respectively (i.e., $\mathbf{M}_i^{\frac{H}{2}} = [\mathbf{J}_i, \mathbf{Q}_i]$), $\mathbf{I}_{M+N_i} + \mathbf{J}_i\mathbf{J}_i^H = \mathbf{E}_i\mathbf{E}_i^H$, $\mathbf{Z}_i = \mathbf{V}_i^H\mathbf{M}_i^{-\frac{H}{2}}\mathbf{E}_i^{-H}$, and $\mathbf{\Pi}_i = \mathbf{G}_i\mathbf{L}\mathbf{Q}_i^H\mathbf{E}_i^{-H}$. From (22), it can be shown that

$$\begin{aligned} &\mathrm{tr}(\mathbf{Z}_i(\mathbf{I}_{M+N_i} + \mathbf{\Pi}_i^H\mathbf{\Pi}_i)^{-1}\mathbf{Z}_i^H)) \\ &= \min_{\mathbf{R}_i} \mathrm{tr}\Big(\mathbf{Z}_i\big[(\mathbf{R}_i^H\mathbf{\Pi}_i - \mathbf{I}_{M+N_i})(\mathbf{R}_i^H\mathbf{\Pi}_i - \mathbf{I}_{M+N_i})^H \\ &\qquad + \mathbf{R}_i^H\mathbf{R}_i\big]\mathbf{Z}_i^H\Big). \end{aligned} \quad (23)$$

In fact, the right-hand side (RHS) of (23) is a quadratic programming problem and the optimal solution is

$$\mathbf{R}_i = (\mathbf{\Pi}_i\mathbf{\Pi}_i^H + \mathbf{I}_{N_i})^{-1}\mathbf{\Pi}_i. \quad (24)$$

By substituting (24) back into the RHS of (23) we obtain (22).

Based on (22) and (24), the problems (20) and (21) can be converted to

$$\begin{aligned} \min_{\{\mathbf{R}_i\},\mathbf{L}} \quad & \sum_{i=1}^{K} \mathrm{tr}\big(\mathbf{Z}_i[(\mathbf{R}_i^H\mathbf{G}_i\mathbf{L}\mathbf{Q}_i^H\mathbf{E}_i^{-H} - \mathbf{I}_{M+N_i}) \\ & \times(\mathbf{R}_i^H\mathbf{G}_i\mathbf{L}\mathbf{Q}_i^H\mathbf{E}_i^{-H} - \mathbf{I}_{M+N_i})^H + \mathbf{R}_i^H\mathbf{R}_i]\mathbf{Z}_i^H\big) \quad (25) \\ \text{s.t.} \quad & \mathrm{tr}(\mathbf{L}\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \mathbf{I}_L)^{-1}\mathbf{H}\mathbf{L}^H) \le P_r \quad (26) \end{aligned}$$

where $\{\mathbf{R}_i\} = \{\mathbf{R}_i, i = 1, \ldots, K\}$. The problems (25) and (26) can be solved by iteratively updating $\{\mathbf{R}_i\}$ and $\mathbf{L}$. In each iteration, $\{\mathbf{R}_i\}$ is first optimized as given by (24) using $\mathbf{L}$ from the previous iteration. Then based on the $\{\mathbf{R}_i\}$ obtained in the current iteration, $\mathbf{L}$ can be optimized through the Lagrange multiplier method as shown below.

Let us introduce $\mathbf{\Phi}_i = \mathbf{R}_i^H\mathbf{G}_i$, $\mathbf{\Theta}_i = \mathbf{Q}_i^H\mathbf{E}_i^{-H}$, and $\mathbf{\Omega} = \mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \mathbf{I}_L)^{-1}\mathbf{H}$. The problem of optimizing $\mathbf{L}$ can be rewritten as

$$\min_{\mathbf{L}} \quad \sum_{i=1}^{K} \mathrm{tr}(\mathbf{Z}_i(\mathbf{\Phi}_i\mathbf{L}\mathbf{\Theta}_i - \mathbf{I}_{M+N_i})(\mathbf{\Phi}_i\mathbf{L}\mathbf{\Theta}_i - \mathbf{I}_{M+N_i})^H\mathbf{Z}_i^H) \quad (27)$$

$$\text{s.t.} \quad \mathrm{tr}(\mathbf{L}\mathbf{\Omega}\mathbf{L}^H) \le P_r. \quad (28)$$

By using the identities of $\mathrm{vec}(\mathbf{A}\mathbf{B}\mathbf{C}) = (\mathbf{C}^T \otimes \mathbf{A})\,\mathrm{vec}(\mathbf{B})$ and $\mathrm{tr}(\mathbf{A}^H\mathbf{B}\mathbf{A}\mathbf{C}) = (\mathrm{vec}(\mathbf{A}))^H(\mathbf{C}^T \otimes \mathbf{B})\mathrm{vec}(\mathbf{A})$ [11], where $\mathrm{vec}(\cdot)$ denotes the operator which stacks all columns of a matrix on top of each other into a vector and $\otimes$ represents the Kronecker product, we have

$$\begin{aligned}&\mathrm{tr}(\mathbf{Z}_i(\mathbf{\Phi}_i\mathbf{L}\mathbf{\Theta}_i-\mathbf{I}_{M+N_i})(\mathbf{\Phi}_i\mathbf{L}\mathbf{\Theta}_i-\mathbf{I}_{M+N_i})^H\mathbf{Z}_i^H)\\&=(\mathbf{U}_i\mathrm{vec}(\mathbf{L})-\mathrm{vec}(\mathbf{Z}_i))^H(\mathbf{U}_i\mathrm{vec}(\mathbf{L})-\mathrm{vec}(\mathbf{Z}_i)) \qquad (29)\end{aligned}$$

$$\mathrm{tr}(\mathbf{L}\mathbf{\Omega}\mathbf{L}^H)=(\mathrm{vec}\,(\mathbf{L}))^H\left(\mathbf{\Omega}^T\otimes\mathbf{I}_L\right)\mathrm{vec}\,(\mathbf{L}) \qquad (30)$$

where $\mathbf{U}_i=\mathbf{\Theta}_i{}^T\otimes(\mathbf{Z}_i\mathbf{\Phi}_i)$. Using (29) and (30), the solution to the problems (27) and (28) is given by

$$\mathrm{vec}(\mathbf{L})=\left(\sum_{i=1}^{K}\mathbf{U}_i^H\mathbf{U}_i+\lambda\,\mathbf{\Omega}^T\otimes\mathbf{I}_L\right)^{-1}\sum_{i=1}^{K}\mathbf{U}_i^H\,\mathrm{vec}\,(\mathbf{Z}_i) \qquad (31)$$

where $\lambda\geq 0$ is the Lagrangian multiplier, which can be calculated by substituting (31) into $\mathrm{tr}(\mathbf{L}\mathbf{\Omega}\mathbf{L}^H)=P_r$ and solving the obtained equation through the bisection search [12]. We would like to mention that the conditional updates of $\{\mathbf{R}_i\}$ and $\mathbf{L}$ may either reduce or maintain the value of the objective function (25) but never increase it. Thus, a monotonic convergence of $\mathbf{L}$ toward (at least) a stationary point of (25) can be shown based on this observation.

After the convergence of the iterative algorithm, the optimal $\mathbf{F}$ can be obtained by substituting the solution $\mathbf{L}$ into (13). Finally, the receiver matrices $\{\mathbf{W}_i\}$ can be calculated by (15).

IV. Numerical Examples

In this section, we study the performance of the proposed transceiver design algorithm through numerical simulations. We simulate an interference MIMO relay system with $K=2$ pairs of transmitter–receiver, where $M_1=M_2=N_1=N_2=2$ and $L=8$. The channel matrices $\mathbf{H}_i$ and $\mathbf{G}_i$ have i.i.d. complex Gaussian entries with zero mean and variance σ_1^2, while the direct link channel matrices $\mathbf{T}_{ij}$ have i.i.d. complex Gaussian entries of zero mean and variance σ_2^2. We define SNR_1, SNR_2, and SNR_3 as the signal-to-noise ratio (SNR) for the source–relay, relay–destination, and source–destination links, respectively. We set $\mathrm{SNR}_1=15\,\mathrm{dB}$, $\mathrm{SNR}_2=\mathrm{SNR}$, and $\mathrm{SNR}_3=10\,\mathrm{dB}$. Quadrature phase-shift keying constellations are employed to modulate the source symbols. All simulation results are averaged through 10^4 independent channel realizations.

We compare the performance of the proposed transceiver design algorithm with both Algorithms 1 and 2 in [9]. For a fair comparison, the source precoding matrices in [9] are set to be scaled discrete Fourier transform matrices. Figs. 2 and 3 show, respectively, the MSE and BER comparisons of the three algorithms and the proposed algorithm without direct links at the second iteration and convergence. It can be clearly seen that the direct links are indeed beneficial to the system performance. We also observe from Figs. 2 and 3 that by exploiting the structure of the optimal relay precoding matrix, the proposed algorithm shows a better performance in MSE and BER than both algorithms in [9] at the high SNR region. Moreover, the BER gap between the proposed algorithm and the approaches in [9] increases with the SNR, suggesting that the diversity gain of the proposed algorithm is larger than that of the approaches in [9].

Fig. 4 shows the MSE comparison of the three algorithms versus the number of iterations when SNR = 20 dB. It can be seen that the proposed algorithm has a faster convergence rate than both algorithms in [9].

Finally, we discuss the computational complexity of the three algorithms tested. For the simplicity of notation, we assume $M_i=N_i=N$, $i=1,\ldots,K$, and thus $M=KN$. In each iteration of the proposed algorithm, matrix inversions in (24) and (31) need to be computed, which have a complexity order of $\mathcal{O}(KN^3)$ and $\mathcal{O}(M^3L^3)$, respectively. Therefore, the per iteration computational complexity of the proposed

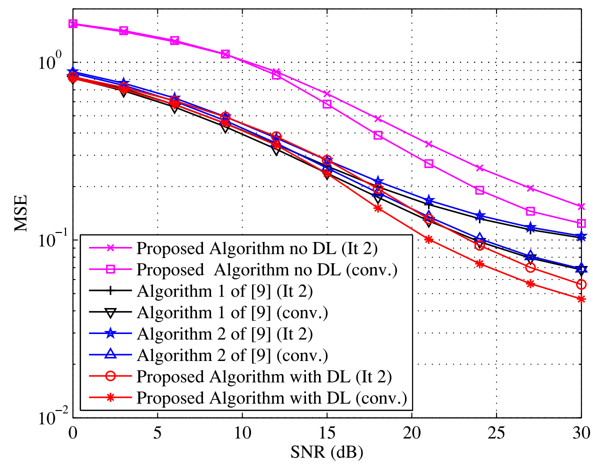

Fig. 2. MSE comparison of the proposed algorithm and the algorithms in [9].

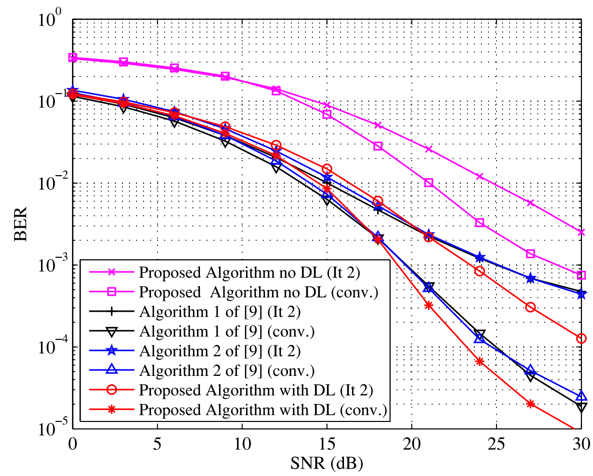

Fig. 3. BER comparison of the proposed algorithm and the algorithms in [9].

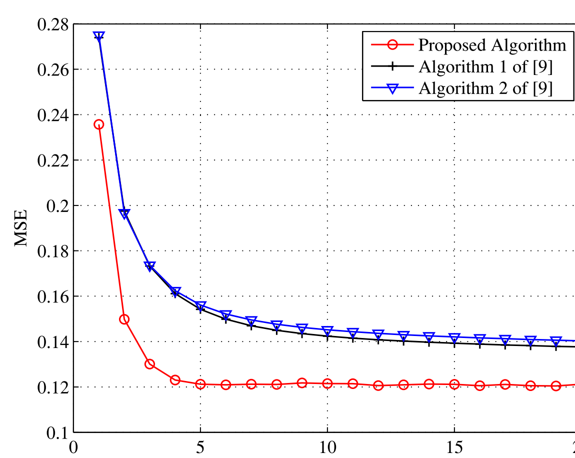

Fig. 4. MSE comparison of the proposed algorithm and the algorithms in [9] at different number of iterations.

TABLE I
AVERAGE NUMBER OF ITERATIONS REQUIRED UNTIL CONVERGENCE BY THREE ALGORITHMS

SNR (dB)	0	5	10	15	20	25	30
Proposed algorithm	3	3	4	4	4	4	4
Algorithm 1 of [9]	8	9	12	12	12	12	12
Algorithm 2 of [9]	7	9	12	12	11	12	12

algorithm is $\mathcal{O}(KN^3 + M^3L^3)$. According to [9], both Algorithms 1 and 2 have a per iteration complexity order of $\mathcal{O}(KN^3 + L^6)$. Therefore, when $L \geq M$, the proposed algorithm has a lower per iteration complexity than both algorithms in [9].

The overall complexity of the three algorithms depends also on the number of iterations required until convergence. Table I shows the average number of iterations needed by three algorithms at various SNRs to reach a convergence criterion that the MSE improvement is less than 10^{-3} between two iterations. It can be clearly seen from Table I that the proposed algorithm requires less iterations. Therefore, the proposed algorithm has a lower computational complexity than both algorithms in [9].

V. CONCLUSION

We have proposed a novel iterative minimum sum MSE based transceiver design algorithm for interference MIMO relay systems with direct links. Simulation results demonstrate that by properly exploiting the optimal structure of the relay precoding matrix, the system MSE and BER performances are improved, and the computational complexity of the transceiver optimization is reduced. Extending the proposed algorithm to interference MIMO relay systems with multiple relay nodes is an interesting and challenging future topic.

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How Much Bandpass Filtering is Required in Massive MIMO Base Stations?

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***Abstract*—In this paper, we study the impact of aliased out-of-band (OOB) interference signals on the information sum rate of the maximum ratio combining receiver in massive multiple-input multiple-output (MIMO) uplink, with both perfect and imperfect channel estimates, in order to determine the required out-of-band attenuation in radio-frequency (RF) bandpass filters (BPFs). With imperfect channel estimates, our study reveals that as the number of base-station (BS) antennas (M) increases, the required attenuation at the BPFs increases as $\mathcal{O}(\sqrt{M})$ with $M \to \infty$, provided the desired information sum rate (both in the presence and in the absence of aliased OOB interferers) remains fixed. This implies a practical limit on the number of BS antennas due to the increase in BPF design complexity and power consumption with increasing M.**

***Index Terms*—Aliasing, attenuation, bandpass filter (BPF), information sum-rate, massive multiple-input multiple-output (MIMO), out-of-band (OOB) interference.**

I. INTRODUCTION

In the development of the next-generation [fifth-generation (5G)] wireless communication systems, massive multiple-input multiple-output (MIMO) has been visualized as a key technology, which would supplement other 5G technologies forming an integrated communication network, which has high energy and spectral efficiency and low latency [1]. The vision for massive MIMO is to equip the base station (BS) with a large antenna array (of the order of hundreds) to support a few tens of users in the same time-frequency resource [2]. For massive MIMO systems, it has been suggested that low-complexity signal processing can achieve good information rate performance due to the averaging of noise and hardware imperfections across the M BS antennas (as $M \to \infty$) [3], [4].

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