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Iterative Transceiver Beamformer Design for Multi-Pair Two-Way Distributed Relay Networks

Jingxiao Ma, Wei Liu, and Jie Zhang
Communications Research Group
Department of Electronic and Electrical Engineering
University of Sheffield, UK
{jma15, w.liu, jie.zhang}@sheffield.ac.uk

***Abstract*—In this paper the transceiver beamforming design problem for multipair two-way distributed relay networks is studied, where each multi-antenna user in one user group communicate with its partner in the other user group via distributed single-antenna relay nodes. To achieve a satisfactory performance while relieving relay nodes of the usual computation task, two iteration-based transceiver beamforming schemes are proposed to coordinate the operation of the users from the two user groups, where the beamforming vectors are determined at the user nodes through an iterative process. Simulation results indicate that both schemes can achieve considerable SINR improvement after only a few iterations compared to the existing zero-forcing scheme.**

I. Introduction

Distributed relay networks can exploit the spatial diversity of network nodes, reduce the deployment cost and extend system coverage [1–4]. Various devices in a certain area can be utilized as relay nodes, including mobile devices whose antenna number is strictly limited by the physical size and transmitting power.

One specific area of research is *multipair relay networks*, where multiple users can communicate with their destination pair simultaneously with the assistance of one or multiple relay nodes [5, 6]. In [5, 6], the basic multi-pair one-way relay networks were studied with the assumption of full relay nodes cooperation. To further improve the throughput and achievable rate of traditional four-phase relay networks, two-way relay systems were proposed based on the concept of analog network coding [7]. In [8–10], the multipair two-way relay networks with one multi-antenna relay node centrally processing the received signals were studied. Distributed single-antenna relay networks with multipair two-way scenarios were considered in [11–14]. A distributed relay beamforming design aiming at minimizing the total transmission power at the relay nodes subject to a signal-to-interference-plus-noise ratio (SINR) constraint at each user node was proposed in [11] , which was formulated as a convex semidefinite programming problem. In [12], a similar structure was studied, but with an individual transmission power constraint. With the assumption that the channel state information (CSI) for all relay-user connections are known at every relay, [13] proposed a distributed beamforming scheme where relay weight coefficients are decided at each relay. By contrast, a scheme not requiring cooperation between the relay nodes was investigated in detail in [14]; however, the relay number should be large enough to achieve a good performance.

For a multipair two-way relay network, the main bottleneck is the inter-pair interference (IPI) caused by simultaneous signal transmission of multiple user pairs. In [10, 13, 15], beamforming methods base on zero forcing (ZF) were proposed for IPI cancellation. On the other hand, [16] studied the scheme of block-diagonalization (BD), which is employed at one central relay node with multiple antennas. In [17], a coordinated eigen-beamforming scheme was proposed where multi-antenna user node and multi-antenna relay node are assumed, and the beamforming weights at user nodes and relay node are jointly determined to maximize the effective channel gain between user pairs. A similar scheme was studied in [18], where the user pairs are also equipped with multiple antennas, and the signal space alignment (SSA) method is used for transceiver beamforming to reduce the effective number of interference, with an enhanced ZF method for relay beamforming.

In the aforementioned literature, the main signal processing procedures and beamforming weights determination processes are performed at the relay nodes, and this will take significant resources from the relay nodes, such as time, computational capacity and processing power. If the resources requirement for the relay is reduced, more devices can potentially be utilized as distributed relay nodes, and help forward signals for user pairs with their spare resources.

Motivated by this, in this paper, we focus on a multipair two-way distributed relay network with multi-antenna users from one user group simultaneously transmitting signals to their user partners in the other user group via distributed relay nodes working in the simple amplify-and-forward (AF) mode, and two iteration-based transceiver beamforming schemes are proposed for coordination of the user pairs, where the beamforming vectors are decided at the user side, instead of the relay nodes.

This paper is organized as follows. In Section II, the overall system model is introduced. The proposed iterative zero-forcing scheme and iterative SINR optimization scheme are presented in Section III. Simulation results and relevant discussions are provided in Section IV and conclusions are drawn in Section V.

II. System Model

We consider a time-slotted dual-hop multipair two-way distributed relay network consisting of K multi-antenna communication pairs (each is equipped with N antennas) which are divided into two groups (X_a, X_b) as shown in Fig. 1. We assume that the distance between the two groups are long enough compared to their transmission power that the direct link does not exist, and thus, the transmission between user pairs is assisted by M single-antenna distributed relay nodes between them.

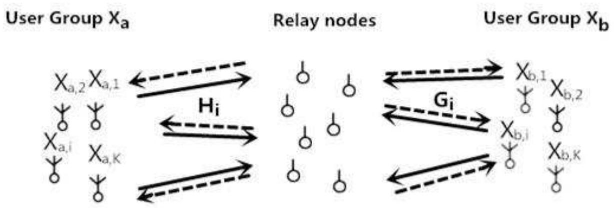

Fig. 1. Model for the distributed relay network.

Two transmission phases are considered. In the uplink phase, the users transmit information stream to the relay nodes simultaneously with transmit beamforming. Then in the downlink phase, the distributed relay nodes use low-complexity AF protocols to broadcast the signals back to the user nodes. The transmission channels are assumed to be Rayleigh fading, reciprocal and quasi-stationary, so that the channel gains remain unchanged during the two time slot phases.

In the first time slot, the transmitted signal from user $X_{a,i}$ and $X_{b,i}$ ($i = 1, ..., K$) to the relay nodes are

$$\mathbf{x}_{a,i} = \mathbf{a}_i x_{a,i}, \quad \mathbf{x}_{b,i} = \mathbf{b}_i x_{b,i}, \quad i \in \{1, \cdots, K\}, \tag{1}$$

where $x_{a,i}$ and $x_{b,i}$ are the data symbol. $\mathbf{a}_i, \mathbf{b}_i \in \mathbb{C}^{N\times 1}$ are the transmit beamforming vectors, which satisfy the total transmit power constraint $||\mathbf{a}_i||^2 \leq P_s$ and $||\mathbf{b}_i||^2 \leq P_s$, with P_s being the upper bound. Then the signals received at the relay can be represented by an $M \times 1$ vector $\mathbf{r}$, given by

$$\mathbf{r} = \sum_{i=1}^{K} \mathbf{H}_i \mathbf{a}_i x_{a,i} + \sum_{i=1}^{K} \mathbf{G}_i \mathbf{b}_i x_{b,i} + \mathbf{n}_R, \tag{2}$$

where $\mathbf{H}_i, \mathbf{G}_i \in \mathbb{C}^{M\times N}$ are the channel matrix from user $X_{a,i}$ and $X_{b,i}$ to the relay nodes, respectively. $\mathbf{n}_R \in \mathbb{C}^{M\times 1}$ denotes the complex Gaussian noise vector of relay nodes with the distribution $\mathcal{CN}(\mathbf{0}, \sigma_r^2\mathbf{I})$. Then, each relay node amplifies the received signal to generate the transmit signal $\mathbf{r}_T$ as

$$\mathbf{r}_T = \mathbf{W}\mathbf{r}, \tag{3}$$

where $\mathbf{W} \in \mathbb{C}^{M\times M}$ is diagonal, and $\mathbf{r}_T$ is subject to a total power constraint P_R.

In the second time slot, the relay nodes broadcast the scaled versions of the received signals to all users. Let $\mathbf{y}_{a,i}$ and $\mathbf{y}_{b,i}$ represent the signal received at the user node $X_{a,i}$ and $X_{b,i}$, respectively. Due to the reciprocal channel assumption, we have

$$\begin{aligned}\mathbf{y}_{a,i} =& \underbrace{\mathbf{H}_i^T\mathbf{W}\mathbf{G}_i\mathbf{b}_i x_{b,i}}_{\text{Desired signal}} + \underbrace{\mathbf{H}_i^T\mathbf{W}\mathbf{H}_i\mathbf{a}_i x_{a,i}}_{\text{Self Interference}} + \underbrace{\mathbf{H}_i^T\mathbf{W}\mathbf{n}_R + \mathbf{n}_{a,i}}_{\text{Noise}} \\ &+ \underbrace{\mathbf{H}_i^T\mathbf{W}\sum_{j\neq i}^{K}(\mathbf{H}_j\mathbf{a}_j x_{a,j} + \mathbf{G}_j\mathbf{b}_j x_{b,j})}_{\text{IPI}},\end{aligned} \tag{4}$$

$$\begin{aligned}\mathbf{y}_{b,i} =& \underbrace{\mathbf{G}_i^T\mathbf{W}\mathbf{H}_i\mathbf{a}_i x_{a,i}}_{\text{Desired signal}} + \underbrace{\mathbf{G}_i^T\mathbf{W}\mathbf{G}_i\mathbf{b}_i x_{b,i}}_{\text{Self Interference}} + \underbrace{\mathbf{G}_i^T\mathbf{W}\mathbf{n}_R + \mathbf{n}_{b,i}}_{\text{Noise}} \\ &+ \underbrace{\mathbf{G}_i^T\mathbf{W}\sum_{j\neq i}^{K}(\mathbf{H}_j\mathbf{a}_j x_{a,j} + \mathbf{G}_j\mathbf{b}_j x_{b,j})}_{\text{IPI}},\end{aligned} \tag{5}$$

where $\mathbf{n}_{a,i}, \mathbf{n}_{b,i} \in \mathbb{C}^{N\times 1}$ denote the complex Gaussian noise vectors of user node $X_{a,i}$ and $X_{b,i}$, respectively, with the distribution $\mathcal{CN}(\mathbf{0}, \sigma_u^2\mathbf{I})$. Here the expressions for the desired signal, self interference (SI), IPI and noise are obtained. Since each user knows its own transmitted signal, the SI signal can be removed from $\mathbf{y}_{a,i}$ and $\mathbf{y}_{b,i}$ through some standard adaptive filtering techniques and for simplicity, we will omit them in the following derivation. The estimated desired symbol after cancelling SI and applying receive beamforming can be expressed as

$$\bar{y}_{a,i} = \mathbf{c}_i^H\mathbf{y}_{a,i}, \quad \bar{y}_{b,i} = \mathbf{d}_i^H\mathbf{y}_{b,i}, \tag{6}$$

where $\mathbf{c}_i, \mathbf{d}_i \in \mathbb{C}^{N\times 1}$ denote the beamforming vectors, and they are assumed to be unit vectors in our work ($||\mathbf{c}_i||^2 = 1, ||\mathbf{d}_i||^2 = 1$).

III. Problem Formulation

In the following, two transceiver beamforming designs will be proposed for the multipair two-way distributed relay beamforming network. In our first design, an iterative zero-forcing-based scheme is proposed aiming at eliminating the IPI, where an iterative algorithm is used to achieve coordination of beamforming vectors of the two user groups. In the second design, it is focused on an iterative transceiver beamforming scheme by maximizing the SINR at each user node.

A. Iterative Zero-Forcing Design

In order to derive the expression for IPI, we first define the overall uplink channel matrix of the IPI (containing the transmit beamforming vectors) of the *i*th user pair as $\mathbf{\Omega}_i \in \mathbb{C}^{M\times 2K-2}$, which is given by

$$\widetilde{\mathbf{\Omega}}_i = [\mathbf{\Omega}_1 \ \cdots \mathbf{\Omega}_{i-1} \ \mathbf{\Omega}_{i+1} \ \cdots \mathbf{\Omega}_K], \tag{7}$$

where $\mathbf{\Omega}_i = [\mathbf{H}_i\mathbf{a}_i \ \ \mathbf{G}_i\mathbf{b}_i] \in \mathbb{C}^{M\times 2}$ is the uplink channel matrix of the *i*th pair. Then from (4), (5), (6) and (7), we can obtain the IPI signal received at the *i*th user pair as

$$\begin{aligned} y_{a,i}^{IPI} &= \mathbf{c}_i^H\mathbf{H}_i^T\mathbf{W}\widetilde{\mathbf{\Omega}}_i\widetilde{\mathbf{x}}_i, \\ y_{b,i}^{IPI} &= \mathbf{d}_i^H\mathbf{G}_i^T\mathbf{W}\widetilde{\mathbf{\Omega}}_i\widetilde{\mathbf{x}}_i, \end{aligned} \tag{8}$$

where $\widetilde{\mathbf{x}}_i = [x_{a,1}\ x_{b,1} \cdots x_{a,i-1}\ x_{b,i-1}\ x_{a,i+1}\ x_{b,i+1} \cdots x_{a,K}\ x_{b,K}]$ consists of all the transmit user symbols other than those coming from the ith user pair.

According to (8), in order to completely eliminate the IPI, $\mathbf{c}_i$ and $\mathbf{d}_i$ should lie in the null space of $\mathbf{H}_i^T \mathbf{W} \widetilde{\mathbf{\Omega}}_i$ and $\mathbf{G}_i^T \mathbf{W} \widetilde{\mathbf{\Omega}}_i$, respectively. And the null space exists when the condition that $N > 2K - 2$ is satisfied. We can define the singular value decomposition (SVD) of the two matrix products as

$$\begin{aligned} \mathbf{\Psi}_{X_{a,i}} &= \mathbf{H}_i^T \mathbf{W} \widetilde{\mathbf{\Omega}}_i = [\mathbf{U}_{X_{a,i}}^{(1)} \mathbf{U}_{X_{a,i}}^{(0)}] \Sigma_{X_{a,i}} \mathbf{V}_{X_{a,i}}^H, \\ \mathbf{\Psi}_{X_{b,i}} &= \mathbf{G}_i^T \mathbf{W} \widetilde{\mathbf{\Omega}}_i = [\mathbf{U}_{X_{b,i}}^{(1)} \mathbf{U}_{X_{b,i}}^{(0)}] \Sigma_{X_{b,i}} \mathbf{V}_{X_{b,i}}^H, \end{aligned} \tag{9}$$

where $\mathbf{U}_{X_{a,i}}^{(1)}$ and $\mathbf{U}_{X_{b,i}}^{(1)}$ hold the left singular vectors of non-zero singular values of the corresponding left-hand-side matrices, while $\mathbf{U}_{X_{a,i}}^{(0)}$ and $\mathbf{U}_{X_{b,i}}^{(0)}$ hold the left singular vectors of zero singular values of $\mathbf{\Psi}_{X_{a,i}}$ and $\mathbf{\Psi}_{X_{b,i}}$, respectively.

To cancel IPI completely, for the receive beamforming vectors $\mathbf{c}_i$ and $\mathbf{d}_i$ we can choose any column vectors of $\mathbf{U}_{X_{a,i}}^{(0)}$ and $\mathbf{U}_{X_{b,i}}^{(0)}$. However, the undetermined transmit beamforming vectors $\mathbf{a}_i$ and $\mathbf{b}_i$ will affect the values of $\mathbf{U}_{X_{a,i}}^{(0)}$ and $\mathbf{U}_{X_{b,i}}^{(0)}$, and we also need to find appropriate values for $\mathbf{a}_i$ and $\mathbf{b}_i$ for a complete solution. To avoid iteration, an effective method is to apply the eigen-beamforming approach at the transmitter side. In detail, $\mathbf{a}_i$ and $\mathbf{b}_i$ are generated as the eigenvectors corresponding to the largest eigenvalues of $\mathbf{H}_i^H \mathbf{H}_i$ and $\mathbf{G}_i^H \mathbf{G}_i$, respectively.

However, to obtain a better performance, $\mathbf{a}_i$ and $\mathbf{b}_i$ should maximize the real equivalent channel gain, taking into consideration the effect of $\mathbf{c}_i$ and $\mathbf{d}_i$. From (4), (5) and (6), we can formulate the problem as follows,

$$\begin{aligned} \max_{\mathbf{b}_i} \quad & C_{X_a,i} = |\mathbf{c}_i^H \mathbf{H}_i^T \mathbf{W} \mathbf{G}_i \mathbf{b}_i|^2, \\ s.t. \quad & ||\mathbf{b}_i||^2 \le P_s, \\ \max_{\mathbf{a}_i} \quad & C_{X_b,i} = |\mathbf{d}_i^H \mathbf{G}_i^T \mathbf{W} \mathbf{H}_i \mathbf{a}_i|^2, \\ s.t. \quad & ||\mathbf{a}_i||^2 \le P_s, \end{aligned} \tag{10}$$

where $C_{X_{a,i}}$ and $C_{X_{b,i}}$ represents the overall equivalent channel gain for the desired signal received at user nodes $X_{a,i}$ and $X_{b,i}$, respectively. It is difficult to derive a closed-form solution for (9) and (10), and here we propose an iterative algorithm to alternately optimize the transmit and receiver beamforming vectors, and make sure no update is required during the iteration for either the relay node or the user node from the other group.

To start with, we employ the uniform AF mode at the relay node.

$$\mathbf{W} = \lambda_R \cdot \mathbf{I}_M, \tag{11}$$

where $\mathbf{I}_M \in \mathbb{C}^{M \times M}$ is the unity matrix, and λ_R is a power-control scalar resulting from the total relay power constraint, which can be expressed as

$$\lambda_R = \sqrt{\frac{P_R}{tr(\mathbf{H}_i \mathbf{a}_i \mathbf{a}_i^H \mathbf{H}_i^H + \mathbf{G}_i \mathbf{b}_i \mathbf{b}_i^H \mathbf{G}_i^H + \sigma_r^2 \cdot \mathbf{I}_M)}}\ . \tag{12}$$

Note that the value of λ_R does not affect the solution of (9) and (10), we can consider it at the final step of our iteration process. First, the initial value of the receive beamforming vectors $\mathbf{c}_i$ and $\mathbf{d}_i$ are assigned values as $[\delta_M \delta_M \cdots \delta_M] \in \mathbb{C}^{1 \times N}$, where $\delta_M = \sqrt{M}$. Then we can calculate $\mathbf{a}_i$ and $\mathbf{b}_i$ at each user node based on (10), given by

$$\mathbf{a}_i = \lambda_{a,i} \cdot \mathbf{H}_i^H \mathbf{G}_i^* \mathbf{d}_i, \quad \mathbf{b}_i = \lambda_{b,i} \cdot \mathbf{G}_i^H \mathbf{H}_i^* \mathbf{c}_i, \tag{13}$$

where $\lambda_{a,i}$ and $\lambda_{b,i}$ are the power-control scalars resulting from the transmit power constraint, which are given as

$$\lambda_{a,i} = \sqrt{\frac{P_S}{||\mathbf{H}_i^H \mathbf{G}_i^* \mathbf{d}_i||^2}}, \quad \lambda_{b,i} = \sqrt{\frac{P_S}{||\mathbf{G}_i^H \mathbf{H}_i^* \mathbf{c}_i||^2}}\ . \tag{14}$$

Next, the updated values of $\mathbf{c}_i$ and $\mathbf{d}_i$ can be obtained at each user node from $\mathbf{U}_{X_{a,i}}^{(0)}$ and $\mathbf{U}_{X_{b,i}}^{(0)}$ in (9). This update process keeps going until some predefined stopping criterion is met, such as a preset maximum iteration number or convergence requirement. When the final updates of the beamforming vectors are obtained, the power-control scalar λ_R is decided from (12). The iterative zero-forcing method is summarized in **Iteration Algorithm Summary**.

Although the iterative zero-forcing method can not guarantee a globally optimum solution due to the non-convexity of the problem, it still outperforms the non-iterative zero-forcing method significantly, as will be shown in our simulations.

B. Iterative Algorithm for SINR Optimizing

The proposed iterative zero-forcing method can completely eliminate the IPI signal received at each user node. However, such a beamformer may lead to undesired amplification of noise, degrading the overall performance. In this section, we propose an iterative algorithm aiming at maximizing the SINR at each user node, which has a better performance compared to the zero-forcing based one.

Without loss of generality, we take user $X_{a,i}$ as an example. From (4) and (6), the SINR at this user node can be expressed as

$$SINR_{a,i} = \frac{\mathbf{c}_i^H \mathbf{H}_i^T \mathbf{Q}_{a,i}^{(S)} \mathbf{H}_i^* \mathbf{c}_i}{\sigma_u^2 + \mathbf{c}_i^H \mathbf{H}_i^T \mathbf{Q}_{a,i}^{(N)} \mathbf{H}_i^* \mathbf{c}_i + \mathbf{c}_i^H \mathbf{H}_i^T \mathbf{Q}_{a,i}^{(I)} \mathbf{H}_i^* \mathbf{c}_i}\ , \tag{15}$$

where,

$$\begin{aligned} \mathbf{Q}_{a,i}^{(N)} &= \lambda_R^2 \sigma_r^2 \cdot \mathbf{I}_M, \\ \mathbf{Q}_{a,i}^{(S)} &= \lambda_R^2 Ps \cdot \mathbf{G}_i \mathbf{b}_i \mathbf{b}_i^H \mathbf{G}_i^H, \\ \mathbf{Q}_{a,i}^{(I)} &= \lambda_R^2 Ps \cdot \sum_{j \neq i}^{K} (\mathbf{H}_j \mathbf{a}_j \mathbf{a}_j^H \mathbf{H}_j^H + \mathbf{G}_j \mathbf{b}_j \mathbf{b}_j^H \mathbf{G}_j^H). \end{aligned} \tag{16}$$

Apparently, if we only need to consider user node $X_{a,i}$, an ideal way to maximize the $SINR$ is to completely eliminate the IPI by $\mathbf{a}_j$ and $\mathbf{b}_j$ ($j = 1 \cdots K, j \neq i$), and maximize the remaining part by $\mathbf{c}_i$ and $\mathbf{b}_i$. However, the optimal choice of $\mathbf{a}_j$ and $\mathbf{b}_j$ for user node $X_{a,i}$ will unlikely result in an optimal $SINR$ for other user nodes. In fact, it is very difficult, if not impossible, to obtain an analytical global solution for maximizing $SINR$ at every user node for this transceiver beamforming scenario.

As an alternative, we propose an iterative algorithm which can achieve a desirable sub-optimal SINR, while being performed locally at each user node.

First, we initialize the beamforming vectors $\mathbf{c}_i$ and $\mathbf{d}_i$ as unity vectors $[\delta_M \delta_M \cdots \delta_M] \in \mathbb{C}^{1\times N}$, where $\delta_M = \sqrt{M}$. Note that in practice, this initialization step may not be necessary, since the update process can always continue as long as the transmission keeps going, and when the channel state change slowly, the iteration number required to achieve convergence can be further reduced.

Then we update $\mathbf{a}_i$ and $\mathbf{b}_i$ based on maximizing power of the desired signal received at each user node, which is also the numerator of the $SINR$ expression. For user node $X_{a,i}$, the SINR expression is given in (15) and (16), and the case for user node $X_{b,i}$ is similar. Applying the individual transmit power constraint, after some simple derivations, we can express the updated values for the two transmit beamforming vectors as

$$\mathbf{a}_i = \lambda_{a,i} \cdot \mathbf{H}_i^H \mathbf{G}_i^* \mathbf{d}_i, \quad \mathbf{b}_i = \lambda_{b,i} \cdot \mathbf{G}_i^H \mathbf{H}_i^* \mathbf{c}_i, \tag{17}$$

which are the same as (13) in the earlier scheme, and $\lambda_{a,i}$ and $\lambda_{b,i}$ have been defined in (14). Next, the following SINR optimization problem for user node $X_{a,i}$ can be solved locally to obtain the receive beamforming vector $\mathbf{c}_i$.

$$\begin{aligned} \max_{\mathbf{c}_i} \quad & SINR_{a,i} = \mathbf{c}_i^H \mathbf{\Theta}_{a,i} \mathbf{c}_i, \\ s.t. \quad & ||\mathbf{c}_i||^2 = 1, \end{aligned} \tag{18}$$

where

$$\begin{aligned} \mathbf{\Theta}_{a,i} &= (\mathbf{\Xi}_{a,i})^{-1} \mathbf{H}_i^T \mathbf{Q}_{a,i}^{(S)} \mathbf{H}_i^*, \\ \mathbf{\Xi}_{a,i} &= \sigma_u^2 \mathbf{I}_N + \mathbf{H}_i^T \mathbf{Q}_{a,i}^{(N)} \mathbf{H}_i^* + \mathbf{H}_i^T \mathbf{Q}_{a,i}^{(I)} \mathbf{H}_i^*. \end{aligned} \tag{19}$$

Then the closed-form solution to this eigenvector problem leads to the updated value for $\mathbf{c}_i$, and similarly for $\mathbf{d}_i$ as well, as expressed in the following

$$\mathbf{c}_i = \rho\{\mathbf{\Theta}_{a,i}\}, \quad \mathbf{d}_i = \rho\{\mathbf{\Theta}_{b,i}\}, \tag{20}$$

where $\rho\{\cdot\}$ denotes the principle eigenvector of a matrix.

As summarized in **Iteration Algorithm Summary**, this iteration is repeated until reaching some stopping criterion, which can be defined by a preset maximum iteration number or convergence requirement. The relay nodes weights with the power-control scalar λ_R is decided from (11) and (12) at the final step.

Iteration Algorithm Summary

Iterative Zero-Forcing:
1) Initialization: $\mathbf{c}_i, \mathbf{d}_i = [\delta_M \delta_M \cdots \delta_M] \in \mathbb{C}^{1\times N}$, where $\delta_M = \sqrt{M}$, and set t=1.
2) Update $\mathbf{a}_i$ and $\mathbf{b}_i$ based on (13) and (14).
3) Update $\mathbf{c}_i$ and $\mathbf{d}_i$ based on $\mathbf{U}_{X_{a,i}}^{(0)}$ and $\mathbf{U}_{X_{b,i}}^{(0)}$ in (9).
4) If $|\mathbf{x}_i^{(t)} - \mathbf{x}_i^{(t-1)}|/\mathbf{x}_i^{(t)} < \varepsilon$ or $t > n$ (ε is a predetermined value for convergence check of the iterative process, $\mathbf{x} \leftarrow \mathbf{c}$ for users from group X_a and $\mathbf{x} \leftarrow \mathbf{d}$ for users from group X_b), go to the next step. Otherwise, $t = t + 1$ and go back to step 2).
5) Decide $\mathbf{W}$ based on (11) and (12).

Iterative SINR Optimization:
1) Initialization: $\mathbf{c}_i, \mathbf{d}_i = [\delta_M \delta_M \cdots \delta_M] \in \mathbb{C}^{1\times N}$, where $\delta_M = \sqrt{M}$, and set t=1.
2) Update $\mathbf{a}_i$ and $\mathbf{b}_i$ based on (17) and (14).
3) Update $\mathbf{c}_i$ and $\mathbf{d}_i$ based on (19) and (20).
4) If $|\mathbf{x}_i^{(t)} - \mathbf{x}_i^{(t-1)}|/\mathbf{x}_i^{(t)} < \varepsilon$ or $t > n$ (ε is a predetermined value for convergence check of the iterative process, $\mathbf{x} \leftarrow \mathbf{c}$ for users from group X_a and $\mathbf{x} \leftarrow \mathbf{d}$ for users from group X_b), go to the next step. Otherwise, $t = t + 1$ and go back to step 2).
5) Decide $\mathbf{W}$ based on (11) and (12).

Discussions: For both algorithms, knowledge of all the receive beamforming vectors $\mathbf{c}_i$ and $\mathbf{d}_i$ is required to update $\mathbf{a}_j$ and $\mathbf{b}_j$ for the *j*th user pair ($j = 1, ..., K$). They can all be calculated by each user within the *j*th user pair to avoid any communication with any other users, or shared within each user group (X_a and X_b) using limited backhaul resources to reduce the computation complexity. Another way to reduce the computation complexity is to utilize a central node (it can be one of the users) for each user group (i.e. one central node for X_a, and one central node for X_b) to perform all the calculations and inform the users in its own group the resultant beamforming vectors. The computational complexity of the second algorithm is higher than the first one, since the calculations of $\mathbf{c}_i$ and $\mathbf{d}_i$ are more complicated for it. The number of multiplications needed for updating $\mathbf{c}_i$ or $\mathbf{d}_i$ using the iterative zero-forcing algorithm and the SINR optimizing algorithm is roughly $N(2K-2)(2M-1) + \mathcal{O}(N(2K-2)^2)$ and $6M^2N + 6MN^2 - 3MN - 2N^2 + \mathcal{O}(N^3)$, respectively. As an example, for $M = N = 2K - 2 = \Lambda$, this number is $(2\Lambda^3 - 2\Lambda^2 + \mathcal{O}(\Lambda^3))$ and $(12\Lambda^3 - 5\Lambda^2 + \mathcal{O}(\Lambda^3))$, respectively.

IV. Simulation results

In this section, numerical results are provided demonstrating the performance of the two proposed transceiver beamforming strategies for multipair two-way distributed relay networks. The channels are assumed to be i.i.d. Rayleigh fading, i.e., the elements of each channel vector are complex Gaussian random variables with zero mean and unit variance. We also assume that the transmit power P_S is normalized to 1 (compensating the unconsidered path-loss), and the noise powers at all nodes are identical to 1 ($\sigma_r^2 = \sigma_u^2 = 1$). The SNR_R is defined to

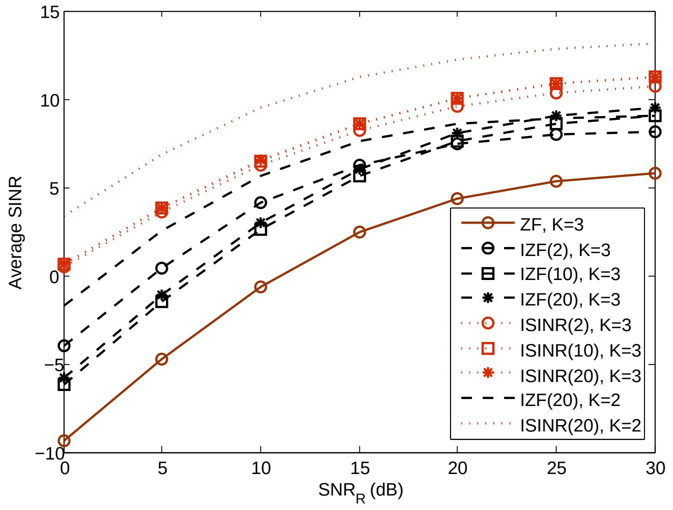

Fig. 2. SINR performance of ZF, IZF and ISINR method with different iteration number (M=6, N=5, K=2, 3).

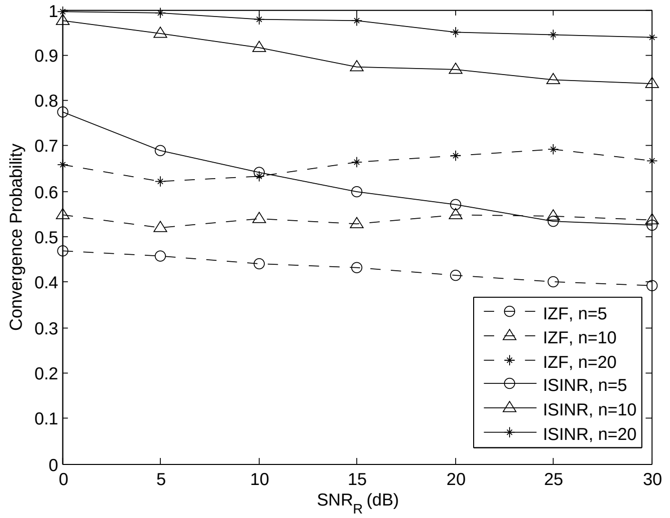

Fig. 3. Convergence performance of the IZF and ISINR methods with different iteration number (M=6, N=5, K=3).

be the ratio of relay node output power to the noise variance, i.e., $SNR_R = P_R/\sigma_r^2$. The value of $\varepsilon = 0.01$ is chosen to determine the convergence of the iterative process.

In Fig. 2, we present the average SINR performance of the proposed iterative zero-forcing method (denoted by "IZF") and the iterative SINR optimization method (denoted by "ISINR"), with $M = 6$, $N = 5$ and $K = 2, 3$. The performance of the two methods with different iteration numbers are provided in comparison with the non-iterative zero-forcing method (denoted by "ZF"). As can be seen, our proposed iterative methods have outperformed the non-iterative zero-forcing method with only 2 iterations, especially for the iterative SINR optimization method, where the improvement is more significant. Clearly, although the iterative SINR optimization method will not necessarily result in the optimum SINR, performance improvement has been achieved for all iteration number settings; moreover, when the iteration number is increased to 10, the SINR performance is further enhanced. However, further increase of the iteration number leads to much less gain in the result and considering the associated cost for each iteration, the iteration process can then be stopped.

Next, we study the convergence performance of the two schemes in terms of their convergence probability, which is defined as the probability of a simulation result meeting the convergence requirement. The results with different preset maximum iteration number and SNR_R are shown in Fig. 3. We can see that the convergence probability of the iterative SINR optimization scheme is always better than the corresponding IZF scheme, especially when the SNR_R is low. As SNR_R increases, the convergence probability of the second scheme decreases; meanwhile the iterative zero-forcing scheme is not much affected. When the iteration number is large enough, the influence of SNR_R becomes less significant for the iterative SINR optimization scheme. Combined with Fig. 2, it also indicates that the improvement of SINR performance does not necessarily require the scheme to converge. Moreover, in some cases, the beamforming vectors will keep swapping between two values, both of which will lead to a similar and desirable SINR.

V. Conclusions

In this paper, the transceiver beamforming problem for multipair two-way distributed relay networks has been studied, where the relay nodes are employed with very simple settings and all the computation processes and the main signal processing procedures are performed at the user nodes. In order to achieve a desirable performance, the transmit and receive beamforming vectors from the two separated user groups are designed using two iterative methods, where the first one aims to eliminate the IPI and the second one considers maximizing the SINR at each user node. Both of them can be performed locally at each user node. However, if data exchange within the same group is allowed, utilization of limited backhaul resources can lead to reduction of the computational complexity. Simulations have been provided to evaluate the performance of the two transceiver beamforming designs in terms of both SINR and convergence speed, and the results indicate that both work effectively and can achieve a better performance with a small iteration number compared to the existing zero-forcing scheme.

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