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A Distributed Iterative Transceiver Beamforming Algorithm for Multipair Two-Way Relay Networks

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***Abstract*—The transceiver beamforming design problem is studied in this paper for a multi-pair two-way distributed relay network, where multi-antenna users in one user group communicate with their partners in the other user group via distributed single-antenna relay nodes. An iterative algorithm is proposed where transmit and receive beamformings are performed at user nodes, and relay nodes have their own simple strategies for deciding the weights. The computation tasks are distributed among each user and relay node, through which high computation efficiency can be ensured. By coordinating them together, satisfactory performance is obtained when relay number is low and significant performance enhancement is also achieved for a large relay number.**

I. Introduction

Due to its advantages in coverage extension, mitigating the effect of fading and enhancement of network throughput, distributed relay assisted networks have attracted much attention in the past decade [1–6]. In such networks, distributed relay nodes create a virtual multiple-input multiple-output (MIMO) environment, where beamforming techniques could be applied to regulate the performance of the network.

Multipair relay network is one of the considered configurations, where multiple source nodes simultaneously communicate with their destination nodes, leading to great advantage in spatial efficiency and overall throughput [7–10]. In [7, 9, 11], multipair two-way relay network with a central multi-antenna relay node was studied, with zero-focring (ZF) based solutions proposed to cancel out the inter-pair interference (IPI). In [12, 13], block diagonalization (BD) was employed at the central relay node to reduce the IPI. The work in [8] studied the distributed single-antenna relay networks with multipair two-way communication, where a relatively complicated ZF method was applied to eliminate the IPI completely and guided the relay weights setting, while in [10], a similar network was considered, and the implementation of the relay nodes was simplified. However, both methods require a very large relay number.

Reducing IPI and noise in a multipair two-way relay network requires relatively heavy task of computation. In the aforementioned designs, the computation tasks were globally performed, and were assigned to either the users side, or a central relay node. In some of the schemes, the same computation process has to be repeated at each user. Motivated by this issue, we considered an iterative transceiver beamforming design for multi-pair two-way relay networks in [14], where the iteration process was performed and completed at the user nodes to optimize the SINR performance at each user. The main drawback of that design was that almost all the signal processing and computation were shifted to the transmit/receive pairs, and although it relieves the relay nodes of their computation tasks greatly, the trade-off is the unavoidable system performance loss. In this paper, we distribute the iteration process to all the users and the relay nodes, where the main computation tasks are assigned to each of the user nodes. In particular, an amplify-and-forward (AF) strategy is designed for the relay nodes which allows them to decide their own weights with simple computation process using their local channel state information (CSI) only.

Notations: $[\cdot]^T$, $[\cdot]^H$ and $[\cdot]^*$ stand for transpose, Hermitian transpose and conjugate, respectively. $||\cdot||$ denotes the Frobenius norm of a vector and $|\cdot|$ the absolute value of a scalar. $\mathbb{E}[\cdot]$ represents the expectation operator and Var$[\cdot]$ the variance operator. $\mathbf{I}_N$ is the $N \times N$ identity matrix.

II. System Model And Problem Description

Consider a time-slotted dual-hop multipair two-way distributed relay network consisting of $2K$ multi-antenna users (antenna number = N) which forms K communication pairs (X_a, X_b), and the transmission of information streams takes place in both directions and in two transmission phases assisted by M single-antenna relay nodes, as shown in Fig. 1. We assume that there is no direct link between any user pairs.

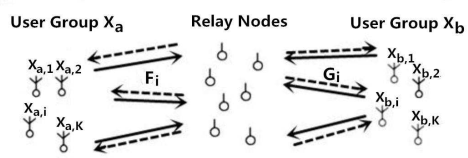

Fig. 1. The considered dual-hop multipair two-way relay network.

In phase I, all users simultaneously transmit their information streams to the relay nodes with transmit beamforming vectors $\mathbf{a}_i$ and $\mathbf{b}_i(\in \mathbb{C}^{N\times 1}$, i=1, ..., K), for each user from group X_a and X_b, respectively. Total transmit power constraint puts a limit on the transmit beamforming vectors by $||\mathbf{a}_i||^2 \leq P_S$ and $||\mathbf{b}_i||^2 \leq P_S$, with P_S being the maximally allowed transmitted signal power. In phase II, the distributed relay nodes amplify-and-forward the information streams back to the users with a set of weights, denoted by w_m for the m-*th* relay nodes, $m = 1, ..., M$. Following that, the received signal undergoes receive beamforming, denoted by two $N \times 1$ normalized vectors $\mathbf{c}_i$ and $\mathbf{d}_i$ ($||\mathbf{c}_i||^2$=$||\mathbf{d}_i||^2$=1), at $X_{a,i}$ and $X_{b,i}$ sides, respectively.

Let $\mathbf{F}_i, \mathbf{G}_i \in \mathbb{C}^{M\times N}$ represent the channel matrix from $X_{a,i}$ and $X_{b,i}$ to the relay nodes, respectively, with $f_{m,n,i}$ and $g_{m,n,i}$ (m=1, ..., M, n=1, ..., N) being the (m,n)-*th* element. We assume a rich-scattering environment with independent reciprocal channels modeled as Rayleigh fading with the distribution $f_{m,n,i} \sim \mathcal{CN}(0,1)$ and $g_{m,n,i} \sim \mathcal{CN}(0,1)$.

Then the signals received at the relay nodes at phase I can be represented by $\mathbf{r} \in \mathbb{C}^{M\times 1}$

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

where $x_{a,i}$ and $x_{b,i}$ denote the data symbol and $\mathbf{n}_R \in \mathbb{C}^{M\times 1}$ is the complex Gaussian noise vector of relay nodes with the distribution $\mathcal{CN}(\mathbf{0}, \sigma_r^2 \mathbf{I})$.

Then, the signals are scaled at relay nodes using the AF protocol, which is given by

$$\mathbf{r}_T = \mathbf{W}\mathbf{r}, \tag{2}$$

where $\mathbf{W} \in \mathbb{C}^{M\times M}$ is diagonal, with w_m being the (m,m)-*th* element of $\mathbf{W}$.

Then, in phase II, the received signal at $X_{a,i}$ can be expressed as (expression for $X_{b,i}$ is similar)

$$\begin{aligned} y_{a,i} = & \underbrace{\mathbf{c}_i \mathbf{F}_i^T \mathbf{W} \mathbf{G}_i \mathbf{b}_i x_{b,i}}_{\text{Desired signal}} + \underbrace{\mathbf{c}_i \mathbf{F}_i^T \mathbf{W} \mathbf{F}_i \mathbf{a}_i x_{a,i}}_{\text{Self Interference}} + \mathbf{c}_i \mathbf{F}_i^T \mathbf{W} \mathbf{n}_R \\ & + \mathbf{c}_i \mathbf{n}_{a,i} + \underbrace{\mathbf{c}_i \mathbf{F}_i^T \mathbf{W} \sum_{j\neq i}^{K} (\mathbf{F}_j \mathbf{a}_j x_{a,j} + \mathbf{G}_j \mathbf{b}_j x_{b,j})}_{\text{IPI}}, \end{aligned} \tag{3}$$

where $\mathbf{n}_{a,i} \in \mathbb{C}^{N\times 1}$ is the additive white complex Gaussian noise vector at the user node, with the distribution $\mathcal{CN}(\mathbf{0}, \sigma_u^2 \mathbf{I})$. Since the user knows its own transmitted signal, the self interference (SI) in (3) can be removed through some standard adaptive filtering techniques. For simplicity, they are ignored in the following derivation.

III. Distributed Iterative Beamforming Algorithm for SINR Optimization

In this section, motivated by the iSINR method proposed in [14], a distributed iteration algorithm for SINR optimization (noted as *distributed iSINR*) is proposed, where the iteration is divided into three parts: the transmitter part, the relay part and the receiver part. And the computation performed at each user node and relay node will only update their own beamforming weights. Therefore, the power usage for performing the required tasks is much more efficient.

Taking user $X_{a,i}$ as an example. From (3), the SINR at this user can be expressed as follows,

$$SINR_{a,i} = \frac{\mathbf{c}_i^H \mathbf{F}_i^T \mathbf{Q}_{a,i}^{(S)} \mathbf{F}_i^* \mathbf{c}_i}{\sigma_u^2 + \mathbf{c}_i^H \mathbf{F}_i^T \mathbf{Q}_{a,i}^{(N)} \mathbf{F}_i^* \mathbf{c}_i + \underbrace{\mathbf{c}_i^H \mathbf{F}_i^T \mathbf{Q}_{a,i}^{(I)} \mathbf{F}_i^* \mathbf{c}_i}_{\text{IPI}}}, \tag{4}$$

where,

$$\begin{aligned} \mathbf{Q}_{a,i}^{(I)} &= Ps \cdot \sum_{j\neq i}^{K} (\mathbf{W}\mathbf{F}_j \mathbf{a}_j \mathbf{a}_j^H \mathbf{F}_j^H \mathbf{W}^H + \mathbf{W}\mathbf{G}_j \mathbf{b}_j \mathbf{b}_j^H \mathbf{G}_j^H \mathbf{W}^H), \\ \mathbf{Q}_{a,i}^{(N)} &= \sigma_r^2 \cdot \mathbf{W}\mathbf{W}^H, \ \mathbf{Q}_{a,i}^{(S)} = Ps \cdot \mathbf{W}\mathbf{G}_i \mathbf{b}_i \mathbf{b}_i^H \mathbf{G}_i^H \mathbf{W}^H. \end{aligned} \tag{5}$$

As can be seen, if maximizing $SINR_{a,i}$ is the only objective, $\mathbf{a}_j$ and $\mathbf{b}_j$ ($j = 1\cdots K, j \neq i$) could be carefully chosen to completely eliminate the IPI part, and the remaining part can be maximized by $\mathbf{c}_i$ and $\mathbf{b}_i$. However, the optimal choice of $\mathbf{a}_j$ and $\mathbf{b}_j$ for user $X_{a,i}$ will unlikely result in a sufficiently good SINR for other users, as the beamforming vectors of one user not only affects its own SINR, but also others. In fact, it is very difficult, if not impossible, to obtain an analytical solution for maximizing SINR at all user nodes for this transceiver beamforming scenario.

Therefore, as an alternative, an iterative process composed of the three parts mentioned earlier is employed to achieve a sub-optimal SINR.

A. Iteration Step on the Transmit Part

Throughout this paper, we assume that the CSI is either estimated at the user or fed back to it by the relay nodes via low rate feedback channels, so that the beamforming vectors can be decided at the user nodes.

The first iteration step is applied to the user nodes to decide their transmit beamforming vectors $\mathbf{a}_i$ and $\mathbf{b}_i$, for user $X_{a,i}$ and $X_{b,i}$, respectively. At this step, the receive beamforming vectors $\mathbf{c}_i$, $\mathbf{d}_i$ and relay weights $\mathbf{W}$ are fixed to an updated value through previous steps; otherwise, an initial value should be assigned to them. Then, we try to optimize $\mathbf{a}_i$ and $\mathbf{b}_i$ based on maximizing the power of the desired signal received at $X_{a,i}$ and $X_{b,i}$, respectively, under a transmit power constraint.

$$\begin{aligned} &\max_{\mathbf{b}_i} \ |\mathbf{c}_i^H \mathbf{F}_i^T \mathbf{W} \mathbf{G}_i \mathbf{b}_i|^2, \ s.t. \ ||\mathbf{b}_i||^2 \leq P_S, \\ &\max_{\mathbf{a}_i} \ |\mathbf{d}_i^H \mathbf{G}_i^T \mathbf{W} \mathbf{F}_i \mathbf{a}_i|^2, \ s.t. \ ||\mathbf{a}_i||^2 \leq P_S. \end{aligned} \tag{6}$$

These two problems have closed-form solutions, given by

$$\mathbf{a}_i = \lambda_{a,i} \cdot \mathbf{F}_i^H \mathbf{W}^H \mathbf{G}_i^* \mathbf{d}_i, \quad \mathbf{b}_i = \lambda_{b,i} \cdot \mathbf{G}_i^H \mathbf{W}^H \mathbf{F}_i^* \mathbf{c}_i, \tag{7}$$

where $\lambda_{a,i}$ and $\lambda_{b,i}$ are the power-control scalars

$$\lambda_{a,i} = \sqrt{\frac{P_S}{||\mathbf{F}_i^H \mathbf{W}^H \mathbf{G}_i^* \mathbf{d}_i||^2}}, \quad \lambda_{b,i} = \sqrt{\frac{P_S}{||\mathbf{G}_i^H \mathbf{W}^H \mathbf{F}_i^* \mathbf{c}_i||^2}}. \tag{8}$$

The obtained transmit beamforming vectors should be forwarded to their user pairs through the relay nodes in order to perform the updates of the other beamforming weights. Until receiving an update for $\mathbf{c}_i$ and $\mathbf{d}_i$, the transmit beamforming vectors should remain constant.

B. Iteration Step on the Relay Part

The second step is applied to the relay nodes where $\mathbf{c}_i$, $\mathbf{d}_i$, $\mathbf{a}_i$ and $\mathbf{b}_i$ are fixed to their previously updated value. Let $\mathbf{f}_{i,m}, \mathbf{g}_{i,m} \in \mathbb{C}^{1\times N}$ represents the m-*th* row of $\mathbf{F}_i$ and $\mathbf{G}_i$, respectively. We propose the following phase rotating rule for the m-*th* relay node ($m = 1, ..., M$).

$$w_m = \lambda_m(\sum_{i=1}^{K} \mathbf{f}_{i,m}^* \mathbf{c}_i \mathbf{b}_i^H \mathbf{g}_{i,m}^H + \mathbf{g}_{i,m}^* \mathbf{d}_i \mathbf{a}_i^H \mathbf{f}_{i,m}^H)$$
$$= \lambda_m(\sum_{i=1}^{K} \hat{u}_{i,m}^* v_{i,m}^* + \hat{v}_{i,m}^* u_{i,m}^*), \quad (9)$$

where $\hat{u}_{i,m} \triangleq \mathbf{f}_{i,m}^* \mathbf{c}_i$, $u_{i,m} \triangleq \mathbf{f}_{i,m} \mathbf{a}_i$, $\hat{v}_{i,m} \triangleq \mathbf{g}_{i,m}^* \mathbf{d}_i$ and $v_{i,m} \triangleq \mathbf{g}_{i,m} \mathbf{b}_i$. And λ_m is a power-control parameter which limits the output power of each relay node, given by

$$\lambda_m = \sqrt{\frac{P_{R,m}/|\sum_{i=1}^{K} \bar{u}_{i,m}^* v_{i,m}^* + \bar{v}_{i,m}^* u_{i,m}^*|^2}{\sigma_r^2 + \sum_{i=1}^{K} |u_{i,m}|^2 + |v_{i,m}|^2}}, \quad (10)$$

where $P_{R,m}$ is the individual power budget at the m*-th* relay.

As will be observed from the updating process for $\mathbf{c}_i$ in Section III-C, $\mathbf{c}_i$ is not directly determined by $\mathbf{f}_{i,m}$ in our scenario, and in fact their correlation is very weak, especially when M and K are large. As a result, we can consider them as the two independent variables. We have $||\mathbf{c}_i||^2$=1, and accordingly $\hat{u}_{i,m}$ has the distribution of $\mathcal{CN}(0, 1)$, and so are $u_{i,m}$, $\hat{v}_{i,m}$ and $v_{i,m}$.

In order to provide further insight for choosing the phase rotating coefficient on the relay node, we rewrite (3) after removing the self interference part, in terms of $u_{i,m}$, $v_{i,m}$, $\hat{u}_{i,m}$ and $\hat{v}_{i,m}$.

$$\hat{y}_{a,i} = \underbrace{\sum_{m=1}^{M} \hat{u}_{i,m} w_m v_{i,m} x_{b,i}}_{\text{Desired signal}} + \underbrace{\sum_{m=1}^{M} \hat{u}_{i,m} w_m n_{R,m} + n_{a,i}}_{\text{Noise}}$$
$$+ \underbrace{\sum_{m=1}^{M} \sum_{j \neq i}^{K} (\hat{u}_{i,m} w_m u_{j,m} x_{a,j} + \hat{u}_{i,m} w_m v_{j,m} x_{b,j})}_{\text{IPI}}$$
$$= \mathcal{G}_{a,i}^{(S)} x_{b,i} + \mathcal{G}_{a,i}^{(Noise)} n_{R,m} + n_{a,i}$$
$$+ \sum_{j \neq i}^{K} (\mathcal{G}_{ab,ij}^{(IPI)} x_{a,j} + \mathcal{G}_{ba,ij}^{(IPI)} x_{b,j}), \quad (11)$$

where $\mathcal{G}_{a,i}^{(S)}$, $\mathcal{G}_{a,i}^{(Noise)}$, $\mathcal{G}_{ab,ij}^{(IPI)}$ and $\mathcal{G}_{ba,ij}^{(IPI)}$ represents the gain of each component, $n_{R,m}$ represents the complex Gaussian noise of the m*-th* relay node with the distribution $\mathcal{CN}(0, \sigma_r^2)$ and $n_{a,i}$=$\mathbf{d}_i \mathbf{n}_{b,i}$. Due to the fact that in our scheme, $\mathbf{d}_i$ is normalized vector ($||\mathbf{d}_i||^2$=1), $n_{a,i}$ will have a distribution given by $\mathcal{CN}(0, \sigma_u^2)$.

Let $\hat{y}_{a,i}^{(S)}$, $\hat{y}_{a,i}^{(IPI)}$ and $\hat{y}_{a,i}^{(Noise)}$ denote the desired signal, IPI and noise part in (11), respectively. We have

$$\hat{y}_{a,i}^{(S)} = \sum_{m=1}^{M} \lambda_m \hat{u}_{i,m} (\sum_{i=1}^{K} \hat{u}_{i,m}^* v_{i,m}^* + \hat{v}_{i,m}^* u_{i,m}^*) v_{i,m} x_{b,i} \quad (12)$$

Since $\hat{u}_{i,m}$, $\hat{v}_{i,m}$, $\hat{u}_{i',m}(i' \neq i)$ and $\hat{u}_{i,m'}(m' \neq m)$ can be considered as zero mean mutually uncorrelated random variables, with $\mathbb{E}[x^2] = \sigma^2$, where $x \sim \mathcal{CN}(0, \sigma^2)$, we have

$$\mathbb{E}[\mathcal{G}_{a,i}^{(S)}] = \mathbb{E}[\sum_{m=1}^{M} \lambda_m ||\hat{u}_{i,m}||^2 ||v_{i,m}||^2] = \sum_{m=1}^{M} \lambda_m \quad (13)$$

Denote $\gamma_{i,m}$=$\hat{u}_{i,m} w_m v_{i,m}$ for m=$1, ..., M$. Since all $\gamma_{i,m}$ are independent random variables, we can apply the Tchebyshev's inequality theorem [15], and for any constant ζ obtain

$$\Pr[\left| \frac{\mathcal{G}_{a,i}}{M} - \frac{\mathbb{E}[\mathcal{G}_{a,i}]}{M} \right| \geq \zeta] \leq \frac{\mathrm{Var}[\hat{y}_{a,i}^{(S)}]/M^2}{\zeta^2} \quad (14)$$

where $\Pr[\cdot]$ represents the probability operator. Apparently $\hat{y}_{a,i}^{(S)}/M$ will be more likely to approach $\mathbb{E}[\hat{y}_{a,i}^{(S)}/M]$=$x_{a,i}\mathbb{E}[\mathcal{G}_{a,i}]/M$=$\overline{\lambda}_m x_{a,i}$ ($\overline{\lambda}_m$ denotes the average value of λ_m) as M increases. As a result, the asymptotic value of $|\hat{y}_{a,i}^{(S)}|^2$ is proportional to M^2, when M is large.

Similarly, we can derive that $\mathbb{E}[\mathcal{G}_{a,i}^{(Noise)}]$=0, $\mathbb{E}[\mathcal{G}_{ab,ij}^{(IPI)}]$=0 and $\mathbb{E}[\mathcal{G}_{ba,ij}^{(IPI)}]$=0, and when M is large, $\hat{y}_{a,i}^{(IPI)}/M$ and $\hat{y}_{a,i}^{(Noise)}/M$ will have a high probability of taking a value around 0.

In another word, the $\lambda_m \hat{u}_{i,m} \hat{u}_{i,m}^* v_{i,m}^* v_{i,m} x_{b,i}$ part in $\hat{y}_{a,i}^{(S)}$ is the only component in $\hat{y}_{a,i}$ that can grow steadily through accumulation as M increases; meanwhile, the other parts will grow much more slowly. The situation is similar for $\hat{y}_{b,i}$ (received signal at $X_{b,i}$).

C. Iteration Step on Receiver Part

In the third step, based on the updated values of $\mathbf{a}_i$, $\mathbf{b}_i$ and $\mathbf{W}$, we determine the receive beamforming vector $\mathbf{c}_i$ (similar process for $\mathbf{d}_i$) by solving the following SINR optimization problem for the user node $X_{a,i}$. From (4) and (5) we have

$$\max_{\mathbf{c}_i} \quad SINR_{a,i} = \mathbf{c}_i^H \mathbf{\Theta}_{a,i} \mathbf{c}_i, \ s.t. \ ||\mathbf{c}_i||^2 = 1, \quad (15)$$

where

$$\mathbf{\Theta}_{a,i} = (\mathbf{\Xi}_{a,i})^{-1} \mathbf{F}_i^T \mathbf{Q}_{a,i}^{(S)} \mathbf{F}_i^*,$$
$$\mathbf{\Xi}_{a,i} = \sigma_u^2 \mathbf{I}_N + \mathbf{F}_i^T \mathbf{Q}_{a,i}^{(N)} \mathbf{F}_i^* + \mathbf{F}_i^T \mathbf{Q}_{a,i}^{(I)} \mathbf{F}_i^*. \quad (16)$$

This eigenvector problem can be solved locally at each user node with the close-form solution given by

$$\mathbf{c}_i = \rho\{\mathbf{\Theta}_{a,i}\}, \quad \mathbf{d}_i = \rho\{\mathbf{\Theta}_{b,i}\}, \quad (17)$$

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

It can be seen that in order to determine $\mathbf{c}_i$ at user $X_{a,i}$, transmit beamforming vectors of all the other users are required. In our scheme, we assume this information is gathered at the relay node first, and then broadcast to all the users with the relay weights information.

D. Summary of the Distributed Iteration Algorithm

In the proposed distributed iteration algorithm, $\mathbf{a}_i$ and $\mathbf{b}_i$ are first decided, by assigning an initial value for the relay weights and the receive beamforming vectors, as indicated in **Summary of Iteration Steps**. Then, $\mathbf{a}_i$ and $\mathbf{b}_i$ remains fixed until the next round of iteration begins.

Each relay node updates its AF weight based on the proposed strategy, only when it has received the complete set of updated $\mathbf{a}_i$ and $\mathbf{b}_i$. Their updated weights should be broadcasted back to the user nodes, and until the updated values of $\mathbf{c}_i$ and $\mathbf{d}_i$ arrive, their weights remain unchanged.

The user nodes perform the iteration step to decide $\mathbf{c}_i$ and $\mathbf{d}_i$ after they received the updates of all relay weights. After that, the new receive beamforming vectors are sent back to their user pairs through the relay nodes; however, this will not trigger the weight updating process of the relay nodes, which ensures that the relay nodes only update their weights once at each iteration round.

For user $X_{a,i}$, when it receives the transmit beamforming vector updates from its user pair, namely $\mathbf{c}_i$, as well as all updated weights of the relay nodes, a new round of iteration begins. We assume that the channels are quasi-stationary for t_{max} rounds of iterations. In detail, we denote $f^{(t)}_{m,n,i}$ and $g^{(t)}_{m,n,i}$, where $t \in (1, t_{max})$, as the channel coefficients of the t-th round of iteration, and we assume that $\Delta f_{m,n,i}$=$f^{(t+1)}_{m,n,i} - f^{(t+1)}_{m,n,i}$ and $\Delta g_{m,n,i}$=$g^{(t+1)}_{m,n,i} - g^{(t+1)}_{m,n,i}$ are i.i.d., and bounded by an upper value ξ. After t_{max} rounds of iterations, the channel coefficients are assigned with newly estimated values. The values of t_{max} and ξ together define the level of stationarity of the channel.

Note that during the iteration steps, the instantaneous output power at some relays may exceed their budgets. However, it can be prevented if the individual power constraint is set well below their output power capability. In fact, supported by the simulation results, the required transmit power at each relay node is modest to give a satisfactory performance, especially when the relay number is large.

As can be seen, the computation task assigned for each user node only determines their own beamforming vectors, while in the iSINR method proposed in ([14]), each user node has to compute the beamforming vectors of its own and its user pair's at least. Moreover, for the iSINR method, several iteration steps are required for determining the beamforming vectors before convergence is reached, which is not required in the proposed method. Therefore, the computation complexity of the iSINR method is at least $2t_{conv}$ times that of the proposed method (t_{conv} denotes the iteration steps required to reach/approach convergence).

Summary of Iteration Steps

1) Initialization: $\mathbf{c}_i$=$\mathbf{d}_i$=$[\delta_N \delta_N \cdots \delta_N]^T$, where δ_N=$\sqrt{N}$, w_m=$\sqrt{P_{R,m}/(1+\sigma_r^2)}$ (derived from the expectation of relay output power), and set t=1.
2) Update $\mathbf{a}_i$ and $\mathbf{b}_i$ based on (7) and (8).
3) Update w_m based on (9) and (10).
4) Update $\mathbf{c}_i$ and $\mathbf{d}_i$ based on (16) and (17).
5) Go to step 1) if $t \geq t_{max}$; otherwise, set $t = t + 1$ and go to step 2).

IV. Simulation Results

In this section, simulation results are provided for performance evaluation of the proposed method. For simplicity, we set P_S=1 (compensating for the unconsidered path-loss); all relay nodes have the same output power budget of P_R/M, to ensure the same total relay output power for different relay number settings. P_R/M is determined by SNR_R, which is the ratio of relay output power constraint to the noise variance, i.e., SNR_R=$P_R/(M\sigma_r^2)$.

Figs. 2 and 3 show the average received SINR versus $\mathrm{SNR_R}$ with different number of relay nodes, where a perfect quasi-stationary channel is assumed (ξ=0). In Fig. 2 the iSINR method from [14] is used as a comparison. Moreover, results based on a non-iterative ZF method (denoted by "ZF") used in [14] are also provided. Specifically, in this ZF method, real CSI is considered, $\mathbf{a}_i$ and $\mathbf{b}_i$ are generated as the eigenvectors corresponding to the largest eigenvalues of $\mathbf{F}_i^H\mathbf{F}_i$ and $\mathbf{G}_i^H\mathbf{G}_i$, respectively, and together with $\mathbf{c}_i$ and $\mathbf{d}_i$, the IPI parts are

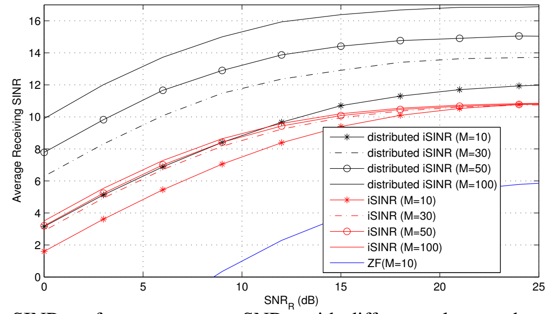

Fig. 2. SINR performance versus $\mathrm{SNR_R}$ with different relay number settings (t_{max}=10, N=5, K=3, ξ=0).

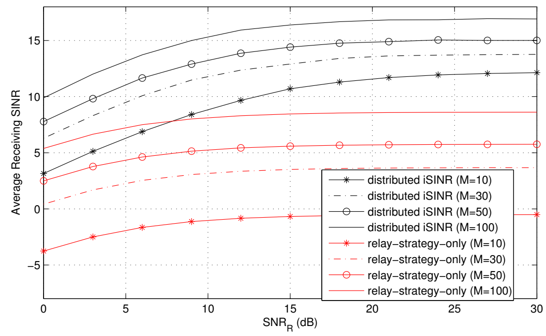

Fig. 3. SINR performance versus $\mathrm{SNR_R}$ with "relay-strategy-only" method as comparison (t_{max}=10, N=5, K=3, ξ=0).

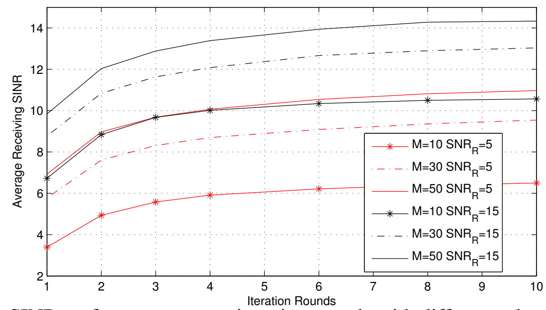

Fig. 4. SINR performance versus iteration rounds with different relay number settings (N=5, K=3, ξ=0.1).

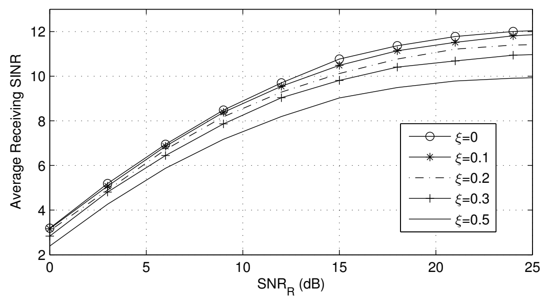

Fig. 5. SINR performance of the proposed method with channels of different stableness level (t_{max}=10, M=10, N=5, K=3).

eliminated completely without any iteration. Both iteration based methods have outperformed the ZF method significantly and the performance of our proposed scheme is the best, at both the low-relay-power and high-relay-power regions. The improvement is more obvious when the relay number is large, and it can also increase the asymptotic SINR by employing more relay nodes in the network, while the original iSINR method can not achieve that.

In Fig. 3, a "relay-strategy-only" method is used as a comparison where the beamforming vectors $\mathbf{a}_i$, $\mathbf{b}_i$, $\mathbf{c}_i$ and $\mathbf{d}_i$ are fixed to their initial values. The figure shows that when only the relay strategy is used in our scheme, the average SINR increases as more relay nodes are employed in the network. However, without the iterative transceiver beamforming steps, the performance is very limited when the relay number is small and the SINR improvement introduced by the transceiver beamforming is significant with any relay number settings.

Fig. 4 illustrates the average SINR of the proposed method after certain rounds of iterations. As can be seen, although the proposed method does not have the best performance immediately after the initialization step, the average SINR will quickly approach its asymptotic value only after a few rounds of iterations. And this pattern applies for different relay number settings and differen total relay power budgets.

Fig. 5 shows the performance for channels with different stationarity levels. By introducing the random channel difference between different iteration rounds, variance of the global channel states will be affected, which will make the comparison unfair. Accordingly, the simulations are performed after compensating the variance changes. The results demonstrate that our proposed scheme will be affected by the channel stationarity level; however, the degradations are within an acceptable range. When the channel states change smoothly (ξ=0.1), the performance degradation is hardly noticeable, compared with the perfect quasi-stationary channel ξ=0.

V. Conclusions

An iterative transceiver beamforming algorithm has been proposed for multipair two-way distributed relay networks, where the iteration steps are distributed among user nodes and relay nodes. As a result, the overall computation complexity can be effectively reduced. On the other hand, a relay strategy is designed for the relay nodes which can significantly increase the SINR performance without the need of extra total relay power, and it only requires simple signal processing operations and local CSI for each relay node. Simulation results indicate that the proposed method is quite robust to channel state changes between different rounds of iterations.

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