

Power Allocation Optimization for Spectrum-Efficient Multi-Pair Two-Way Massive MIMO Full-Duplex Relay over Ricean Channels

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How to cite this paper: Wu, C.Y., Yue, T., Xu, K. and Xie, W. (2017) Power Allocation Optimization for Spectrum-Efficient Multi-Pair Two-Way Massive MIMO Full-Duplex Relay over Ricean Channels. *Int. J. Communications, Network and System Sciences*, 10, 35-47.

<https://doi.org/10.4236/ijcns.2017.108B005>

Received: March 20, 2017

Accepted: August 11, 2017

Published: August 14, 2017

Abstract

In this paper we investigate the power allocation optimization for spectrum efficient multi-pair two-way massive MIMO (TWMM) amplify-and-forward (AF) full-duplex (FD) relay over Ricean fading channels, where multiple user-pairs exchange information within pair through a AF-FD relay with very large number of antennas, while each user equipped with a single antenna. First, the zero-forcing reception/zeroforcing transmission and maximum-ratio combining/maximum ratio transmission processing matrices with imperfect channel state information at the relay are presented. Then, the unified asymptotic signal-to-interference-plus-noise ratio (SINR) expression of the system at general power scaling schemes are investigated. Finally, the joint user-relay power allocation (JURPA) scheme is proposed to improve the spectral efficiency of TWMM-AF-FD relay system. Simulation results show that the proposed JURPA scheme outperforms traditional user-side only power allocation scheme.

Keywords

Massive MIMO, Ricean Fading Channels, Processing Matrices, Power Allocation

1. Introduction

The ever growing challenges for significant traffic growth driven by mobile Internet and Internet of things have made system capacity enhancement one of the most important features in next generation wireless communication systems. The general consensus is that the aggregate data rate will increase by roughly

1000X by 2020. Massive multiple-input multiple output (MIMO) [1] is identified as one of the key enabling technologies to achieve this goal due to its strong potential in boosting the spectral efficiency (SE) of wireless networks [1] [2].

The term massive MIMO indicates that the base station (BS) or relay employs a number of antennas much larger than the number of active data streams per time-frequency resource. Massive MIMO was originally designed for time division duplex (TDD) system [1]-[7], since by exploiting the channel reciprocity in TDD setting, the required channel state information (CSI) for downlink transmission at the BS can be easily obtained via uplink training [1]. The training overhead scales linearly with the number of user equipments (UEs) and is independent with the number of BS antennas. As frequency division duplex (FDD) dominates the current wireless cellular systems, the application of massive MIMO in FDD system is even more desirable. In FDD massive MIMO, the downlink training and corresponding CSI feedback yield an unacceptably high overhead. One attempt of practical FDD massive MIMO is called joint spatial division and multiplexing (JSDM) [8], where the correlation between channels is exploited to reduce the training and feedback dimensions. Another scheme is called beam division multiple access (BDMA) [9], which gets rid of the need of CSI at transmitter and provides strong potential to realize massive MIMO gain in FDD system.

In TDD and FDD massive MIMO systems (namely halfduplex (HD) massive MIMO systems), the uplink and downlink UEs must be allocated with orthogonal time slots or frequency bands, which results in insufficient utilization of time-frequency resources. Inspired by the recent development of full-duplex (FD) communication [10], co-time co-frequency uplink and downlink (CCUD) transmission becomes another option in the cellular system. Although attractive in SE, CCUD transmission is considered challenging due to the strong self interference (SI) caused by the signal leakage between BS/relay transmitter and receiver, especially when the BS is equipped with large-scale antenna arrays. To support the CCUD transmission, the BS employs a separate antenna configuration where two separate large-scale antenna arrays are used for transmission and reception, respectively [11]. In this case, the downlink channel reciprocity is commonly considered as unavailable [12]. Without reciprocity, the training overhead to obtain the downlink CSI scales linearly with the number of BS antennas, which poses another big challenge.

Note that the CCUD transmission in the cellular system with massive MIMO BS/relay has been investigated recently in several works (See [13] [14] [15] and the references therein). The authors in [13] studied the SE performance of CCUD transmission in both macro-cell and small-cell environments. The linear beamforming design of the BS for CCUD transmission has been considered in [14]. The power allocation scheme for user-side only has considered in [15] to optimize the system spectral efficiency.

In this paper, we investigate the power allocation optimization for spectrum efficient multi-pair two-way massive MIMO (TWMM) amplify-and-forward (AF) FD relay over Ricean fading channels. The unified asymptotic signal-to-

interferenceplus-noise ratio (SINR) expression of the system based on the beamforming matrixes of MRC/MRT and ZFR/ZFT at the relay, at the power-scaling ($P_{s,k} = E_{s,k} / M^a, P_{R,k} = E_{R,k} / M^b, 0 \leq a, b \leq 1$ $E_{s,k}$ and $E_{R,k}$ are fixed) are investigates. Moreover, the joint user-relay power allocation scheme is proposed to improve the spectral efficiency of TWMM-AF-FD relay system.

Notation: X^T ; X^H ; X^* ; X^{-1} ; $Tr(X)$ to denote the transpose, conjugate-transpose, conjugate, inverse and the trace of X . respectively. I_M denotes an $M \times M$ identity matrix. $\mathbb{E}\{\cdot\}$ is the expectation operator, $\|\cdot\|$ represents the Euclidean norm.

2. System Model

We consider the $K(K \geq 2)$ user-pairs two-way AF relay system, where K pairs of single-antenna users (U_{2l-1}, U_{2l})($l=1, \dots, K$) on two sides try to exchange information within pair through a massive antenna relay (R) with M antennas, as is illustrated in **Figure 1**. Without loss of generality, a pair of source nodes U_{2l-1} and U_{2l} are too far apart to communicate directly. Meanwhile, one source can be inevitably interfered by others on the same side.

2.1. Channel Model

Let $G \in C^{M \times 2K}$ and $F^T \in C^{2K \times M}$ denote the channel from all users transmit antennas to relay R receive antenna array and the channel from the relay R transmit antennas to all users receive antennas, respectively. Specifically, G and F can be expressed as $G = H_1 D_1^{1/2}$ and $F = H_2 D_2^{1/2}$, $H_i \in C^{M \times 2K}$ ($i=1, 2$) is the channel matrix representing fast fading and $D_i \in C^{2K \times 2K}$ ($i=1, 2$) is the diagonal matrix representing large-scale fading with $[D_1]_{KK} = \beta_{g,k}$, $[D_2]_{KK} = \beta_{f,k}$. The fast fading matrix can be written as [4], $H_i = \overline{H}_i [\Omega(\Omega + I_{2K})^{-1}]^{-1/2} + H_{i,w} [\Omega(\Omega + I_{2K})^{-1}]^{-1/2}$, where Ω is a $2K \times 2K$ Ricean K -factor diagonal matrix with $\Omega_{ii} = K_i$. $H_{i,w}$ contains the independent identically distributed (i.i.d.) $CN(0,1)$ entries. \overline{H}_i denotes the deterministic component, and let $\overline{H}_i = e^{-j(m-1)(2\pi/\lambda)\sin\theta_k}$, where θ_k denotes the arrival angle

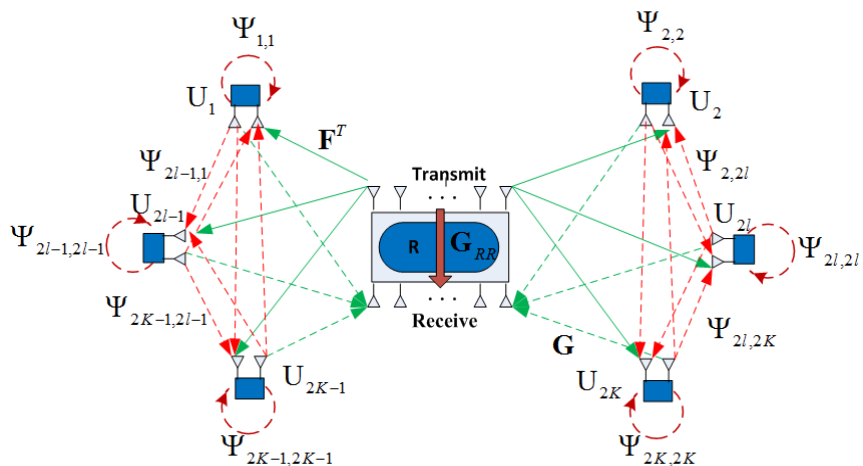


Figure 1. Illustration of the TWMM-AF-FD relay system.

of the k -th user, λ is the wavelength, and d is the antenna spacing. For convenience, we set $d = \lambda / 2$.

Let $G_{RR} \in C^{M \times M}$ denote the echo interference (EI) channel matrix between the relay transmit and receive arrays with (i.i.d.) $CN(0, \sigma_{rr}^2)$ elements. $\Psi_{k,k}$ and $\Psi_{k,i}$ represent the self-loop interference coefficient at U_k and the inter-user interference channel coefficient from U_i to U_k . $S_k = \{1, 3, \dots, 2K - 1\}$ or $\{2, 4, \dots, 2K\}$ denotes the set of users on the same side. $\Psi_{k,k}$ and $\Psi_{k,i}$ can be modeled as i.i.d $CN(0, \Phi_{k,k})$ and $CN(0, \Phi_{k,i})$ random variables.

2.2. Channel Estimation and Data Transmission

Since it is impossible for the relay to obtain the complete channel state information from all the channels, so it is necessary to estimate the channel matrix. For the Ricean fading channel model, the Ricean K factor matrix and the LOS transmission signal component are fully known in the relay and the user. So we only need to estimate $G_w = H_{1,w} D_1^{1/2}$ and $F_w = H_{2,w} D_2^{1/2}$, since G_w and F_w are i.i.d, we can use minimum mean square error (MMSE) estimator. Let T be the length of the coherence interval and let be τ the number of symbols used for uplink pilots. In the training part of the coherent interval, all users receive and transmit antennas simultaneously send symbols of length τ to the relay.

The received pilots matrices at the R's receive and transmit antenna arrays are given by $Y_r = \sqrt{\tau P_T} G \Phi_1 + \sqrt{\tau P_T} \hat{F} \Phi_2 + N_r$ and $Y_t = \sqrt{\tau P_T} \hat{G} \Phi_1 + \sqrt{\tau P_T} F \Phi_2 + N_t$, among them $\Phi_i \in C^{2K \times \tau}$ ($i=1, 2$). are pilot sequences transmitted from all users transmit antennas and all users receive antennas, respectively. $\tilde{F} \in C^{M \times 2K}$ is the channel matrices from all users transmit antennas to R's transmit antenna array and $\tilde{F} \in C^{M \times 2K}$ is from all users receive antennas to R's receive antenna array. N_r and N_t are additive white Gaussian noise (AWGN) matrices with (i.i.d.) $CN(0, \sigma)$ elements. And P_T is the transmit power of each pilot symbol. All pilot sequences are assumed to be paired and distributed independently, $\overline{\Phi_i \Phi_i^H} = I_{2K}$ and $\overline{\Phi_i \Phi_j^H} = O_{2K}$ ($i \neq j$) where $\overline{\Phi_i} \triangleq [(\Omega + I_{2K})^{-1}]^{1/2} \Phi_i$. This requires $\tau = 4K$. and we set $\tau = 4K$ in this paper. We consider the LOS component assumed to be known and can be removed, the remaining terms of the received matrices are $Y_{r,w} = \sqrt{\tau P_T} G_w \overline{\Phi_1} + \sqrt{\tau P_T} \tilde{F}_w \overline{\Phi_2} + N_r$, and $Y_{t,w} = \sqrt{\tau P_T} \tilde{G}_w \overline{\Phi_1} + \sqrt{\tau P_T} F_w \overline{\Phi_2} + N_t$. The MMSE estimate of G_w and F_w are $\hat{G}_w = (1/\sqrt{\tau P_T}) Y_{r,w} \overline{\Phi_1}^H (I_{2K} + (1/\sqrt{\tau P_T}) D_1^{-1})^{-1}$ and $\hat{F}_w = (1/\sqrt{\tau P_T}) Y_{t,w} \overline{\Phi_2}^H (I_{2K} + (1/\sqrt{\tau P_T}) D_2^{-1})^{-1}$.

According to MMSE estimates, the actual channel can be expressed as $G = \hat{G} + \Delta G$ and $F = \hat{F} + \Delta F$, \hat{G} (\hat{F}) and ΔG (ΔF) denote the available channel estimate and estimation error, respectively. The elements of the i -th column of ΔG and ΔF are RVs with zero means and variances $\varepsilon_{g,i}^2 = \beta_{g,i} \sigma / (\sigma + P_p \beta_{g,i}) (K_i + 1)$ and $\varepsilon_{f,i}^2 = \beta_{f,i} \sigma / (\sigma + P_p \beta_{f,i}) (K_i + 1)$ where $P_p = \tau P_T$. In addition, due to the nature of MMSE estimates, ΔG is independent of \hat{G} and ΔF is independent of \hat{F} .

At time instant t , all sources transmit their symbols to R and R forwards the

amplified signal to destinations. The received signals at the relay and the k 'th user are given by:

$$y_R(t) = Gx(t) + G_{RR}x_R(t) + n_R(t). \tag{1}$$

$$y_k(t) = f_k^T x_R(t) + \sum_{i \in S_k} \psi_{k,i} x_i(t) + n_k(t). \tag{2}$$

where $x(t) = [x_1(t), x_2(t), \dots, x_{2K}(t)]^T$ and $\mathbb{E}\{x(t)^H x(t)\} = \text{diag}(P_{S,1}, \dots, P_{S,2K}) = \Lambda$, $x_R(t)$ denotes the transmit vector of R . $n_R(t)$ and $n_K(t)$ represent the noise vector at R and U_k . the elements of $n_R(t)$ and $n_K(t)$ are assumed to be (i.i.d.) $CN(0, \sigma_n)$ and $CN(0, \sigma_n)$.

The transmit vector of R at time instant t can be expressed as:

$$x_R(t) = Wy_R(t-d) \tag{3}$$

where $W \in C^{M \times M}$ is the beamforming matrix, and d denotes the processing delay at R . In this paper, some loop interference cancellation methods [16] can be adopted at the relay before carrying out (3), so we can regard the residual loop interference at R as additional noise. So, we replace $x_R(t)$ in the loop interference term in (1) by a Gaussian noise source $\tilde{x}_R(t)$ with the same power limitation to represent the residual loop interference signal. Then (2) can be expressed as:

$$y_k(t) = (\hat{f}_k^T + \Delta f_k^T)W[(\hat{G} + \Delta \hat{G})x(t-d) + G_{RR}\tilde{x}_R(t-d) + n_R(t-d)] + \sum_{i \in S_k} \psi_{k,i} x_i(t) + n_k(t) \tag{4}$$

3. Beamforming Design and SINR Analysis

In this section, we introduce the ZFR/ZFT-based and MRC/MRT-based beamforming design, the end-to-end SINR of the proposed transceiver schemes for multi-pair two-way MM-AFFDR system are analyzed.

Lemma 1: By the law of large numbers, if M is large enough, the inner product of any two columns in the estimate channel matrix \hat{G} can be expressed as [17]:

$$\frac{\hat{g}_n^H \hat{g}_i}{M} \xrightarrow[M \rightarrow \infty]{a.s.} \begin{cases} \frac{\beta_{g,n}}{K_n + 1} (K_n + \frac{P_p \beta_{g,n}}{\sigma + P_p \beta_{g,n}}) = \eta_{g,n}, & i = n \\ 0, & i \neq n \end{cases} \tag{5}$$

and

$$\frac{\hat{f}_n^H \hat{f}_i}{M} \xrightarrow[M \rightarrow \infty]{a.s.} \begin{cases} \frac{\beta_{f,n}}{K_n + 1} (K_n + \frac{P_p \beta_{f,n}}{\sigma + P_p \beta_{f,n}}) = \eta_{f,n}, & i = n \\ 0, & i \neq n \end{cases} \tag{6}$$

According to *Lemma 1*, it can be easily obtained that:

$$\frac{\hat{G}^H \hat{G}}{M} \xrightarrow[M \rightarrow \infty]{a.s.} \text{diag}\{\eta_{g,1}, \eta_{g,2}, \dots, \eta_{g,2k}\} = Q_1 \tag{7}$$

and

$$\frac{\hat{F}^H \hat{F}}{M} \xrightarrow[M \rightarrow \infty]{a.s.} \text{diag}\{\eta_{f,1}, \eta_{f,2}, \dots, \eta_{f,2k}\} = Q_2 \tag{8}$$

3.1. ZFR/ZFT Beamforming

The ZFR/ZFT beamforming matrix can be expressed as [18]:

$$W_{zf} = a_{zf} \widehat{F}^* (\widehat{F}^T \widehat{F}^*)^{-1} P (\widehat{G}^H \widehat{G})^{-1} \widehat{G}^H \tag{9}$$

where $P = \text{diag}(P_1, \dots, P_K)$, $P_l = \begin{pmatrix} 0 & 1 \\ 1 & 0 \end{pmatrix}$, $l = 1, \dots, K$. a_{zf} is the amplification factor, which can be expressed as:

$$a_{zf} \xrightarrow[M \rightarrow \infty]{a.s.} \sqrt{\frac{P_{R,k}}{\frac{1}{M} \sum_{i=1}^{2k} P_{s,i} \eta_{f,i}^{-1} \eta_{f,i} + \frac{1}{M^2} (\sigma + \sigma_{rr}^2 P_{R,k} + \sum_{j=1}^{2k} P_{s,j} \varepsilon_{g,j}^2) \sum_{i=1}^{2k} \eta_{f,i}^{-1} \eta_{g,i}^{-1}}} \tag{10}$$

where (i, i') is a pair of user.

Theorem 1: Using ZFR/ZFT beamforming matrix with imperfect CSI from MMSE estimation, the end-to-end SINR at the k 'th user can be expressed asymptotically (in M) as:

$$\gamma_k = \frac{P_{s,k} a_{zf}^2}{\mathbb{E} \left[|y_k^{ee}|^2 \right] + \mathbb{E} \left[|y_k^{ei}|^2 \right] + \frac{a_{zf}^2}{M} \sigma \eta_{g,k}^{-1} + \sum_{i=S_k} P_{s,i} \Phi_{k,i} + \sigma_n} \tag{11}$$

the power for channel estimation error $\mathbb{E} \left[|y_k^{ee}|^2 \right]$ and power for echo interference can $\mathbb{E} \left[|y_k^{ei}|^2 \right]$ be expressed asymptotically (in M) as:

$$\begin{aligned} \mathbb{E} \left[|y_k^{ee}|^2 \right] &\xrightarrow[M \rightarrow \infty]{a.s.} \frac{a_{zf}^2 \eta_{g,k}^{-1}}{M} \sum_{i=1}^{2k} P_{s,i} \varepsilon_{g,i}^2 + \frac{a_{zf}^2 \varepsilon_{f,k}^2}{M} \sum_{i=1}^{2k} P_{s,i} \eta_{f,i}^{-1} \\ &+ \frac{a_{zf}^2 \varepsilon_{f,k}^2}{M^2} \sum_{i=1}^{2k} P_{s,j} \varepsilon_{g,j}^2 \sum_{i=1}^{2k} \eta_{f,i}^{-1} \eta_{g,i}^{-1} + \frac{a_{zf}^2 \sigma}{M^2} \varepsilon_{f,k}^2 \sum_{i=1}^{2k} \eta_{f,i}^{-1} \eta_{g,i}^{-1} \\ \mathbb{E} \left[|y_k^{ei}|^2 \right] &\xrightarrow[M \rightarrow \infty]{a.s.} \frac{P_{R,k} a_{zf}^2}{M^2} \varepsilon_{f,k}^2 \sigma_{rr}^2 \sum_{i=1}^{2k} \eta_{f,i}^{-1} \eta_{g,i}^{-1} + \frac{P_{R,k} a_{zf}^2}{M} \sigma_{rr}^2 \eta_{g,k}^{-1} \end{aligned} \tag{12}$$

Using ZFR/ZFT beamforming matrix with imperfect CSI from MMSE estimation, at the power ($P_{s,k} = E_{s,k} / M^a$, $P_{R,k} = E_{R,k} / M^a$, $0 < a, b < 1$), the asymptotic SINR when $M \rightarrow \infty$ can be expressed as:

$$\gamma_k \xrightarrow[M \rightarrow \infty]{a.s.} \frac{E_{s,k'}}{(E_{R,k} M^{-b} \sigma_{rr}^2 + \sigma) M^{a-1} \eta_{g,k'}^{-1}} + (M^{-a} \sum_{i=S_k} E_{s,i} \Phi_{k,i} + \sigma_n) \frac{M^{b-1} \sum_{i=1}^{2K} E_{s,i} \eta_{f,i}^{-1}}{E_{R,k}} + \frac{M^{a+b-1}}{E_{R,k}} \sigma_n \sigma \sum_{i=1}^{2K} \eta_{f,i}^{-1} \eta_{g,i}^{-1} \tag{13}$$

3.2. MRC/MRT Beamforming

The MRC/MRT beamforming matrix can be expressed as [18]:

$$W_{mrc} = a_{mrc} \widehat{F}^* P \widehat{G}^H \tag{14}$$

where $P = \text{diag}(P_1, \dots, P_K)$, $P_l = \begin{pmatrix} 0 & 1 \\ 1 & 0 \end{pmatrix}$, $l = 1, \dots, K$. a_{mrc} is the amplification factor, which can be expressed as:

$$a_{mrc} \xrightarrow[M \rightarrow \infty]{a.s.} \sqrt{\frac{P_{R,k}}{M^3 \sum_{i=1}^{2K} P_{s,i} \eta_{f,i} \eta_{f,i}^2 + M^2 (\sigma + \sigma_{rr}^2 P_{R,k} + \sum_{j=1}^{2k} P_{s,j} \varepsilon_{g,j}^2) \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i}}} \tag{15}$$

Theorem 2: Using MRC/MRT beamforming matrix with imperfect CSI from MMSE estimation, the end-to-end SINR at the k 'th user can be expressed asymptotically (in M) as:

$$\gamma_k = \frac{P_{s,k} a_{mrc}^2 \eta_{f,k}^2 \eta_{g,k}^2 M^4}{\mathbb{E}[|y_k^{ee}|^2] + \mathbb{E}[|y_k^{ei}|^2] + a_{mrc}^2 M^3 \sigma \eta_{f,k}^2 \eta_{g,k} + \sum_{i \in S_k} P_{s,i} \Phi_{k,i} + \sigma_n} \quad (16)$$

the power for channel estimation error $\mathbb{E}[|y_k^{ee}|^2]$ and power for echo interference can $\mathbb{E}[|y_k^{ei}|^2]$ be expressed asymptotically (in M) as:

$$\begin{aligned} \mathbb{E}[|y_k^{ee}|^2] &\xrightarrow[M \rightarrow \infty]{a.s.} M^3 a_{mrc}^2 \eta_{f,k}^2 \eta_{g,k} \sum_{i=1}^{2k} P_{s,i} \varepsilon_{g,i}^2 + M^2 a_{mrc}^2 \varepsilon_{f,k}^2 \sum_{j=1}^{2k} P_{s,j} \varepsilon_{g,j}^2 \sum_{i=1}^{2k} \eta_{f,i} \eta_{g,i} \\ &+ M^3 a_{mrc}^2 \varepsilon_{f,k}^2 \sum_{i=1}^{2k} \eta_{f,i} \eta_{g,i}^2 P_{s,i} + M^2 a_{mrc}^2 \sigma \varepsilon_{f,k}^2 \sum_{i=1}^{2k} \eta_{f,i} \eta_{g,i} \\ \mathbb{E}[|y_k^{ei}|^2] &\xrightarrow[M \rightarrow \infty]{a.s.} P_{R,k} M^3 a_{mrc}^2 \sigma_{rr}^2 \eta_{f,k}^2 \eta_{g,k} + P_{R,k} M^2 a_{mrc}^2 \varepsilon_{f,k}^2 \sigma_{rr}^2 \sum_{i=1}^{2k} \eta_{f,i} \eta_{g,i} \end{aligned} \quad (17)$$

Using MRC/MRT beamforming matrix with imperfect CSI from MMSE estimation, at the power ($P_{s,k} = E_{s,k} / M^a, P_{R,k} = E_{R,k} / M^a, 0 < a, b < 1$), the asymptotic SINR when $M \rightarrow \infty$ can be expressed as:

$$\gamma_k \xrightarrow[M \rightarrow \infty]{a.s.} \frac{E_{s,k} \eta_{f,k}^2 \eta_{g,k}^2}{(E_{R,k} M^{-b} \sigma_{rr}^2 + \sigma) M^{a-1} \eta_{f,k}^2 \eta_{g,k} + (M^{-a} \sum_{i \in S_k} E_{s,i} \Phi_{k,i} + \sigma_n) \frac{M^{b-1}}{E_{R,k}} \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i}^2 E_{s,i} + \frac{M^{a+b-2}}{E_{R,k}} \sigma_n \sigma \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i}} \quad (18)$$

Specially, when $\eta_{f,i} = \eta_{g,i} = \eta$, the deterministic equivalent for γ_k of ZFR/ZFT becomes the same as that of MRC/MRT.

4. Spectral Efficiency Optimization by Power Allocation

In the previous sections, we have assumed that all the sources and relay use the same transmit power. However, the spectral efficiency achieved in reality may not be optimal since each source-destination pair suffers from different fading environments. As a result, power control scheme at sources and relay is need to improve the system performance and optimize the SE subject to the maximum power constrains $P_{s,k}^{\max}$ at U_k and $P_{R,k}^{\max}$ at R . The SE is defined as:

$$SE = \frac{T - \tau}{2T} \mathbb{E}[\sum_{i=1}^{2K} \log_2(1 + \gamma_i)] \quad (19)$$

For convenience to analysis, we rewrite the end-to-end SINR in *Theorem 1, 2* under ZFR/ZFT and MRC/MRT beamforming schemes as a unified expression as:

$$\gamma_k = \frac{K_i P_{s,k}}{\sum_{i=1}^{2K} A_{k,i} P_{s,i} + \sum_{i \in S_k} a_{k,i} P_{s,i} + B_k P_{R,k} + \sum_{i=1}^{2K} c_i \frac{P_{s,i}}{P_{R,i}} + \sum_{i \in S_k} c_{k,i} \frac{P_{s,i}}{P_{R,i}} + \sum_{j \in S_k} \phi_{k,j} P_{s,j} \sum_{i=1}^{2K} d_i \frac{P_{s,i}}{P_{R,i}} + e_k P_{R,k}^{-1} + f_k} \quad (20)$$

For ZFR/ZFT beamforming scheme, we have:

$$A_{k,i} = \frac{\eta_{g,k}^{-1} \varepsilon_{g,i}^2}{M} + \frac{\eta_{f,k}^2 \eta_{f,i}^{-1}}{M} + \frac{\eta_{f,k}^2 \varepsilon_{g,i}^2}{M^2} \sum_{j=1}^{2K} \eta_{f,j}^{-1} \eta_{g,j}^{-1}$$

$$\begin{aligned}
 a_{k,i} &= \frac{\Phi_{k,i} \sigma_{rr}^2 \sum_{j=1}^{2K} \eta_{f,j}^{-1} \eta_{g,j}^{-1}}{M^2} \\
 B_k &= \frac{\varepsilon_{f,k}^2 \sigma_{rr}^2 \sum_{i=1}^{2K} \eta_{f,i}^{-1} \eta_{g,i}^{-1}}{M^2} + \frac{\sigma_{rr}^2 \eta_{g,k}^{-1}}{M} \\
 c_i &= \frac{\sigma_n \eta_{f,i}^{-1}}{M} + \frac{\varepsilon_{g,i}^2 \sigma_n \sum_{j=1}^{2K} \eta_{f,j}^{-1} \eta_{g,j}^{-1}}{M^2} \\
 c_{k,i} &= \frac{\sigma_n \eta_{f,i}^{-1}}{M} + \frac{\varepsilon_{g,i}^2 \sigma_n \sum_{j=1}^{2K} \eta_{f,j}^{-1} \eta_{g,j}^{-1}}{M^2} \\
 d_i &= \frac{\eta_{f,i}^{-1}}{M} + \frac{\varepsilon_{g,i}^2}{M^2} \sum_{ii=1}^{2K} \eta_{f,ii}^{-1} \eta_{g,ii}^{-1} \\
 e_k &= \frac{\sigma_n \sigma \sum_{i=1}^{2K} \eta_{f,i}^{-1} \eta_{g,i}^{-1}}{M^2} \\
 f_k &= \frac{\sigma}{M^2} \varepsilon_{f,k}^2 \sum_{i=1}^{2K} \eta_{f,i}^{-1} \eta_{g,i}^{-1} + \frac{\sigma_{rr}^2 \sigma_n \sum_{i=1}^{2K} \eta_{f,i}^{-1} \eta_{g,i}^{-1}}{M^2} + \frac{\sigma \eta_{g,k}^{-1}}{M}
 \end{aligned} \tag{21}$$

For MRC/MRT beamforming scheme, we have:

$$\begin{aligned}
 A_{k,i} &= \frac{\eta_{g,k}^{-1} \varepsilon_{g,i}^2}{M} + \frac{\eta_{f,i} \varepsilon_{f,k}^2 \varepsilon_{g,i}^2}{\eta_{f,k}^2 \eta_{g,k}^2 M} + \frac{\varepsilon_{f,k}^2 \varepsilon_{g,i}^2 \sum_{j=1}^{2K} \eta_{f,j}^{-1} \eta_{g,j}^{-1}}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} \\
 a_{k,i} &= \frac{\Phi_{k,i} \sigma_{rr}^2 \sum_{j=1}^{2K} \eta_{f,j} \eta_{g,j}}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} \\
 B_k &= \frac{\varepsilon_{f,k}^2 \sigma_{rr}^2 \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i}}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} + \frac{\sigma_{rr}^2 \eta_{g,k}^{-1}}{M} \\
 c_i &= \frac{\sigma_n \eta_{f,i} \eta_{g,i}^2}{\eta_{f,k}^2 \eta_{g,k}^2 M} + \frac{\varepsilon_{g,i}^2 \sigma_n \sum_{j=1}^{2K} \eta_{f,j}^{-1} \eta_{g,j}^{-1}}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} \\
 c_{k,i} &= \frac{\Phi_{k,i} \sigma \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i}}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} \\
 d_i &= \frac{\eta_{f,i} \eta_{g,i}^2}{\eta_{f,k}^2 \eta_{g,k}^2 M} + \frac{\varepsilon_{g,i}^2}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} \sum_{ii=1}^{2K} \eta_{f,ii} \eta_{g,ii} \\
 e_k &= \frac{\sigma_n \sigma \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i}}{\eta_{f,k}^2 \eta_{g,k}^2 M^2}
 \end{aligned}$$

$$f_k = \frac{\sigma}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} \varepsilon_{f,k}^2 \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i} + \frac{\sigma_{rr}^2 \sigma_n \sum_{i=1}^{2K} \eta_{f,i} \eta_{g,i}}{\eta_{f,k}^2 \eta_{g,k}^2 M^2} + \frac{\sigma \eta_{g,k}^{-1}}{M} \quad (22)$$

The SE optimization problem can be written as (23). By introducing auxiliary variables $v_k \leq 1 + \gamma_k$, (23) can be written as (24).

$$\begin{aligned} \max_{P_{s,1}, \dots, P_{s,2K}; P_{R,1}, \dots, P_{R,2K}} \prod_{k=1}^{2K} (1 + \gamma_k) &= \min_{P_{s,1}, \dots, P_{s,2K}; P_{R,1}, \dots, P_{R,2K}} \prod_{k=1}^{2K} \frac{1}{1 + \gamma_k} \\ \text{s.t.} \quad & \left\{ \begin{aligned} C_1 : & \frac{K_i P_{s,k}}{\sum_{i=1}^{2K} A_{k,i} P_{s,i} + \sum_{i \in S_k} a_{k,i} P_{s,i} + B_k P_{R,k} + \sum_{i=1}^{2K} c_i \frac{P_{s,i}}{P_{R,i}} + \sum_{i \in S_k} c_{k,i} \frac{P_{s,i}}{P_{R,i}} + \sum_{i \in S_k} \phi_{k,j} P_{s,j} \sum_{i=1}^{2K} d_i \frac{P_{s,i}}{P_{R,i}} + e_k P_{R,k}^{-1} + f_k} \geq \gamma_k \\ C_2 : & 0 \leq P_{s,k} \leq P_{s,k}^{\max}, k = 1, 2, \dots, 2K \\ C_3 : & 0 \leq P_{R,k} \leq P_{R,k}^{\max}, k = 1, 2, \dots, 2K \end{aligned} \right. \quad (23) \end{aligned}$$

$$\begin{aligned} \min_{P_{s,1}, \dots, P_{s,2K}; P_{R,1}, \dots, P_{R,2K}} \prod_{k=1}^{2K} \frac{1}{v_k} \\ \text{s.t.} \quad & \left\{ \begin{aligned} C_1 : & v_k \leq 1 + \gamma_k, k = 1, 2, \dots, 2K \\ C_2 : & \sum_{i=1}^{2K} A_{k,i} P_{s,i} P_{s,k}^{-1} \gamma_k + \sum_{i \in S_k} a_{k,i} P_{s,i} P_{s,k}^{-1} \gamma_k + B_k P_{R,k} P_{s,k}^{-1} \gamma_k + \sum_{i=1}^{2K} c_i \frac{P_{s,i} P_{s,k}^{-1} \gamma_k}{P_{R,i}} + \sum_{i \in S_k} c_{k,i} \frac{P_{s,i} P_{s,k}^{-1} \gamma_k}{P_{R,k}} \\ & + \sum_{i \in S_k} \phi_{k,j} P_{s,j} \sum_{i=1}^{2K} d_i \frac{P_{s,i} P_{s,k}^{-1} \gamma_k}{P_{R,i}} + e_k P_{R,k}^{-1} P_{s,k}^{-1} \gamma_k + f_k P_{s,k}^{-1} \gamma_k \leq 1 \\ C_3 : & 0 \leq P_{s,k} \leq P_{s,k}^{\max}, k = 1, 2, \dots, 2K \\ C_4 : & 0 \leq P_{R,k} \leq P_{R,k}^{\max}, k = 1, 2, \dots, 2K \end{aligned} \right. \quad (24) \end{aligned}$$

Since the target function is not in the posynomial form, the similar method in [19] for solving geometric programming (GP) can be used. For any $\gamma_k > 0$, $1 + \gamma_k$ can be approximated by a posynomial function $\theta_k(\gamma_k)^{\omega_k}$ near $\hat{\gamma}_k$, where $\omega_k = \hat{\gamma}_k / (1 + \hat{\gamma}_k)$, $\theta_k = \hat{\gamma}_k^{-\omega_k} (1 + \hat{\gamma}_k)$. Then the SE optimization problem can be solved by using several GPs, Algorithm 1 is in the following description.

Algorithm 1 The PC Strategy by GP:

Input: The initial value $\hat{\gamma}_k^{(0)}$, ε , $P_{s,k}^{\max}$, $P_{R,k}^{\max}$, $i = 1$.

Export: The optimal $\hat{\gamma}_k$, and its corresponding $P_{s,k}$ and $P_{R,k}$.

while $< i$ Limited Iteration Number **do**

• solve the GP problem:

$$\min (P_{s,1}, \dots, P_{s,2K}; P_{R,1}, \dots, P_{R,2K}) \prod_{k=1}^{2K} \theta_k^{-1} \gamma_k^{-w_k} \text{ s.t. } C_1, C_2, C_3, C_4 \quad \text{where}$$

$$w_k = \hat{\gamma}_k^{(i-1)} / (1 + \hat{\gamma}_k^{(i-1)}), \theta_k = (1 + \hat{\gamma}_k^{(i-1)}) / (\hat{\gamma}_k^{(i-1)})^{w_k}.$$

if $\max(k = 1, \dots, 2K) |\hat{\gamma}_k^i - \hat{\gamma}_k^{(i-1)}| < \varepsilon$ **then**

break

else

$$i = i + 1$$

end if

end while

5. Simulations Results

In this section, we examine the SE of the multipair two-way MM-AF-FDR system when its uplink power and downlink power are distribution at the same time. Without loss of generality, we let $\sigma = \Phi_{k,i} = \sigma_n^2 = 1$, and all users have the same Ricean K-factor. Under such assignments, the asymptotic SE of MRC/MRT and ZFR/ZFT are equal. Then we let $\tau = 4K$, $P_r = -3.8dB$, $T = 196$.

Figure 2 shows the SE v.s. ν . We can see from **Figure 2** that the achievable SE is improved significantly by using the proposed JURPA scheme, which optimize the uplink power and downlink power at the same time. Compared with the user-side only power allocation scheme in [15], the proposed JURPA scheme can obtain significantly SE improvement, and the SE gain increases with the decreasing of ν .

Figure 3 shows the SE v.s. the number of the relay antennas. The proposed JURPA scheme obtains the optimum SE performance when compared with the user-side only power allocation scheme in [15] and the no power allocation scheme. The SE performance of TWMM-AF-FD system with ZFR/ZFT beamforming outperforms MRC/MRT beamforming.

Figure 4 shows the SE performance v.s. σ_{rr}^2 . It is seen from **Figure 4** that the proposed JURPA scheme outperforms the userside only power allocation scheme in [15] and the no power allocation scheme on the system SE performance. Moreover, the SE gain increases with the increasing of σ_{rr}^2 .

6. Conclusion

In this paper we investigate the power allocation optimization for spectrum efficient TWMM-AF-FD relay over Ricean fading channels. First, the ZFR/ZFT and

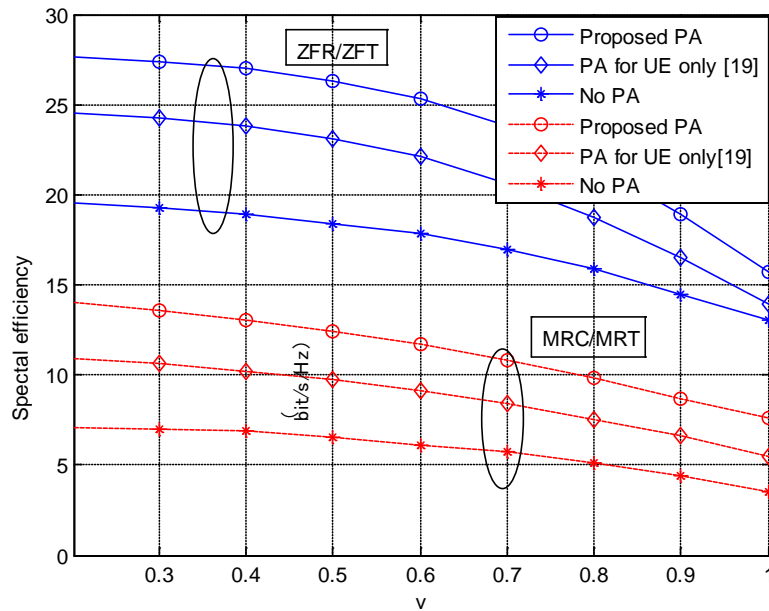


Figure 2. The SE v.s. ν , where $K = 5$, $E_{s,k} = 10dB$, Ricean K - factor = 4, $P_{R,k}^{max} = 2KP_{s,k}^{max}$, $P_{s,k}^{max} = E_{s,k} / M^\nu$, $M = 200$, $\sigma_{rr}^2 = 5$.

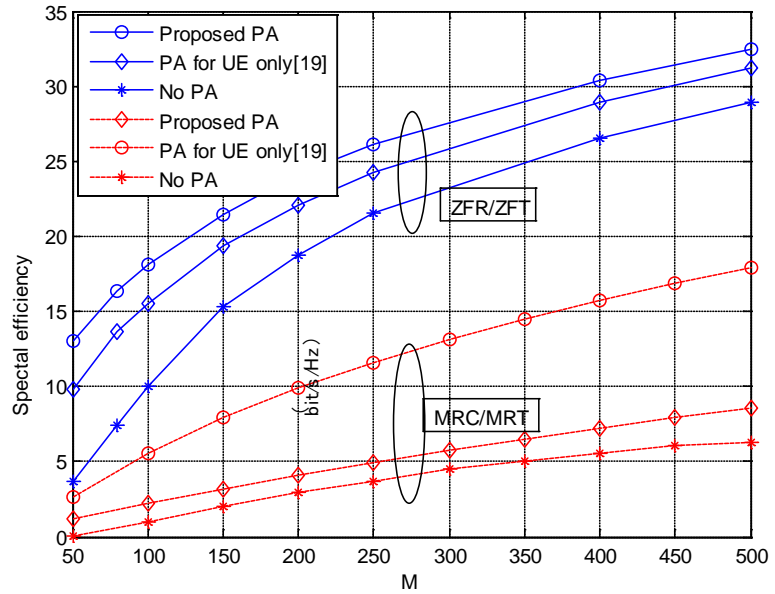


Figure 3. The SE v.s. the number of the relay antennas M , where $K = 5$, $E_{s,k} = 10\text{dB}$, Ricean K -factor = 4, $P_{R,k}^{\max} = 2KP_{s,k}^{\max}$, $P_{s,k}^{\max} = E_{s,k} / M^v$, $v = 0.6$, $\sigma_{rr}^2 = 5$.

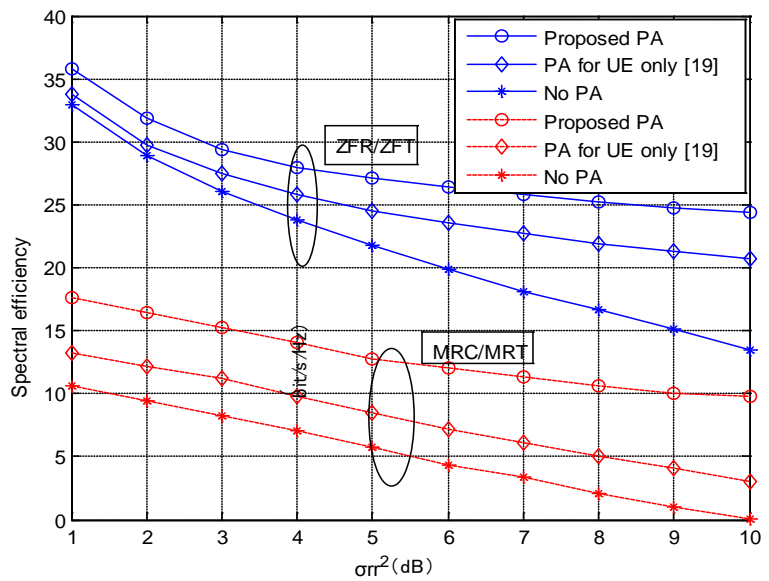


Figure 4. The SE v.s. σ_{rr}^2 , where $K = 5$, $E_{s,k} = 10\text{dB}$, Ricean K -factor = 4, $P_{R,k}^{\max} = 2KP_{s,k}^{\max}$, $P_{s,k}^{\max} = E_{s,k} / M^v$, $v = 0.2$, $M = 200$.

MRC/MRT processing matrices with imperfect channel state information at the relay are presented. Then, the unified asymptotic SINR expression of the system at general power scaling schemes are investigated. Finally, the JURPA scheme is proposed to improve the spectral efficiency of TWMM-AF-FD relay system. Simulation results show that the proposed JURPA scheme outperforms traditional user-side only power allocation scheme.

Acknowledgements

This work is supported by Jiangsu Province Natural Science Foundation under Grant BK20160079, National Natural Science Foundation of China (No. 61671472, 91438115, No. 61371123).

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