Joint Transceiver Designs for Full-Duplex K-Pair MIMO Interference Channel With SWIPT

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***Abstract*— In this paper, we propose joint transceiver design algorithms for the full-duplex K-pair multiple-input multiple-output interference channel with simultaneous wireless information and power transfer. To mitigate and exploit the complex interference, we consider two important utility optimization problems, i.e., the sum power minimization problem and the sum-rate maximization problem. In the first problem, our aim is to minimize the total transmission power under both transmission rate and energy harvesting (EH) constraints. An iterative algorithm based on alternating optimization (AO) and with guaranteed monotonic convergence is proposed to successively optimize the transceiver coefficients. The algorithm consists of three main steps, where the concave–convex procedure (CCCP), the minimum mean-square error (MMSE) criterion, and the semidefinite relaxation technique are, respectively, employed to compute the vectors of power splitting ratios, the receiving matrices, and the transmitting beamforming vectors. Two simplified algorithms based on fixed beamformers, namely, the maximum ratio transmission and the maximum signal-to-interference-leakage beamformers are also proposed. In the second problem, our aim is to maximize the sum-rate under additional power and EH constraints. Due to the highly non-convex nature of this problem, we first reformulate it into an equivalent-weighted MMSE problem by introducing suitable weight factors, such that the global optima of the two problems are identical. Then, by utilizing the concept of AO and CCCP, we show that the equivalent problem can be efficiently solved. Again, with the aid of the fixed beamformers, two simplified algorithms are provided to reduce the computational complexity. Simulation results are presented to validate the effectiveness of the proposed algorithms.**

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***Index Terms*— Transceiver design, full-duplex, energy harvesting, MIMO interference channel, SWIPT.**

I. INTRODUCTION

TO SUPPORT the exponential increasing data demand in wireless communication with the proliferation of wireless services, more powerful communication technologies that exploit the current spectrum more efficiently is needed. Also, the proliferation of wireless communication technologies, e.g., wireless sensor networks (WSNs), calls for higher requirement for the quality of power supply to wireless devices. To meet these new requirements, full-duplex (FD) transmission and energy harvesting (EH) are two new technological solutions that are currently under considerable investigation.

A. Background and Prior Work

The FD transmission is a powerful and promising technique for the next-generation wireless communication systems which can exploit the current spectrum more efficiently by enabling transmission and reception of user signals at the same time and, over the same frequency band [1]–[19], [21]. In traditional half-duplex (HD) communication systems, devices can either transmit or receive on a single frequency band, but not simultaneously. In FD systems, devices can transmit and receive *simultaneously* on the same channel, which can potentially double the link capacity and increase spectral efficiency. The main limitation in FD operation is the self-interference (SI) (also known as loopback interference) due to signal leakage at the front-end of the receiver from the transmit antennas of an FD transceiver node [1]–[4]. In order to efficiently solve this problem, several research groups have developed methods to suppress the SI [5]–[11].

However, in practice, due to many factors such as analog circuit imperfections, I/Q imbalance, etc., the SI cannot be canceled completely. Therefore, many researchers have considered the related optimization problems (such as, resource allocation, transceiver designs, etc.) in FD system design under the existence of residual SI [12]–[20]. Specifically, the sum-rate maximization problem in FD MIMO channels was considered in [13]. Cirik *et al.* [14] studied the ergodic mutual information maximization problem of FD MIMO channels and relay systems in fast fading channels. The impact of residual SI on the sum-rate performance was considered in [15]. Different power allocation algorithms were developed for FD orthogonal frequency division multiplexing (OFDM) links in [12], where the rate regions are compared in both frequency flat and frequency selective fading channels. Transceiver designs for

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FD relay systems were considered in [16] and [17]. In [18], mean square error (MSE) based transceiver design problems for FD MIMO cognitive interference channels were considered, and an iterative algorithm which optimizes the transmit and receive beamforming matrices in an alternating manner was proposed. In [19], the authors proposed transceiver designs to maximize the weighted sum-rate in FD MIMO interference channels. The authors in [20] considered the proportional fair issue in FD MIMO interference channels and a gradient projection method was proposed to solve the non-convex optimization problem.

Recently, energy harvesting communication systems that can harvest energy from radio-frequency signals have attracted considerable interest since they offer a promising solution to providing cost-effective and (ideally) perpetual power supply for wireless networks [22]. With the proliferation of wireless services, high data rates in an EH system are expected. As a result, the joint transmission of information and energy using the same waveform, which is known as simultaneous wireless information and power transfer (SWIPT), becomes a promising solution [23], [24].

Initial works in the field of SWIPT focused on point-to-point single-antenna SWIPT systems [22], [25], [26]. Afterwards, a multi-antenna SWIPT system was investigated in [23] and [27]–[30]. The work [23] characterized the rate-energy region and investigated the optimal transmission schemes in a MIMO broadcast channel. The works [27] and [28] considered the rate-energy region and linear precoder design in a two-user MIMO interference channel, respectively. In [29] and [30], transmission strategies for K-user MIMO interference channels were considered. Almost all the above works considered geometrically separated information receiver and energy receiver. In addition to that, the time switching (TS) and power splitting (PS) schemes are two promising techniques currently under consideration for SWIPT [23]. In the former, the receiver switches between information decoding (ID) and EH. In the latter, the received signal is split into two streams, such that a fraction ρ $(0 \le \rho \le 1)$ of the received signal power is used for ID while the remaining fraction $(1-\rho)$ is used for EH. PS-based transceiver design algorithms have received significant attention recently [24], [31]–[34]. The authors of [24] studied the joint beamforming and power splitting (JBPS) design for a multiuser multiple-input single-output (MISO) broadcast system with SWIPT. The JBPS problem for a K-pair MISO interference channel was considered in [31], where the authors used the semidefinite relaxation (SDR) technique to address the non-convex problem. Different from the approach of [31], an alternative second-order-cone-programming (SOCP) relaxation method was proposed in [32]. In order to harvest more power for reliable device operation, Zong *et al.* [33] considered multiple antennas at the receiver and proposed a transceiver design algorithm for SWIPT over K-pair MIMO interference channels. However, the studies in [33] only consider a common, identical PS ratio among all the receive antennas for each user, which is not an optimum PS scheme [34]. The different PS ratios from multiple receive antennas, which define a so-called PS vector, must be taken into account in the optimization problem rather than a single PS ratio. To the best of our knowledge, this problem has not yet been addressed in the literature.

B. Contributions

A transceiver design approach that considers both FD and SWIPT techniques in MIMO interference channel has been proposed in [35], where an identical PS ratio among all the receive antennas for each user was assumed. In this paper, we take one step further by combining the FD and SWIPT techniques together and studying the joint transceiver design problem for the transmit beamformers, the receive beamformers, and the *PS vectors* at the EH nodes in an FD K-pair MIMO interference channel. Two important and timely utility optimization problems are considered, i.e., the sum power minimization problem and the sum-rate maximization problem.[1]

In the former problem, the aim is to minimize the total transmission power under both transmission rate and EH constraints. An iterative algorithm based on alternating optimization (AO) is here proposed to successively optimize the transceiver coefficients. The algorithm consists of three main steps, aimed at successively optimizing: 1) the PS vectors of the EH nodes; 2) the receive beamforming vectors; and 3) the transmit beamforming vectors. The first step is carried out based on the concave-convex procedure (CCCP) [36], the second step is based on the minimum mean-square error (MMSE) criterion and the third step resorts to using the SDR technique. Since the feasible set of the transmit beamformers is enlarged when optimizing the receive beamforming vectors and the PS vectors, the transmission power is guaranteed to be non-increasing when optimizing the transmit beamformers, implying a monotonic convergence of the proposed algorithm. To reduce complexity, two simplified algorithms are then proposed where the maximum ratio transmission (MRT) [37] and the maximum signal-to-interference-leakage (SIL) beamformers are employed. The main idea of the simplified algorithms is to fix the direction of the transmit beamforming vectors based on the MRT or SIL beamformer, and then properly scale the transmit beamforming vectors to minimize the total power consumption. We prove that the simplified algorithm is guaranteed to reach a Karush-Kuhn-Tucker (KKT) solution of the simplified problem.

In the second problem, our aim is to maximize the sum-rate under additional power and EH constraints. Due to the highly non-convex nature of this problem, we first reformulate it into an equivalent weighted MMSE (WMMSE) problem by introducing suitable weight factors, such that the global optima of the two problems are identical. Then, we show that by utilizing the concept of AO and CCCP, the equivalent problem can be efficiently solved. By dividing the optimization variables into several non-overlapping groups, we show that each group of variables can be optimized alternatively, and that each subproblem either admits a closed-form solution or can

[1] Furthermore, our work is very different from the work in [33], where only the HD mode is considered and all the receive antennas for each user employ a common and identical PS ratio.

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be solved by employing the CCCP method. As above, based on the fixed MRT and SIL beamformers, the corresponding simplified algorithms are provided to reduce the computational complexity.

For each one of these utility optimization problems, the computational complexity is evaluated. Finally, simulation results are presented to validate the effectiveness of the proposed algorithm.

The main contributions of this paper are summarized as follows:

1) We establish a general framework for SWIPT in K-user MIMO interference channels with FD operation. Two important utility optimization problems, i.e. power minimization and sum-rate maximization problem under appropriate constraints, are formulated to jointly design the transmit/receive beamforming vectors and PS vectors.

2) For the power minimization problem, we propose an iterative algorithm which employs the CCCP method, the MMSE criterion, and the SDR technique to successively optimize the vectors of PS ratios, the receiving matrices and the transmit beamforming vectors. For the sum-rate maximization problem, by introducing weight factors, we transformed the original problem into an equivalent weighted MMSE minimization form, which is more suitable for the application of the AO strategy. We show that the resulting problem can be efficiently solved by employing the AO and CCCP methods.

3) For both problems, two simplified algorithms are proposed to reduce the computational complexity. The main idea of the simplified algorithms is to fix the direction of the transmit beamforming vectors by the conventional MRT and ZF techniques, and then to jointly optimize their magnitudes together with the other variables. Specially, for the sum power minimization problem, we prove that the proposed simplified algorithm converges to the set of KKT points of the simplified problem. The asymptotic complexity analysis of proposed algorithms is presented to show the effectiveness of the simplified algorithms.

C. Structure and Notations

The reminder of this paper is organized as follows. Section II presents the FD MIMO interference channel model and the problem formulation of the two optimization problems. In Section III, the transceiver design algorithms for the sum power minimization problem are developed. Section IV discusses the sum-rate maximization problem, including the equivalent reformulation and the proposed algorithms. Each one of these sections include a discussion of the convergence properties of the proposed algorithms along with complexity analysis. Finally, in Section V computer simulations are used to evaluate the performance of the proposed algorithms. Conclusions are drawn in Section VI.

Notations: Scalars, vectors and matrices are respectively denoted by lower case, boldface lower case and boldface upper case letters. For a square matrix $\mathbf{A}$, $\mathrm{Tr}(\mathbf{A})$, $\mathrm{Rank}(\cdot)$, $\mathbf{A}^T$, and $\mathbf{A}^H$ denote its trace, rank, transpose, and conjugate transpose respectively, while $\mathbf{A} \succeq \mathbf{0}$ means that $\mathbf{A}$ is a positive semidefinite matrix. $\mathrm{diag}(d_1, \ldots, d_n)$ refers to a diagonal matrix with diagonal elements $\{d_1, \ldots, d_n\}$. $E[\cdot]$ denotes the statistical expectation. $\mathbb{R}\{\cdot\}$ denotes the real part of a variable. The operator $\mathrm{vec}(\mathbf{A})$ stacks the elements of a matrix in one long column vector, $\|\cdot\|$ and $|\cdot|$ denote the Euclidean norm of a complex vector and the absolute value of a complex scalar. $\mathbb{C}^{m\times n}(\mathbb{R}^{m\times n})$ denotes the space of $m \times n$ complex (real) matrices.

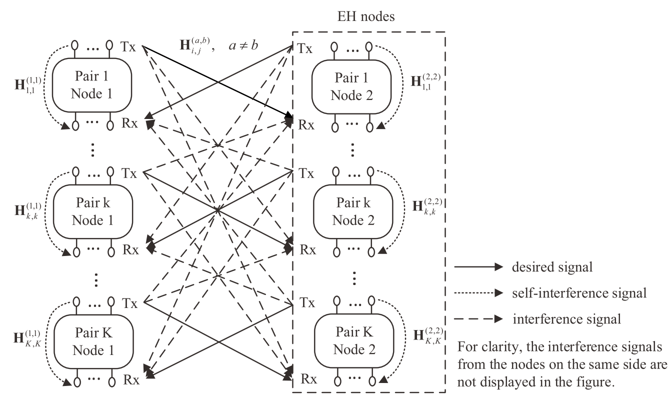

Fig. 1. Full-duplex K-pair MIMO interference channel.

II. System Model and Problem Formulation

In Subsection II-A, we develop the system model equations for the FD MIMO interference channel, along with relevant formulas for the transmission rates and harvested energy. Then, in Subsection II-B and II-C, we consider two important and popular design criteria, i.e. sum power minimization and sum-rate maximization.

A. System Model

We consider an FD MIMO interference channel with K pairs, as illustrated in Fig. 1, where each pair consists of two nodes labeled by $a \in \{1, 2\}$. We assume that the FD transceiver nodes of the ith pair, where $i \in \mathcal{K} \triangleq \{1, \ldots, K\}$, are equipped with T_i and N_i transmit and receive antennas, respectively. Each pair of nodes exchanges information simultaneously in a two way FD communication. Besides, we assume that the node 1 of each pair only contains an ID receiver, while the node 2 of each pair is equipped with both ID and EH receivers.

The received data vector at node $a \in \{1, 2\}$ of the ith pair is given by

$$\mathbf{y}_{a,i} = \mathbf{H}_{i,i}^{(a,b)} \sum_{s=1}^{M} \mathbf{f}_{s,b,i} x_{s,b,i} + \mathbf{H}_{i,i}^{(a,a)} \sum_{s=1}^{M} \mathbf{f}_{s,a,i} x_{s,a,i} + \sum_{j\neq i}^{K} \sum_{c=1}^{2} \mathbf{H}_{i,j}^{(a,c)} \sum_{s=1}^{M} \mathbf{f}_{s,c,j} x_{s,c,j} + \mathbf{n}_{a,i}, \quad a \neq b, \quad (1)$$

where $x_{s,a,i}$ denotes the sth transmit symbol (data stream) of the ath node in pair i with zero mean and $E[|x_{s,a,i}|^2] = 1$, $s \in \mathcal{M} \triangleq \{1, \ldots, M\}$, $i \in \mathcal{K}$, and $\mathbf{f}_{s,a,i} \in \mathbb{C}^{T_i \times 1}$ denotes the transmit beamforming vector for the sth symbol of the ath node in the ith pair. The matrix $\mathbf{H}_{i,i}^{(a,b)} \in \mathbb{C}^{N_i \times T_i}$ denotes the desired channel between node a and b of the ith pair, where $a, b \in \{1, 2\}$ and $a \neq b$. The matrix $\mathbf{H}_{i,i}^{(a,a)} \in \mathbb{C}^{N_i \times T_i}$ denotes the SI channel of node a. The matrix $\mathbf{H}_{i,j}^{(a,b)} \in \mathbb{C}^{N_i \times T_j}$ denotes

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the interference channel from the transmit antennas of node b in pair j to the receive antennas of node a in pair i, where $a, b \in \{1, 2\}$, and $i, j \in \mathcal{K}$, $i \neq j$. $\mathbf{n}_{a,i} \in \mathbb{C}^{N_i \times 1}$ denotes the additive white Gaussian noise vector at the ath node of pair i with zero mean and $E[\mathbf{n}_{a,i}\mathbf{n}_{a,i}^H] = \sigma_{a,i}^2\mathbf{I}$, where $\sigma_{a,i}^2$ is the average noise power. We assume $a \neq b$ hereinafter if a and b appear in the same expression.

Let $\rho_{n,i}$ $(0 \leq \rho_{n,i} \leq 1)$ denotes the PS ratio for the nth receive antenna of node 2 in the ith pair, where $n \in \mathcal{N}_i \triangleq \{1, \ldots, N_i\}$, which means that a portion $\rho_{n,i}$ of the nth receive antenna's signal power is used for signal detection while the remaining portion $1-\rho_{n,i}$ of the power is diverted to an energy harvester. By defining $\mathbf{d}_i = [\sqrt{\rho_{1,i}}, \ldots, \sqrt{\rho_{N_i,i}}]^T$ and $\Lambda_i = \mathrm{diag}(\mathbf{d}_i)$, the signal for ID at node 2 of the ith pair can be expressed as

$$\bar{\mathbf{y}}_i = \Lambda_i \Bigg(\mathbf{H}_{i,i}^{(2,1)} \sum_{s=1}^{M} \mathbf{f}_{s,1,i} x_{s,1,i} + \mathbf{H}_{i,i}^{(2,2)} \sum_{s=1}^{M} \mathbf{f}_{s,2,i} x_{s,2,i} + \sum_{j \neq i}^{K} \sum_{c=1}^{2} \mathbf{H}_{i,j}^{(2,c)} \sum_{s=1}^{M} \mathbf{f}_{s,c,j} x_{s,c,j} + \mathbf{n}_{2,i} \Bigg) + \mathbf{v}_i, \quad (2)$$

where $\mathbf{v}_i \in \mathbb{C}^{N_i \times 1}$ is the additional complex Gaussian circuit noise vector with zero mean and covariance matrix $\omega_i^2\mathbf{I}$, resulting from the phase offsets and non-linearities during baseband conversion [23].

The transmission rate of the sth symbol from node b to node a in the ith pair is given by

$$R_{s,a,i} = \log_2 \left(1 + \Gamma_{s,a,i}\right), \quad (3)$$

where $\Gamma_{s,a,i}$ denotes the corresponding signal-to-interference-plus-noise ratio (SINR), which can be written as

$$\Gamma_{s,a,i} = \frac{|\mathbf{w}_{s,a,i}^H \bar{\mathbf{H}}_{i,i}^{(a,b)} \mathbf{f}_{s,b,i}|^2}{\mathbf{w}_{s,a,i}^H \mathbf{R}_{s,a,i} \mathbf{w}_{s,a,i}}. \quad (4)$$

In this expression, $\mathbf{w}_{s,a,i} \in \mathbb{C}^{N_i \times 1}$ denotes the receive beamforming vector for the sth symbol of node a in the ith pair and

$$\begin{aligned} \mathbf{R}_{s,a,i} = &\sum_{s' \neq s}^{M} \bar{\mathbf{H}}_{i,i}^{(a,b)} \mathbf{f}_{s',b,i} \mathbf{f}_{s',b,i}^H \bar{\mathbf{H}}_{i,i}^{(a,b)H} \\ &+ \sum_{s'=1}^{M} \bar{\mathbf{H}}_{i,i}^{(a,a)} \mathbf{f}_{s',a,i} \mathbf{f}_{s',a,i}^H \bar{\mathbf{H}}_{i,i}^{(a,a)H} \\ &+ \sum_{j \neq i}^{K} \sum_{c=1}^{2} \sum_{s'=1}^{M} \bar{\mathbf{H}}_{i,j}^{(a,c)} \mathbf{f}_{s',c,j} \mathbf{f}_{s',c,j}^H \bar{\mathbf{H}}_{i,j}^{(a,c)H} + \mathbf{U}_{a,i}. \end{aligned} \quad (5)$$

Furthermore, for node $a = 1$ we have $\bar{\mathbf{H}}_{i,i}^{(1,2)} = \mathbf{H}_{i,i}^{(1,2)}$, $\bar{\mathbf{H}}_{i,i}^{(1,1)} = \mathbf{H}_{i,i}^{(1,1)}$, $\bar{\mathbf{H}}_{i,j}^{(1,c)} = \mathbf{H}_{i,j}^{(1,c)}$ and $\mathbf{U}_{1,i} = \sigma_{1,i}^2\mathbf{I}$, where $c \in \{1, 2\}$ while for node $a = 2$ we have $\bar{\mathbf{H}}_{i,i}^{(2,1)} = \Lambda_i \mathbf{H}_{i,i}^{(2,1)}$, $\bar{\mathbf{H}}_{i,i}^{(2,2)} = \Lambda_i \mathbf{H}_{i,i}^{(2,2)}$, $\bar{\mathbf{H}}_{i,j}^{(2,c)} = \Lambda_i \mathbf{H}_{i,j}^{(2,c)}$ and $\mathbf{U}_{2,i} = \sigma_{2,i}^2 \Lambda_i \Lambda_i^H + \omega_i^2\mathbf{I}$.

Besides, the total harvested energy that can be stored by node 2 of the ith pair is given by

$$\begin{aligned} P_i &= \xi_i E[\|\Phi_i \mathbf{y}_{2,i}\|^2] \\ &= \xi_i \Bigg(\sum_{s'=1}^{M} \|\Phi_i \mathbf{H}_{i,i}^{(2,1)} \mathbf{f}_{s',1,i}\|^2 + \sum_{s'=1}^{M} \|\Phi_i \mathbf{H}_{i,i}^{(2,2)} \mathbf{f}_{s',2,i}\|^2 \\ &\quad + \sum_{j \neq i}^{K} \sum_{c=1}^{2} \sum_{s'=1}^{M} \|\Phi_i \mathbf{H}_{i,j}^{(2,c)} \mathbf{f}_{s',c,j}\|^2 + \sigma_{2,i}^2 \|\mathrm{vec}\{\Phi_i\}\|^2 \Bigg), \end{aligned} \quad (6)$$

where $\Phi_i = \mathrm{diag}\{[\sqrt{1-\rho_{1,i}}, \ldots, \sqrt{1-\rho_{N_i,i}}\,]\}$ and $\xi_i \in (0, 1]$ denotes the energy conversion efficiency of the EH unit in pair i.

Remark 1: Note that the considered system model can be viewed as the combination of three innovative techniques, i.e. coordinated multipoint (CoMP) transmission/reception, FD and EH. The nodes 1 in different pairs coordinate their operation with each other in such a way that the transmitted signals from/to other nodes do not incur serious interference, but in fact can be exploited in a meaningful way. The key deployment scenarios for CoMP have been identified as part of various studies performed by the Third Generation Partnership Project (3GPP) for the Long Term Evolution (LTE)-Advanced standard development [38], [39]; furthermore, CoMP will be more widely applied in 5G not only to address the interference, but also to improve the system resource utilization and data rates [40]. Meanwhile, FD and EH are two promising techniques to improve the spectral and energy efficiency under consideration for 5G wireless networks [21], [41]. They are currently the focus of intense research efforts by various groups in academics and industries worldwide.

B. Problem Formulation – Sum Power Minimization

In this subsection, we first formulate the optimization problem for the joint design of $\{\mathbf{w}_{s,a,i}\}$, $\{\mathbf{f}_{s,a,i}\}$ and $\{\mathbf{d}_i\}$ so as to minimize the total power consumption under the constraint that a set of minimum transmission rate and EH targets[2] be satisfied. The optimization problem can be expressed as follows:

$$\begin{aligned} &\min_{\{\mathbf{w}_{s,a,i}\}, \{\mathbf{f}_{s,a,i}\}, \{\mathbf{d}_i\}} \sum_{i=1}^{K} \sum_{a=1}^{2} \sum_{s=1}^{M} \mathbf{f}_{s,a,i}^H \mathbf{f}_{s,a,i} \\ &\text{s.t. } R_{s,a,i} \geq \varphi_{s,a,i},\ P_i \geq \psi_i, \\ &\quad 0 \leq \rho_{n,i} \leq 1, n \in \mathcal{N}_i, i \in \mathcal{K}, s \in \mathcal{M}, \quad a \in \{1, 2\}. \end{aligned} \quad (7)$$

The QoS constraints require that the rate of each data stream should be no smaller than a given positive target $\varphi_{s,a,i}$. In the meantime, the EH constraints call for the harvested energy of each receiver to be no smaller than a positive threshold ψ_i. As we can see, the design variables are highly coupled in the rate and EH constraints, which makes problem (7) very difficult to handle. It is worth mentioning that the received energy at each receiver should be above some threshold in order to activate the EH circuit [42]. This can be guaranteed

[2]In this work, we assume that each data stream supports a different type of service. Therefore, we consider the rate constraint of each data stream.

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by performing proper user scheduling, e.g., according to the distance between the RS and users. This paper focus on joint transceiver design with QoS guarantee for the scheduled users, as in most of the recent literature on EH related beamforming design, e.g. [31], [43], [44]. The same argument also applies to the sum-rate maximization problem presented in Subsection II-C.

C. Problem Formulation – Sum-Rate Maximization

Here, we present another problem formulation for the joint design of $\{\mathbf{w}_{s,a,i}\}$, $\{\mathbf{f}_{s,a,i}\}$ and $\{\mathbf{d}_i\}$. The focus of this design criterion is to maximize the sum-rate utility function under the condition that a set of minimum transmission rate and EH targets are satisfied and the power budget is respected. The optimization problem can be formulated as:

$$\max_{\{\mathbf{w}_{s,a,i}\},\{\mathbf{f}_{s,a,i}\},\{\mathbf{d}_i\}} \sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M} R_{s,a,i} \tag{8a}$$

$$\text{s.t. } R_{s,a,i} \geq \varphi_{s,a,i}, \quad P_i \geq \psi_i, \tag{8b}$$

$$\sum_{s=1}^{M} \mathbf{f}_{s,a,i}^H \mathbf{f}_{s,a,i} \leq P_{a,i}, \tag{8c}$$

$$\sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M} \mathbf{f}_{s,a,i}^H \mathbf{f}_{s,a,i} \leq P_{sum}, \tag{8d}$$

$$0 \leq \rho_{n,i} \leq 1, \quad n \in \mathcal{N}_i, \; i \in \mathcal{K}, \; s \in \mathcal{M}, \; a \in \{1,2\}. \tag{8e}$$

We note that problem (8) is more difficult than problem (7) to address since the objective function of (8) is also non-convex while the $\log_2(\cdot)$ operator from (3) is difficult to handle. The purpose of the constraint (8c) is to restrain the individual transmitter power so as to meet the hardware limitations, while the constraint (8d) can ensure that the power allocation among different nodes is more efficient. This can be intuitively explained by observing that node 1 of a given pair usually requires more power than node 2, since the latter needs to harvest energy from the former.

III. Transceiver Design With Sum Power Minimization

In this section, we address the sum power minimization problem (7), which is clearly not a convex problem and difficult to solve. Our basic idea for solving problem (7) is to optimize the design variables in an alternative manner. In the proposed design, we show that each subset of variables $\{\mathbf{f}_{s,a,i}\}$, $\{\mathbf{w}_{s,a,i}\}$ and $\{\mathbf{d}_i\}$ can be successively optimized in turn with the two other subsets being fixed. Each subproblem for the optimization of a given subset of the complete set of variables can be formulated as a simplier problem which can be solved efficiently, while the monotonic convergence of the sequential AO-based algorithm is guaranteed. Furthermore, two simplified algorithms are proposed based on the use of fixed MRT and SIL beamformers, respectively. We show that the simplified algorithms are guaranteed to achieve a Karush-Kuhn-Tucker (KKT) solution of the corresponding problems and consume much less computational resources compared to the AO-based algorithm while still maintaining a good performance.

A. Proposed Alternating Optimization Design Algorithm

In this subsection, we propose an AO-based approach for the joint transceiver design. Firstly, by fixing the transmit beamforming vectors $\{\mathbf{f}_{s,a,i}\}$ and the PS vectors $\{\mathbf{d}_i\}$, we can obtain the receive beamforming vectors by the MMSE approach[3]:

$$\mathbf{w}_{s,a,i} = (\bar{\mathbf{H}}_{i,i}^{(a,b)} \mathbf{f}_{s,b,i} \mathbf{f}_{s,b,i}^H \bar{\mathbf{H}}_{i,i}^{(a,b)H} + \mathbf{R}_{s,a,i})^{-1} \bar{\mathbf{H}}_{i,i}^{(a,b)} \mathbf{f}_{s,b,i}, \tag{9}$$

where $a \in \{1,2\}$.

Secondly, let us consider the optimization of $\{\mathbf{f}_{s,a,i}\}$ by fixing $\{\mathbf{d}_i\}$ and $\{\mathbf{w}_{s,a,i}\}$. We resort to the SDR technique to solve this problem. By introducing a new variable $\mathbf{F}_{s,a,i} = \mathbf{f}_{s,a,i}\mathbf{f}_{s,a,i}^H$ and ignoring the rank-one constraints for all $\mathbf{F}_{s,a,i}$, problem (7) can be reformulated as a convex SDP problem as shown in (10), at the bottom of this page, where $\gamma_{s,a,i} = 2^{\varphi_{s,a,i}} - 1$,

[3]The MMSE receiver is designed such that the SINR of the receiver is maximized.

$$\begin{aligned}
&\min_{\{\mathbf{F}_{s,a,i}\}} \sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M} \mathrm{Tr}\{\mathbf{F}_{s,a,i}\} \\
&\text{s.t. } \frac{1}{\gamma_{s,b,i}} \mathbf{w}_{s,b,i}^H \bar{\mathbf{H}}_{i,i}^{(b,a)} \mathbf{F}_{s,a,i} \bar{\mathbf{H}}_{i,i}^{(b,a)H} \mathbf{w}_{s,b,i} - \sum_{s' \neq s}^{M} \mathbf{w}_{s,b,i}^H \bar{\mathbf{H}}_{i,i}^{(b,a)} \mathbf{F}_{s',a,i} \bar{\mathbf{H}}_{i,i}^{(b,a)H} \mathbf{w}_{s,b,i} \\
&\quad - \sum_{s'=1}^{M} \mathbf{w}_{s,b,i}^H \bar{\mathbf{H}}_{i,i}^{(b,b)} \mathbf{F}_{s',b,i} \bar{\mathbf{H}}_{i,i}^{(b,b)H} \mathbf{w}_{s,b,i} - \sum_{j \neq i}^{K}\sum_{c=1}^{2}\sum_{s'=1}^{M} \mathbf{w}_{s,b,i}^H \bar{\mathbf{H}}_{i,j}^{(b,c)} \mathbf{F}_{s',c,j} \bar{\mathbf{H}}_{i,j}^{(b,c)H} \mathbf{w}_{s,b,i} \geq \theta_{b,i}, \\
&\quad \sum_{s'=1}^{M} \mathrm{Tr}\{\Phi_i \mathbf{H}_{i,i}^{(2,1)} \mathbf{F}_{s',1,i} \mathbf{H}_{i,i}^{(2,1)H} \Phi_i\} + \sum_{s'=1}^{M} \mathrm{Tr}\{\Phi_i \mathbf{H}_{i,i}^{(2,2)} \mathbf{F}_{s',2,i} \mathbf{H}_{i,i}^{(2,2)H} \Phi_i\} \\
&\quad + \sum_{j \neq i}^{K}\sum_{c=1}^{2}\sum_{s'=1}^{M} \mathrm{Tr}\{\Phi_i \mathbf{H}_{i,j}^{(2,c)} \mathbf{F}_{s',c,j} \mathbf{H}_{i,j}^{(2,c)H} \Phi_i\} \geq \frac{\psi_i}{\xi_i} - \sigma_{2,i}^2 \|\mathrm{vec}\{\Phi_i\}\|^2, \\
&\quad \mathbf{F}_{s,a,i} \succeq \mathbf{0}, \; i \in \mathcal{K}, s \in \mathcal{M}, a, b \in \{1,2\},
\end{aligned} \tag{10}$$

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$\theta_{1,i} = \sigma_{1,i}^2 \mathbf{w}_{s,1,i}^H \mathbf{w}_{s,1,i}$ and $\theta_{2,i} = \sigma_{2,i}^2 \mathbf{w}_{s,2,i}^H \Lambda_i \Lambda_i^H \mathbf{w}_{s,2,i} + \omega_i^2 \mathbf{w}_{s,2,i}^H \mathbf{w}_{s,2,i}$. Therefore, problem (10) can be efficiently solved by off-the-shelf algorithms [45]. It is worth mentioning that, the optimal solution $\{\mathbf{F}_{s,a,i}^*\}$ to problem (10) cannot be proven to be of rank-one in general. However, we have the following lemma to address the rank property of problem (10) when $K = 2$.

Lemma 1: When $K = 2$, there always exist optimal rank-one solutions (i.e. $\text{Rank}(\mathbf{F}_{s,a,i}) = 1, \forall s, a, i$) of problem (10), which are also optimal to the problem before SDR, regardless of the stream number M.

The proof is relegated to Appendix A. To guarantee a rank one solution when $K \geq 3$, we may use a simple rank-one recovery method based on the Gaussian randomization procedure. Interested readers may refer to [46] for details.

Thirdly, let us optimize the PS vectors $\{\mathbf{d}_i\}$ by fixing $\{\mathbf{f}_{s,a,i}\}$ and $\{\mathbf{w}_{s,a,i}\}$. Note that the objective function of (7) does not include $\{\mathbf{d}_i\}$. As a result, directly solving problem (7) with respect to the PS vectors $\{\mathbf{d}_i\}$ (i.e., finding a set of feasible PS vectors) may not improve the objective value. This observation motivates us to devise an alternative strategy, aiming at achieving a nondecreasing transmission power in the next iteration. Specifically, we solve for $\{\mathbf{d}_i\}$ in the following optimization problem:

$$\begin{aligned} &\max_{t,\{\mathbf{d}_i\}} \ t \\ &\text{s.t.} \ \frac{\Gamma_{s,2,i}}{\gamma_{s,2,i}} \geq t, \quad \frac{P_i}{\psi_i} \geq t, \\ &\qquad 0 \leq \rho_{n,i} \leq 1, \quad n \in \mathcal{N}_i, \ i \in \mathcal{K}, \ s \in \mathcal{M}. \end{aligned} \tag{11}$$

It is readily seen that once problem (11) is solved, t must be no less than 1. This implies that the feasible region of problem (7) with respect to $\{\mathbf{f}_{s,a,i}\}$ and $\{\mathbf{w}_{s,a,i}\}$ in the next iteration would become larger, which often leads to a decreased transmission power.

Let us define a vector of parameters $\mathbf{s} = [t, \mathbf{d}_1^T, \ldots, \mathbf{d}_K^T]^T \in \mathbb{R}^{(\sum_{i=1}^K N_i + 1)}$. Then problem (11) can be equivalently rewritten as (12), at the bottom of this page, where the objective function $f(\mathbf{s}) = t$,

$$\begin{aligned} u_{s,i}(\mathbf{s}) = &\sum_{s' \neq s}^{M} |\mathbf{w}_{s,2,i}^H \text{diag}\{\mathbf{H}_{i,i}^{(2,1)} \mathbf{f}_{s',1,i}\} \mathbf{d}_i|^2 \\ &+ \sum_{s'=1}^{M} |\mathbf{w}_{s,2,i}^H \text{diag}\{\mathbf{H}_{i,i}^{(2,2)} \mathbf{f}_{s',2,i}\} \mathbf{d}_i|^2 \\ &+ \sum_{j \neq i}^{K} \sum_{c=1}^{2} \sum_{s'=1}^{M} |\mathbf{w}_{s,2,i}^H \text{diag}\{\mathbf{H}_{i,j}^{(2,c)} \mathbf{f}_{s',c,j}\} \mathbf{d}_i|^2 + \theta_{2,i} \end{aligned} \tag{13}$$

and $\upsilon_{s,i}(\mathbf{s}) = \frac{|\mathbf{w}_{s,2,i}^H \text{diag}\{\mathbf{H}_{i,i}^{(2,1)} \mathbf{f}_{s,1,i}\} \mathbf{d}_i|^2}{t \gamma_{s,2,i}}$. Note that the first constraint (12b) is a difference of convex (DC) constraint and thus, according to the CCCP method, we approximate the convex function $\upsilon_{s,i}(\mathbf{s})$ in the kth iteration by its first order Taylor expansion around the current point $\mathbf{s}^{(k)}$, denoted as $\hat{\upsilon}_{s,i}(\mathbf{s}^{(k)}, \mathbf{s})$. Based on [47] and [48], $\hat{\upsilon}_{s,i}(\mathbf{s}^{(k)}, \mathbf{s})$ is given by

$$\hat{\upsilon}_{s,i}(\mathbf{s}^{(k)}, \mathbf{s}) = \upsilon_{s,i}(\mathbf{s}^{(k)}) + 2\Re\{\nabla \upsilon_{s,i}^H(\mathbf{s}^{(k)})(\mathbf{s} - \mathbf{s}^{(k)})\}, \tag{14}$$

where $\nabla \upsilon_{s,i}(\mathbf{s}^{(k)})$ denotes the conjugate derivative of $\upsilon_{s,i}(\mathbf{s}^{(k)})$. Due to limited space, we omit the detailed results of this calculation for $\hat{\upsilon}_{s,i}(\mathbf{s}^{(k)}, \mathbf{s})$. Note that $\hat{\upsilon}_{s,i}(\mathbf{s}^{(k)}, \mathbf{s})$ is an affine function of $\mathbf{s}$. By introducing the following variables:

$$\begin{aligned} \mathbf{x}_{s,i,i}^{(a,b)} = \Big[&\mathbf{w}_{s,a,i}^H \text{diag}\{\mathbf{H}_{i,i}^{(a,b)} \mathbf{f}_{1,b,i}\} \mathbf{d}_i, \ldots, \\ &\mathbf{w}_{s,a,i}^H \text{diag}\{\mathbf{H}_{i,i}^{(a,b)} \mathbf{f}_{s-1,b,i}\} \mathbf{d}_i, \\ &\mathbf{w}_{s,a,i}^H \text{diag}\{\mathbf{H}_{i,i}^{(a,b)} \mathbf{f}_{s+1,b,i}\} \mathbf{d}_i, \ldots, \\ &\mathbf{w}_{s,a,i}^H \text{diag}\{\mathbf{H}_{i,i}^{(a,b)} \mathbf{f}_{M,b,i}\} \mathbf{d}_i \Big]^T, \end{aligned} \tag{15}$$

$$\begin{aligned} \mathbf{y}_{s,i,j}^{(a,b)} = \Big[&\mathbf{w}_{s,a,i}^H \text{diag}\{\mathbf{H}_{i,j}^{(a,b)} \mathbf{f}_{1,b,j}\} \mathbf{d}_i, \ldots, \\ &\mathbf{w}_{s,a,i}^H \text{diag}\{\mathbf{H}_{i,j}^{(a,b)} \mathbf{f}_{M,b,j}\} \mathbf{d}_i \Big]^T, \end{aligned} \tag{16}$$

$$\begin{aligned} \mathbf{z}_{s,i} = [&\mathbf{y}_{s,i,1}^{(2,1)T}, \ldots, \mathbf{y}_{s,i,i-1}^{(2,1)T}, \mathbf{y}_{s,i,i+1}^{(2,1)T}, \ldots, \\ &\mathbf{y}_{s,i,K}^{(2,1)T}, \mathbf{y}_{s,i,1}^{(2,2)T}, \ldots, \mathbf{y}_{s,i,i-1}^{(2,2)T}, \mathbf{y}_{s,i,i+1}^{(2,2)T}, \ldots, \mathbf{y}_{s,i,K}^{(2,2)T}]^T, \end{aligned} \tag{17}$$

$$\max_{\mathbf{s}} \ f(\mathbf{s}) \tag{12a}$$

$$\text{s.t.} \ u_{s,i}(\mathbf{s}) - \upsilon_{s,i}(\mathbf{s}) \leq 0, \tag{12b}$$

$$\begin{aligned} &\sum_{s'=1}^{M} \|\text{diag}\{\mathbf{H}_{i,i}^{(2,1)} \mathbf{f}_{s',1,i}\} \mathbf{d}_i\|^2 + \sum_{s'=1}^{M} \|\text{diag}\{\mathbf{H}_{i,i}^{(2,2)} \mathbf{f}_{s',2,i}\} \mathbf{d}_i\|^2 + \sum_{j \neq k}^{K} \sum_{c=1}^{2} \sum_{s'=1}^{M} \|\text{diag}\{\mathbf{H}_{i,j}^{(2,c)} \mathbf{f}_{s',c,j}\} \mathbf{d}_i\|^2 + \sigma_{2,i}^2 \|\mathbf{d}_i\|^2 \\ &\leq \sum_{s'=1}^{M} \|\mathbf{H}_{i,i}^{(2,1)} \mathbf{f}_{s',1,i}\|^2 + \sum_{s'=1}^{M} \|\mathbf{H}_{i,i}^{(2,2)} \mathbf{f}_{s',2,i}\|^2 + \sum_{j \neq i}^{K} \sum_{c=1}^{2} \sum_{s'=1}^{M} \|\mathbf{H}_{i,j}^{(2,c)} \mathbf{f}_{s',c,j}\|^2 + \sigma_{2,i}^2 N_i - \frac{t \psi_i}{\xi_i}, \end{aligned} \tag{12c}$$

$$0 \leq \rho_{n,i} \leq 1, \quad n \in \mathcal{N}_i, \ i \in \mathcal{K}, \ s \in \mathcal{M}. \tag{12d}$$

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TABLE I

THE CCCP BASED ITERATIVE ALGORITHM TO SOLVE (12)

1. Define the tolerance of accuracy δ_1 and the maximum number of iteration N_{max}^1. Initialize the algorithm with a feasible point $\mathbf{s}^{(0)}$. Set the iteration number $k=0$.
2. **Repeat**
   - Solve problem (23) with the affine approximation $\hat{v}_{s,i}(\mathbf{s}^{(k)},\mathbf{s})$, and assign the solution to $\mathbf{s}^{(k+1)}$.
   - Update the iteration number : $k=k+1$.
3. **Until** $|f(\mathbf{s}^{(k+1)})-f(\mathbf{s}^{(k)})|\le\delta$ or the maximum number of iterations is reached, i.e., $k>N_{max}^1$.

TABLE II

AO-BASED ALGORITHM FOR THE SUM POWER MINIMIZATION PROBLEM

1. Initialize $\{\mathbf{w}_{s,a,i}\}$ and $\{\mathbf{f}_{s,a,i}\}$, define the tolerance of accuracy δ_2 and the maximum number of iteration N_{max}^2.
2. **Repeat**
   - 2.1 Solve an approximation of problem (12) based on Table I with fixed $\{\mathbf{w}_{s,a,i}\}$ and $\{\mathbf{f}_{s,a,i}\}$ to obtain the updated $\{\mathbf{d}_i\}$.
   - 2.2 Compute (9) with fixed $\{\mathbf{f}_{s,a,i}\}$ and $\{\mathbf{d}_i\}$ to obtain the updated $\{\mathbf{w}_{s,a,i}\}$.
   - 2.3 Solve problem (10) with fixed $\{\mathbf{w}_{s,a,i}\}$ and $\{\mathbf{d}_i\}$ to obtain the updated $\{\mathbf{f}_{s,a,i}\}$.
3. **Until** the total power consumption between two successive iterations is less than δ_2 or the maximum number of iterations is reached.

constraint (12b) can be reformulated as

$$\left\|\left[\mathbf{x}_{s,i,i}^{(2,1)T},\mathbf{y}_{s,i,i}^{(2,2)T},\mathbf{z}_{s,i}^{T},\sigma_{2,i}(\mathrm{diag}\{\mathbf{w}_{s,2,i}\}^H\mathbf{d}_i)^T,\right.\right.$$
$$\left.\left.\omega_i\mathbf{w}_{s,2,i}^T,\frac{\hat{v}_{s,i}(\mathbf{s}^{(k)},\mathbf{s})-1}{2}\right]\right\|\le\frac{\hat{v}_{s,i}(\mathbf{s}^{(k)},\mathbf{s})+1}{2}. \quad (18)$$

Similarly, with the following variables:

$$\widetilde{\mathbf{x}}_{i,j}^{(a,b)}=\Big[(\mathrm{diag}\{\mathbf{H}_{i,j}^{(a,b)}\bar{\mathbf{f}}_{1,b,j}\}\mathbf{d}_i)^T,\ldots,$$
$$(\mathrm{diag}\{\mathbf{H}_{i,j}^{(a,b)}\bar{\mathbf{f}}_{M,b,j}\}\mathbf{d}_i)^T\Big]^T, \quad (19)$$
$$\widetilde{\mathbf{z}}_i=\Big[\widetilde{\mathbf{x}}_{i,1}^{(2,1)T},\ldots,\widetilde{\mathbf{x}}_{i,i-1}^{(2,1)T},\widetilde{\mathbf{x}}_{i,i+1}^{(2,1)T},\ldots,$$
$$\widetilde{\mathbf{x}}_{i,K}^{(2,1)T},\widetilde{\mathbf{x}}_{i,1}^{(2,2)T},\ldots,\widetilde{\mathbf{x}}_{i,i-1}^{(2,2)T},\mathbf{x}_{i,i+1}^{(2,2)T},\ldots,\widetilde{\mathbf{x}}_{i,K}^{(2,2)T}\Big]^T, \quad (20)$$

constraint (12c) can be formulated as

$$\left\|\left[\widetilde{\mathbf{x}}_{i,i}^{(2,1)T},\widetilde{\mathbf{x}}_{i,i}^{(2,2)T},\mathbf{z}_i^T,\sigma_{2,i}\mathbf{d}_i^T,\frac{q_i-\frac{t\psi_i}{\xi_i}-1}{2}\right]\right\|\le\frac{q_i-\frac{t\psi_i}{\xi_i}+1}{2}, \quad (21)$$

where

$$q_i=\sum_{s'=1}^{M}\|\mathbf{H}_{i,i}^{(2,1)}\mathbf{f}_{s',1,i}\|^2+\sum_{s'=1}^{M}\|\mathbf{H}_{i,i}^{(2,2)}\mathbf{f}_{s',2,i}\|^2$$
$$+\sum_{j\neq i}^{K}\sum_{c=1}^{2}\sum_{s'=1}^{M}\|\mathbf{H}_{i,j}^{(2,c)}\mathbf{f}_{s',c,j}\|^2+\sigma_{2,i}^2N_i. \quad (22)$$

Thus, in the kth iteration of the proposed CCCP based algorithm, we have the following SOCP problem

$$\max_{\mathbf{s}}\ f(\mathbf{s})$$
$$\text{s.t. } (18),\ (21),\ 0\le\rho_{n,i}\le 1,\quad n\in\mathcal{N}_i,i\in\mathcal{K},s\in\mathcal{M}. \quad (23)$$

The CCCP based iterative algorithm to solve problem (12) is summarized in Table I, while the proposed alternating optimization algorithm is summarized in Table II.

The proposed alternating optimization design algorithm iterates over steps 2.1-2.3. We note that in step 2.1 and 2.2, the feasible region of the problem is enlarged, and consequently the total transmission power is decreased in step 2.3. Thus, the monotonic convergence of the algorithm is ensured.

Furthermore, the complexity of the alternating optimization design algorithm is dominated by solving problem (10) N_{max}^2 times and solving problem (12) $N_{max}^1N_{max}^2$ times. Similar to the analysis in [46], the complexity of solving (10) can be expressed as $C_1=n_1\sqrt{2KM(N+1)+K}[2KM+K+2KMN^3+n_1(2KM+K+2KMN^2)+n_1^2]$, where $n_1=O(2KMN^2)$, and the complexity of solving problem (12) can be expressed as $C_2=n_2\sqrt{2(M+N)K+2K}[2NK+2NKn_2+MK(2MK+N+2)^2+K(2NMK+N+2)^2+n_2^2]$, where $n_2=O(NK+1)$. Thus, the overall complexity can be expressed as $N_{max}^2C_1+N_{max}^1N_{max}^2C_2=O(N_{max}^2N^{13.5}+N_{max}^1N_{max}^2N^{10})$.[4]

B. Proposed Simplified Algorithm With Fixed Beamformers

In the previous subsection, we proposed an AO-based transceiver design algorithm. As will be shown in Section V, the AO-based algorithm achieves the best performance, but is characterized by high computational complexity. Note that the complexity of this algorithm is dominated by solving the SDP problem (10) and the DC problem (11), where (10) involves the optimization of the transmit beamforming vectors and (11) deals with that of the PS vectors. In this subsection, motivated by these considerations, two simplified algorithms are proposed in which we fix the direction of the transmit beamforming vectors with the aid of the MRT and SIL beamformers, respectively, and only optimize the corresponding amplitude. Moreover, the same PS ratio for all the receive antennas of node 2 ($\rho_{n,i}=\rho_i,\forall n\in\mathcal{N}_i$) is assumed to simplify the original problem (7). Thus, the number of optimization variables can be significantly reduced and so we can obtain the transceiver parameters with lower complexity.

The MRT beamformer maximizes the signal-to-noise ratio (SNR) at each receiver and only requires the knowledge of the direct links $\mathbf{H}_{i,i}^{(a,b)}$. Under the current modelling assumption, the MRT beamformer $\mathbf{f}_{s,b,i}^{\mathrm{MRT}}$ is the solution of the following problem:

$$\mathbf{f}_{s,b,i}^{\mathrm{MRT}}=\max_{\|\mathbf{f}_{s,b,i}\|\le 1}\ \|\mathbf{H}_{i,i}^{(a,b)}\mathbf{f}_{s,b,i}\|^2. \quad (24)$$

Obviously, the solution of (24) is the normalized eigenvector of the matrix $\mathbf{H}_{i,i}^{(a,b)H}\mathbf{H}_{i,i}^{(a,b)}$ corresponding to the

[4]The asymptotic complexity of the proposed algorithms is evaluated when N, K and M are large, i.e. by letting $M=N=K\to\infty$.

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largest eigenvalue. It is worth mentioning that the MRT beamforming scheme does not take the various forms of interference into consideration and therefore it results in lower SINR or data rate. However, although the remaining interference is the major concern for conventional MIMO systems, it could be beneficial for the scenarios with EH.

The SIL beamformer maximizes the SIL ratio at each transmitter. In this case, the channel from node a of the ith pair to all other nodes can be seen as a broadcast channel where node b of the ith pair receives the desired signal and the other nodes receive the interference leakage. Thus, the SIL beamformer $\mathbf{f}_{s,b,i}^{\mathrm{SIL}}$ is the solution of the following problem:

$$\mathbf{f}_{s,b,i}^{\mathrm{SIL}} = \max_{\mathbf{f}_{s,b,i}} \frac{\|\mathbf{H}_{i,i}^{(a,b)}\mathbf{f}_{s,b,i}\|^2}{\sum_{j\neq i}^{K}\|\mathbf{H}_{j,i}^{(a,b)}\mathbf{f}_{s,b,i}\|^2 + \sum_{j=1}^{K}\|\mathbf{H}_{j,i}^{(b,b)}\mathbf{f}_{s,b,i}\|^2}. \quad (25)$$

We note that the optimal SIL beamformer is the maximum of a Rayleigh quotient [49], which admits closed-form solutions. It can be seen that the SIL beamformer provides a tradeoff between the desired signal and the interference.

With the fixed beamformers described above, problem (7) can be transformed into the following design problem:

$$\begin{aligned}
&\min_{\{p_{s,a,i}\},\{\rho_i\},\{\mathbf{w}_{s,a,i}\}} \sum_{s=1}^{M}\sum_{a=1}^{2}\sum_{i=1}^{K} p_{s,a,i}\\
&\text{s.t.}\ \frac{p_{s,2,i}|\mathbf{w}_{s,1,i}^H\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}|^2}{\mathbf{w}_{s,1,i}^H(\mathbf{R}_{s,1,i}^{(1)}+\mathbf{R}_{s,1,i}^{(2)}+\mathbf{R}_{s,1,i}^{(3)}+\sigma_{1,i}^2\mathbf{I})\mathbf{w}_{s,1,i}} \geq \gamma_{s,1,i},\\
&\frac{p_{s,1,i}|\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s,1,i}|^2}{\mathbf{w}_{s,2,i}^H(\mathbf{R}_{s,2,i}^{(1)}+\mathbf{R}_{s,2,i}^{(2)}+\mathbf{R}_{s,2,i}^{(3)}+\sigma_{2,i}^2\mathbf{I}+\frac{\omega_i^2}{\rho_i}\mathbf{I})\mathbf{w}_{s,2,i}}\\
&\geq \gamma_{s,2,i},\quad p_{s,1,i}\|\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s,1,i}\|^2\\
&+\mathrm{Tr}(\mathbf{R}_{s,2,i}^{(1)}+\mathbf{R}_{s,2,i}^{(2)}+\mathbf{R}_{s,2,i}^{(3)}) \geq \frac{\psi_i}{\xi_i(1-\rho_i)} - \sigma_{2,i}^2 N_i,\\
&p_{s,1,i}\geq 0,\quad p_{s,2,i}\geq 0,\ 0\leq\rho_i\leq 1,\ i\in\mathcal{K},\ s\in\mathcal{M},
\end{aligned} \quad (26)$$

where $\widetilde{\mathbf{f}}_{s,a,i} = \mathbf{f}_{s,b,i}^{\mathrm{MRT/SIL}}$ denotes the direction of the fixed beamformer obtained by either one of the MRT and SIL approaches, $p_{s,a,i}$ denotes the corresponding transmit power of the fixed beamformer,

$$\mathbf{R}_{s,a,i}^{(1)} = \sum_{s'\neq s}^{M} p_{s',b,i}\mathbf{H}_{i,i}^{(a,b)}\widetilde{\mathbf{f}}_{s',b,i}\widetilde{\mathbf{f}}_{s',b,i}^H\mathbf{H}_{i,i}^{(a,b)H}, \quad (27)$$

$$\mathbf{R}_{s,a,i}^{(2)} = \sum_{s'=1}^{M} p_{s',a,i}\mathbf{H}_{i,i}^{(a,a)}\widetilde{\mathbf{f}}_{s',a,i}\widetilde{\mathbf{f}}_{s',a,i}^H\mathbf{H}_{i,i}^{(a,a)H}, \quad (28)$$

$$\mathbf{R}_{s,a,i}^{(3)} = \sum_{j\neq i}^{K}\sum_{c=1}^{2}\sum_{s'=1}^{M} p_{s',c,j}\mathbf{H}_{i,j}^{(a,c)}\widetilde{\mathbf{f}}_{s',c,j}\widetilde{\mathbf{f}}_{s',c,j}^H\mathbf{H}_{i,j}^{(a,c)H}. \quad (29)$$

As we can see, the variables $\{p_{s,a,i}\}$, $\{\rho_i\}$ and $\{\mathbf{w}_{s,a,i}\}$ are still coupled together in problem (26). Thus, we propose to successively optimize the variables $\{p_{s,a,i},\rho_i\}$ and $\{\mathbf{w}_{s,a,i}\}$.

TABLE III

SIMPLIFIED ALGORITHM FOR THE SUM POWER MINIMIZATION PROBLEM

1. Calculate the transmit beamforming vectors $\{\mathbf{f}_{s,a,i}\}$ by (24) or (25). Initialize the PS ratios $\{\rho_i\}$ and calculate the receive beamforming vector based on (9). Define the tolerance of accuracy δ_2 and the maximum number of iteration N_{max}^2.
2. **Repeat**
   - 2.1 Solve problem (26) with fixed $\{\mathbf{w}_{s,a,i}\}$ to obtain the updated $\{p_{s,a,i}\}$ and $\{\rho_i\}$.
   - 2.2 Compute (9) with fixed $\{p_{s,a,i}\}$ and $\{\rho_i\}$ to obtain the updated $\{\mathbf{w}_{s,a,i}\}$.
3. **Until** the change in total power consumption between two successive iterations is less than δ_2 or the maximum number of iterations is reached.

We show that each subproblem can be formulated as a convex problem which can be efficiently solved.

Firstly, with fixed $\{p_{s,a,i}\}$ and $\{\rho_i\}$, we can obtain the receive beamforming vectors $\{\mathbf{w}_{s,a,i}\}$ from (9) by setting $\rho_{n,i} = \rho_i$ and $\mathbf{f}_{s,a,i} = \sqrt{p_{s,a,i}}\widetilde{\mathbf{f}}_{s,a,i}$. Secondly, by fixing the receive beamforming vectors $\{\mathbf{w}_{s,a,i}\}$, it can be seen that the resulting optimization problem in $\{p_{s,a,i}\}$ and $\{\rho_i\}$ is a convex problem since both $\frac{1}{\rho_i}$ and $\frac{1}{1-\rho_i}$ are convex functions of ρ_i with $0 < \rho_i < 1$. Actually, the resulting problem can be further formulated as an SOCP problem by introducing certain slack variables; interested readers can refer to [50] for the a similar derivation. The overall simplified algorithm with fixed beamformer is summarized in Table III. We now present two lemmas which will be used later in the convergence proof.

Lemma 2: Define set $\mathcal{L} = \{(s,a,i)|R_{s,a,i} = \varphi_{s,a,i}\}$ which contains all the indexes of (s,a,i) such that $\{\mathbf{w}_{s,a,i}, p_{s,a,i}, \rho_i\}$ satisfy the SINR constraints of problem (26) with equality and $\{\mathbf{w}_{s,a,i}\}$ is the MMSE receiver (9). Then we have

$$\Big(\mathbf{R}_{s,1,i} - \frac{p_{s,2,i}}{\gamma_{s,1,i}}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}\widetilde{\mathbf{f}}_{s,2,i}^H\mathbf{H}_{i,i}^{(1,2)H}\Big)\mathbf{w}_{s,1,i} = \mathbf{0},\quad \forall(s,1,i)\in\mathcal{L}, \quad (30)$$

$$\Big(\mathbf{R}_{s,2,i} - \frac{\rho_i p_{s,1,i}}{\gamma_{s,2,i}}\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s,1,i}\widetilde{\mathbf{f}}_{s,1,i}^H\mathbf{H}_{i,i}^{(2,1)H}\Big)\mathbf{w}_{s,2,i} = \mathbf{0},\quad \forall(s,2,i)\in\mathcal{L}. \quad (31)$$

Proof: Refer to Appendix B. ■

Lemma 3: The iterates generated by the simplified algorithm in Table III produce a non-increasing sequence of objective values, and converge to the set of KKT points of problem (26).

Proof: Please refer to Appendix C. ■

Similar to the previous subsection, the asymptotic complexity of the simplified algorithm can be expressed as $O(N_{max}^1 N_{max}^2 N^7)$. It is also important to note that the proposed simplified algorithm can be applied to any arbitrary beamforming schemes with fixed direction.

IV. TRANSCEIVER DESIGN WITH SUM-RATE MAXIMIZATION

In this section, we consider the sum-rate maximization problem (8) to design transceivers. Here, our goal is to maximize the total transmission rate under a given power budget. In order to satisfy the basic rate requirement of each

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data stream, a set of minimum transmission rates should also be satisfied in problem (8). It is generally not possible to solve the sum-rate maximization problem for its global optimum point in polynomial time. Thus, in this work, we propose suboptimal algorithms to compute a high quality solution for (8). To this end, we first reformulate (8) into an equivalent form, which is easier to handle. Then, we propose an AO-based algorithm to solve the equivalent problem. Finally, simplified algorithms based on the MRT and SIL beamformers are also provided to reduce the computational complexity. The monotonic convergence of the proposed algorithms is guaranteed.

A. Equivalent Reformulation of Problem (8)

As it has been pointed out in Section III-A, the optimal receive beamforming vector for problem (7) and (8) is the MMSE receiver given by (9). Then, by utilizing the WMMSE method [51], we can see that problem (8) is equivalent to the following weighted sum-MSE minimization problem[5]:

$$\min_{\{\mathbf{w}_{s,a,i}\},\{\mathbf{f}_{s,a,i}\},\{\mathbf{d}_i\},\{q_{s,a,i}\}} \sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M} g_{s,a,i} \tag{32a}$$

$$\text{s.t. } R_{s,a,i} \geq \varphi_{s,a,i}, \tag{32b}$$

$$P_i \geq \psi_i, \tag{32c}$$

$$\sum_{s=1}^{M} \mathbf{f}_{s,a,i}^H \mathbf{f}_{s,a,i} \leq P_{a,i}, \tag{32d}$$

$$\sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M} \mathbf{f}_{s,a,i}^H \mathbf{f}_{s,a,i} \leq P_{sum}, \tag{32e}$$

$$0 \leq \rho_{n,i} \leq 1, \quad n \in N_i,\ i \in K,\ s \in M,\ a \in \{1,2\}, \tag{32f}$$

where $g_{s,a,i} = q_{s,a,i}e_{s,a,i} - \log(q_{s,a,i})$, $q_{s,a,i}$ is a weighting factor satisfying $q_{s,a,i} > 0$ and $e_{s,a,i}$ denotes the MSE of the sth data stream of node a in the ith pair, which can be expressed as

$$e_{s,a,i} = |\mathbf{w}_{s,b,i}^H \bar{\mathbf{H}}_{i,i}^{(b,a)} \mathbf{f}_{s,a,i}|^2 + \mathbf{w}_{s,b,i}^H \mathbf{R}_{s,b,i} \mathbf{w}_{s,b,i} - \mathbf{w}_{s,b,i}^H \bar{\mathbf{H}}_{i,i}^{(b,a)} \mathbf{f}_{s,a,i} - \mathbf{f}_{s,a,i}^H \bar{\mathbf{H}}_{i,i}^{(b,a)H} \mathbf{w}_{s,b,i} + 1. \tag{33}$$

By the equivalence between the original problem (8) and the sum-MSE minimization problem (32), we only need to solve the latter. However, problem (32) is still non-convex and highly nontrivial to be solved globally. In the subsequent subsections, we will focus on developing good suboptimal solutions to (32).

B. Proposed Alternating Optimization Design Approach

Since the optimization variables are highly coupled in problem (32), we propose to optimize them successively with the others being fixed. In particular, we optimize problem (32) by sequentially fixing three of the four sets of variables $\{\mathbf{w}_{s,a,i}\}, \{\mathbf{f}_{s,a,i}\}, \{\mathbf{d}_i\}, \{q_{s,a,i}\}$ and updating the remaining set.

[5]Problem (8) and (32) are equivalent in the sense that their global optimal solution are identical.

TABLE IV

AO-Based Algorithm for the Sum-Rate Maximization Problem

1. Initialize $\{\mathbf{f}_{s,a,i}\}$, $\{\mathbf{w}_{s,a,i}\}$ and $\{\mathbf{d}_i\}$. Define the tolerance of accuracy δ_2 and the maximum number of iteration N_{max}^2.
2. **Repeat**
   - 2.1 Update the weight factor $\{q_{s,a,i}\}$ by (34).
   - 2.2 Employ the CCCP method to solve problem (32) with fixed $\{\mathbf{w}_{s,a,i}\}, \{q_{s,a,i}\}, \{\mathbf{d}_i\}$ to obtain the updated $\{\mathbf{f}_{s,a,i}\}$.
   - 2.3 Employ the CCCP method to solve problem (32) with fixed $\{\mathbf{f}_{s,a,i}\}, \{\mathbf{w}_{s,a,i}\}, \{q_{s,a,i}\}$ to obtain the updated $\{\mathbf{d}_i\}$.
   - 2.4 Update the receive beamforming vectors $\{\mathbf{w}_{s,a,i}\}$ by (9).
3. **Until** the sum-rate between two successive iterations is less than δ_2 or the maximum number of iterations is reached.

The update of the weighting factor variables $\{q_{s,a,i}\}$ is given in closed-form by

$$q_{s,a,i} = \frac{1}{e_{s,a,i}}, \tag{34}$$

while the update expression of the receive beamforming vectors $\{\mathbf{w}_{s,a,i}\}$ is given by (9). The update of the transmit beamforming vectors $\{\mathbf{f}_{s,a,i}\}$ admits no closed-form solution. First, we can see that the objective function is convex in the variables $\{\mathbf{f}_{s,a,i}\}$; hence by introducing auxiliary variables $\{t_{s,a,i}\}$ which satisfy

$$q_{s,a,i}e_{s,a,i} - \log(q_{s,a,i}) \leq t_{s,a,i}, \tag{35}$$

the objective function can be further formulated into an second-order cone (SOC) constraint with the new affine objective function $\sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M} t_{s,a,i}$. The SINR constraints (32b) can also be reformulated as SOC constraints by employing the technique proposed in Section III-A. This leaves us the EH constraints (32c) which is non-convex and difficult to handle. However, inspired by the CCCP method we utilized in Section III-A, we can linearize the quantity P_i on the left hand side of (32c) by its first order Taylor expansion. Then (32c) can be further formulated as SOC constraint and the optimization of the transmit beamforming vectors can proceed by solving a sequence of SOCP problems. Finally, the optimization of the PS vectors $\{\mathbf{d}_i\}$ can be similarly addressed as that of the transmit beamforming vectors. The details are omitted.

To summarize, the AO-based algorithm for the sum-rate maximization problem is listed in Table IV. Also, we note that due to the nature of the CCCP and AO approaches, the monotonic convergence of the algorithm is guaranteed. The asymptotic complexity of the proposed algorithm can be expressed as $O(N_{max}^1 N_{max}^2 N^{10})$.

C. Proposed Simplified Algorithm With Fixed Beamformers

Similar to Section III-B, we fix the direction of the beamforming vector based on the MRT and SIL beamformers. Also, we employ the same PS ratio for all the receive antennas of node 2 (i.e., $\rho_{n,i} = \rho_i, \forall n \in \mathcal{N}_i$). Thus, problem (32) can be

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simplified as

$$\begin{aligned}
&\min_{\{p_{s,a,i}\},\{\rho_i\},\{q_{s,a,i}\},\{\mathbf{w}_{s,a,i}\}} \sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M}\widetilde{g}_{s,a,i}\\
&\text{s.t. } \frac{p_{s,2,i}|\mathbf{w}_{s,1,i}^H\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}|^2}{\mathbf{w}_{s,1,i}^H(\mathbf{R}_{s,1,i}^{(1)}+\mathbf{R}_{s,1,i}^{(2)}+\mathbf{R}_{s,1,i}^{(3)}+\sigma_{1,i}^2\mathbf{I})\mathbf{w}_{s,1,i}}\geq\gamma_{s,1,i},\\
&\quad \frac{p_{s,1,i}|\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s,1,i}|^2}{\mathbf{w}_{s,2,i}^H(\mathbf{R}_{s,2,i}^{(1)}+\mathbf{R}_{s,2,i}^{(2)}+\mathbf{R}_{s,2,i}^{(3)}+\sigma_{2,i}^2\mathbf{I}+\frac{\omega_i^2}{\rho_i}\mathbf{I})\mathbf{w}_{s,2,i}}\geq\gamma_{s,2,i},\\
&\quad \left(p_{s,1,i}|\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s,1,i}|^2+\mathrm{Tr}(\mathbf{R}_{s,2,i}^{(1)}+\mathbf{R}_{s,2,i}^{(2)}+\mathbf{R}_{s,2,i}^{(3)})\right)\\
&\quad \geq\frac{\psi_i}{\xi_i(1-\rho_i)}-\sigma_{2,i}^2N_i,\\
&\quad \sum_{s=1}^{M}p_{s,a,i}\widetilde{\mathbf{f}}_{s,a,i}^H\widetilde{\mathbf{f}}_{s,a,i}\leq P_{a,i},\\
&\quad \sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M}p_{s,a,i}\widetilde{\mathbf{f}}_{s,a,i}^H\widetilde{\mathbf{f}}_{s,a,i}\leq P_{sum},\\
&\quad p_{s,1,i}\geq 0,\quad p_{s,2,i}\geq 0,\ 0\leq\rho_i\leq 1,\ i\in K,\ s\in M,
\end{aligned}\tag{36}$$

where $\widetilde{g}_{s,a,i}=q_{s,a,i}\widetilde{e}_{s,a,i}-\log(q_{s,a,i})$. The quantities $\widetilde{e}_{s,1,i}$ and $\widetilde{e}_{s,2,i}$ can be written as

$$\begin{aligned}
\widetilde{e}_{s,1,i}=\ &p_{s,1,i}\rho_i|\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s,1,i}|^2+\mathbf{w}_{s,2,i}^H\widetilde{\mathbf{R}}_{s,2,i}\mathbf{w}_{s,2,i}+1\\
&-\sqrt{p_{s,1,i}\rho_i}\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,1)}\mathbf{f}_{s,1,i}\\
&-\sqrt{p_{s,1,i}\rho_i}\mathbf{f}_{s,1,i}^H\mathbf{H}_{i,i}^{(2,1)H}\mathbf{w}_{s,2,i},
\end{aligned}\tag{37}$$

$$\begin{aligned}
\widetilde{e}_{s,2,i}=\ &p_{s,2,i}|\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}|^2+\mathbf{w}_{s,1,i}^H\widetilde{\mathbf{R}}_{s,1,i}\mathbf{w}_{s,1,i}+1\\
&-\sqrt{p_{s,2,i}}\mathbf{w}_{s,1,i}^H\mathbf{H}_{i,i}^{(1,2)}\mathbf{f}_{s,2,i}\\
&-\sqrt{p_{s,2,i}}\mathbf{f}_{s,2,i}^H\mathbf{H}_{i,i}^{(1,2)H}\mathbf{w}_{s,1,i},
\end{aligned}\tag{38}$$

where

$$\begin{aligned}
\widetilde{\mathbf{R}}_{s,2,i}=\ &(\rho_i\sigma_{2,i}^2+\omega_i^2)\mathbf{I}\\
&+\sum_{s'\neq s}^{M}\rho_i\,p_{s',1,i}\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s',1,i}\widetilde{\mathbf{f}}_{s',1,i}^H\mathbf{H}_{i,i}^{(2,1)H}\\
&+\sum_{s'=1}^{M}\rho_i\,p_{s',2,i}\mathbf{H}_{i,i}^{(2,2)}\widetilde{\mathbf{f}}_{s',2,i}\widetilde{\mathbf{f}}_{s',2,i}^H\mathbf{H}_{i,i}^{(2,2)H}\\
&+\sum_{j\neq i}^{K}\sum_{c=1}^{2}\sum_{s'=1}^{M}\rho_i\,p_{s',c,j}\mathbf{H}_{i,j}^{(2,c)}\widetilde{\mathbf{f}}_{s',c,j}\widetilde{\mathbf{f}}_{s',c,j}^H\mathbf{H}_{i,j}^{(2,c)H},
\end{aligned}\tag{39}$$

$$\begin{aligned}
\widetilde{\mathbf{R}}_{s,1,i}=\ &\sigma_{1,i}^2\mathbf{I}+\sum_{s'\neq s}^{M}p_{s',2,i}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s',2,i}\widetilde{\mathbf{f}}_{s',2,i}^H\mathbf{H}_{i,i}^{(1,2)H}\\
&+\sum_{s'=1}^{M}p_{s',1,i}\mathbf{H}_{i,i}^{(1,1)}\widetilde{\mathbf{f}}_{s',1,i}\widetilde{\mathbf{f}}_{s',1,i}^H\mathbf{H}_{i,i}^{(1,1)H}\\
&+\sum_{j\neq i}^{K}\sum_{c=1}^{2}\sum_{s'=1}^{M}p_{s',c,j}\mathbf{H}_{i,j}^{(1,c)}\widetilde{\mathbf{f}}_{s',c,j}\widetilde{\mathbf{f}}_{s',c,j}^H\mathbf{H}_{i,j}^{(1,c)H}.
\end{aligned}\tag{40}$$

As in the case of problem (32), we propose to solve problem (36) using a sequential approach. It can be observed that the optimization of $\{q_{s,a,i}\}$ and $\{\mathbf{w}_{s,a,i}\}$ admits closed-form solutions. The only remaining problem for the proposed algorithm is how to optimize $\{p_{s,a,i}\}$ and $\{\rho_i\}$. Firstly, we observe that the constraints in (36) and the functions $\{\widetilde{e}_{s,2,i}\}$ are jointly convex over the variables $\{p_{s,a,i}\}$ and $\{\rho_i\}$, which can be easily handled with fixed $\{\mathbf{w}_{s,a,i}\}$. Secondly, we address the non-convex function $\{\widetilde{e}_{s,1,i}\}$ by introducing the auxiliary variables $\{\varrho_{s,i}\}$, $\{\varsigma_{s,i}\}$, $\{\tau_{s,i}\}$ and $\{\upsilon_{s,i}\}$ which satisfy

$$\sum_{s'=1}^{M}p_{s',1,i}\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,1)}\widetilde{\mathbf{f}}_{s',1,i}\widetilde{\mathbf{f}}_{s',1,i}^H\mathbf{H}_{i,i}^{(2,1)H}\mathbf{w}_{s,2,i}\leq\frac{(\varrho_{s,i})^2}{\rho_i},\tag{41}$$

$$\sum_{s'=1}^{M}p_{s',2,i}\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,2)}\widetilde{\mathbf{f}}_{s',2,i}\widetilde{\mathbf{f}}_{s',2,i}^H\mathbf{H}_{i,i}^{(2,2)H}\mathbf{w}_{s,2,i}\leq\frac{(\varsigma_{s,i})^2}{\rho_i},\tag{42}$$

$$\begin{aligned}
&\sum_{j\neq i}^{K}\sum_{c=1}^{2}\sum_{s'=1}^{2}p_{s',c,j}\left(\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,j}^{(2,c)}\widetilde{\mathbf{f}}_{s',c,j}\widetilde{\mathbf{f}}_{s',c,j}^H\mathbf{H}_{i,j}^{(2,c)H}\mathbf{w}_{s,2,i}\right)\\
&\qquad\leq\frac{(\tau_{s,i})^2}{\rho_i},
\end{aligned}\tag{43}$$

$$p_{s,1,i}\geq\frac{(\upsilon_{s,i})^2}{\rho_i}.\tag{44}$$

Notice that the right hand sides of the inequalities (41)-(44) are jointly convex over the variables $\{\rho_i,\,p_{s,1,i}\}$. Then, inspired by the concept of CCCP, we can approximate $\frac{(\varrho_{s,i})^2}{\rho_i}$ in (41) by $\frac{(\varrho_{s,i}^{(k)})^2}{\rho_i^{(k)}}+2\frac{\varrho_{s,i}^{(k)}}{\rho_i^{(k)}}(\varrho_{s,i}-\varrho_{s,i}^{(k)})-\frac{\varrho_{s,i}^{(k)}}{(\rho_i^{(k)})^2}(\rho_i-\rho_i^{(k)})$ in the kth iteration of the proposed algorithm. In cthis way, inequality (41) can be approximated by a convex inequality. Equations (42) and (43) can be handled in a similar way and $\widetilde{e}_{s,1,i}$ can be approximated by

$$\begin{aligned}
\widetilde{e}_{s,1,i}\approx\ &(\varrho_{s,i})^2+(\varsigma_{s,i})^2+(\tau_{s,i})^2+1\\
&+\mathbf{w}_{s,2,i}^H(\rho_i\sigma_{2,i}^2+\omega_i^2)\mathbf{w}_{s,2,i}-\upsilon_{s,i}\mathbf{w}_{s,2,i}^H\mathbf{H}_{i,i}^{(2,1)}\mathbf{f}_{s,1,i}\\
&-\upsilon_{s,i}\mathbf{f}_{s,1,i}^H\mathbf{H}_{i,i}^{(2,1)H}\mathbf{w}_{s,2,i},
\end{aligned}\tag{45}$$

which is a convex function. Finally, we can see that problem (36) with fixed $\{q_{s,a,i},\,\mathbf{w}_{s,a,i}\}$ can be solved iteratively with guaranteed convergence (to a stationary point) [36], where each subproblem can be formulated as an SOCP problem which can be solved efficiently by off-the-shelf algorithms. The details are omitted.

To summarize, the simplified algorithm for the sum-rate maximization problem is listed in Table V. Similar to the previous subsection, we can infer that the proposed algorithm is monotonically convergent and the asymptotic complexity is $O(N_{max}^1N_{max}^2N^7)$.

V. Simulation Results

In this section, we compare the proposed FD joint transceiver design algorithms with the HD AO-based algorithm.[6]

[6]The HD AO-based algorithm can be obtained straightforwardly based on the proposed algorithms in Table II and IV for the FD case.

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TABLE V

SIMPLIFIED ALGORITHM FOR THE SUM-RATE MAXIMIZATION PROBLEM

1. Calculate the transmit beamforming vectors $\{\mathbf{f}_{s,a,i}\}$ by (24) or (25). Initialize the PS ratios $\{\rho_i\}$ and calculate the receive beamforming vector based on (9). Define the tolerance of accuracy δ_2 and the maximum number of iteration N_{max}^2.
2. **Repeat**
   - 2.1 Update the weight factor $\{q_{s,a,i}\}$ by (34).
   - 2.2 Employ the CCCP method to solve problem (36) with fixed $\{\mathbf{w}_{s,a,i}\}, \{q_{s,a,i}\}$ to obtain the updated $\{p_{s,a,i}\}$ and $\{\rho_i\}$.
   - 2.2 Compute (9) with fixed $\{p_{s,a,i}\}$ and $\{\rho_i\}$ to obtain the updated $\{\mathbf{w}_{s,a,i}\}$.
3. **Until** the sum-rate between two successive iterations is less than δ_2 or the maximum number of iterations is reached.

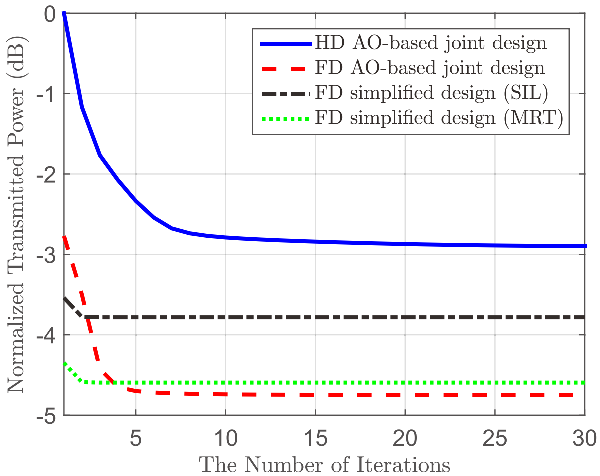

Fig. 2. Convergence behavior of the proposed algorithms for the case of $\varphi = 20/6$bps/Hz and $\psi = 0$dBm.

The nominal system configuration is defined by the following choice of parameters: $T = T_i = 4$, $N = N_i = 4$, $K = 3$, $M = 2$, $\xi_i = 1$, $\sigma_{a,i}^2 = \sigma^2 = -30$dBm, and $\omega_i^2 = \omega^2 = -20$dBm, $\forall i \in \mathcal{K}$.[7] In addition, we assume equal transmission rate and EH targets at the destination receivers, i.e., $\varphi_{s,a,i} = \varphi$, $\psi_i = \psi$, $\forall i \in \mathcal{K}$, $\forall s \in \mathcal{M}$, $\forall a \in \{1, 2\}$, for simplicity. In the implementations of the various algorithms, the tolerance parameters are chosen as $\delta_1 = 1 \times 10^{-4}$, $\delta_2 = 1 \times 10^{-3}$, and the maximum numbers of iterations are chosen as $N_{max}^1 = 20$, $N_{max}^2 = 30$. Furthermore, we assume that the interference channel gain from the nodes in the jth pair to the nodes in the ith pair is given by -10dB, $i \neq j$, and that the residual SI channel gain is given by -20dB. All convex problems are solved by CVX [52] on a desktop Intel (i3-2100) CPU running at 3.1GHz with 4GB RAM. The tolerance level of the solver is set as default.

A. Results for the Sum Power Minimization Problem

In Fig. 2, we illustrate the typical convergence behavior of the proposed algorithms for the case of $\varphi = 20/6$bps/Hz

[7]Similar to [13], [19], we assume that only T_i (N_i) antennas can be used for transmission (reception) in the HD mode, despite the fact that the nodes in the ith pair have $N_i + T_i$ antennas. In other words, the nodes in the ith pair only have T_i transmission front-ends and N_i receiving front-ends. This is because RF front-ends are more expensive than antennas and are therefore considered as scarce resources in practical systems.

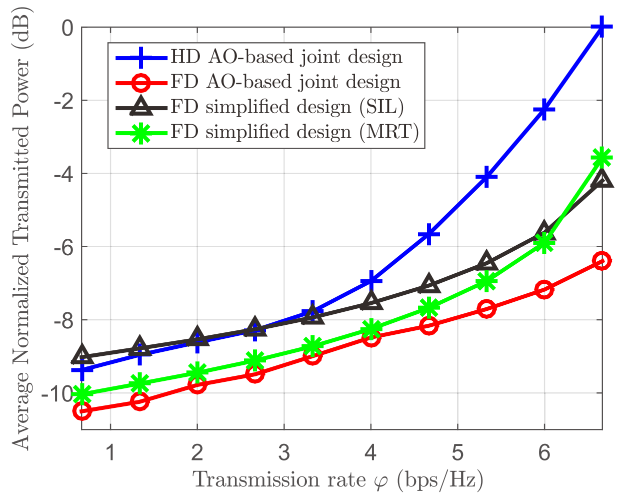

Fig. 3. The average normalized transmission power versus transmission rate φ.

and $\psi = 0$dBm, which is obtained by averaging over 50 independent channel realizations. It can be observed that the proposed algorithms converge in a few steps and they do so monotonically. Specifically, due to the fact that the number of optimization variables of the simplified algorithms is smaller than that of the AO-based algorithm, the former converges faster than the latter. However, the FD AO-based algorithm achieves the best steady-state performance, followed by the MRT-based simplified algorithm, the SIL-based simplified algorithm and the HD AO-based algorithm.

Fig. 3 shows the average power consumption versus transmission rate for the analyzed sum power minimization problem, where we fix $\psi = 0$dBm. We can see from the figure that the performance of the proposed FD transceiver design algorithms is much better than that of the HD design algorithm, and that the gain becomes larger for higher transmission rate. In particular, compared with the HD case, the AO-based joint transceiver design algorithm can lead to a power saving of almost 5dB when the transmission rate φ is set as 6bps/Hz. The power consumption of the simplified algorithms with the MRT and SIL beamformer is still 3dB less than that of the HD scheme. It is important to mention that the performance of the MRT-based simplified algorithm is better than that of the SIL-based simplified algorithm in the lower transmission rate region since the interference is beneficial when the rate requirement is not too high. However, in the high transmission rate region, large interference can limit the rate performance and thus the SIL-based algorithm, which takes the interference into account, can perform better than the MRT-based algorithm.

Fig. 4 shows the average power consumption versus node 2 EH target ψ for the FD and HD joint transceiver design algorithms, where the transmission rate is given by $\varphi = 20/6$ bps/Hz.[8] From the results, we can see that the proposed FD AO-based joint transceiver design algorithm outperforms its HD counterpart. In particular, the

[8]Since we choose $K = 3$ and $M = 2$ in the system configuration, thus we choose $\varphi = 20/6$ bps/Hz in this simulation.

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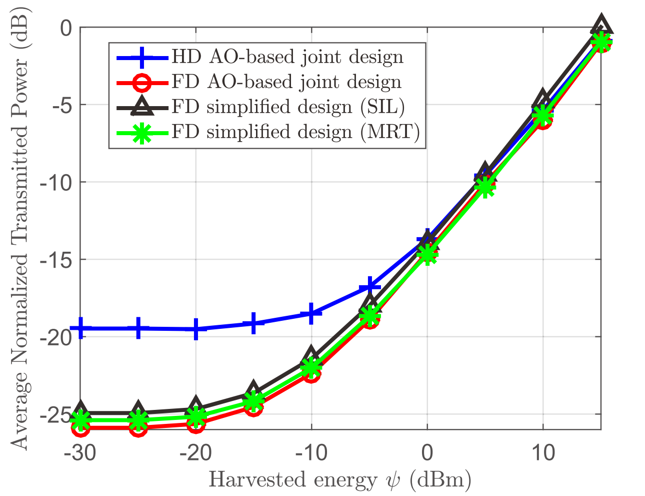

Fig. 4. The average normalized transmission power versus harvested power ψ.

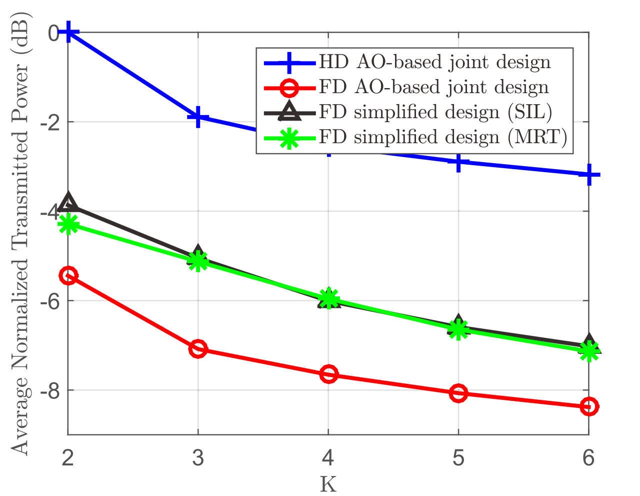

Fig. 5. The average normalized transmission power versus the number of pairs K ($N = T = K + 1$).

proposed design algorithm can save 5dB in transmitted power for $\psi = -15$dBm, compared to the HD case. The performance of the simplified algorithms is very close to the AO-based algorithm in this case. Furthermore, the performance of the FD and HD schemes becomes closer while increasing the EH requirement. This can be intuitively explained by noting that in order to support large harvested energy, the transmission power of node 1 in a given pair must be much larger than that of its corresponding node 2, and therefore the advantage of the FD scheme would vanish in this case.

Fig. 5 illustrates the impact of N, T and K, on the minimum transmission power for all proposed solutions with $\varphi = 2$bps/Hz and $\psi = 0$dBm. In order to make the problem feasible (i.e., reserve enough degrees of freedom), we set $N = T = K+1$. It is observed that when N, T and K increase in this way, the total transmission power is substantially decreased for all solutions. One can infer that if we fix the number of pairs and increase the number of antennas, the total transmission power can be further decreased since additional multiplexing gain would be available. This demonstrates the significant benefit of using large antenna arrays to

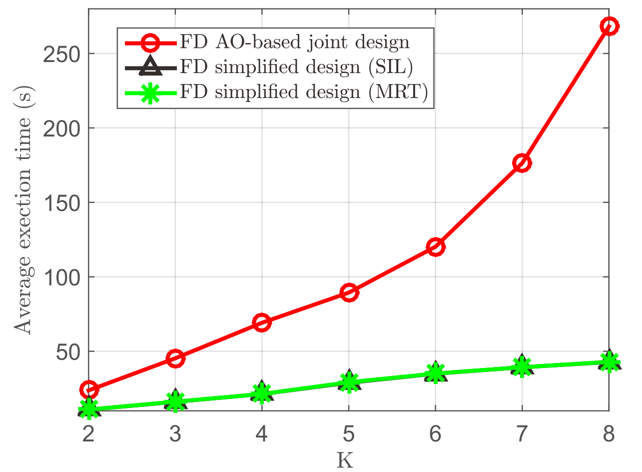

Fig. 6. The average execution time versus the number of pairs K ($N = T = K + 1$) for the sum power minimization problem (7).

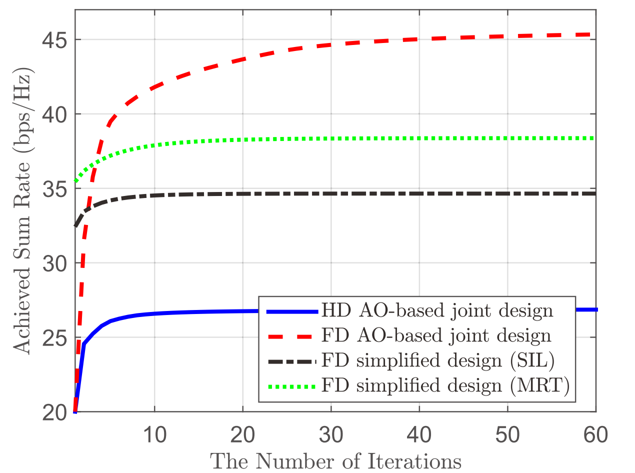

Fig. 7. Convergence behavior of the proposed algorithms for the case of $\varphi = 2$bps/Hz, $\psi = 0$dBm and $P_{sum} = -2$dBm.

efficiently implement FD MIMO interference SWIPT systems in practice.

Finally, in Fig. 6, we compare the the average execution time of the AO-based algorithm and the simplified algorithms over 20 channel realizations under the same simulation setup as that in Fig. 5. It can be observed that the time consumed by the simplified algorithms is much less than that of the AO-based algorithm, especially as the number of user pairs K increases.

B. Results for the Sum-Rate Maximization Problem

In this subsection, we illustrate the performance of the proposed algorithms for the sum-rate maximization problem. Firstly, in Fig. 7, we illustrate the convergence behavior of the proposed algorithms for the case of $\varphi = 2$bps/Hz, $\psi = 0$dBm and $P_{sum} = -2$dBm, which is obtained by averaging over 50 independent channel realizations. We can see from the figure that the proposed algorithms converge in a few tens of steps and they do so monotonically. In particular, the simplified algorithms converge faster than the AO-based algorithm due to their reduced number of optimization variables. However, the FD AO-based algorithm achieves the best steady-state

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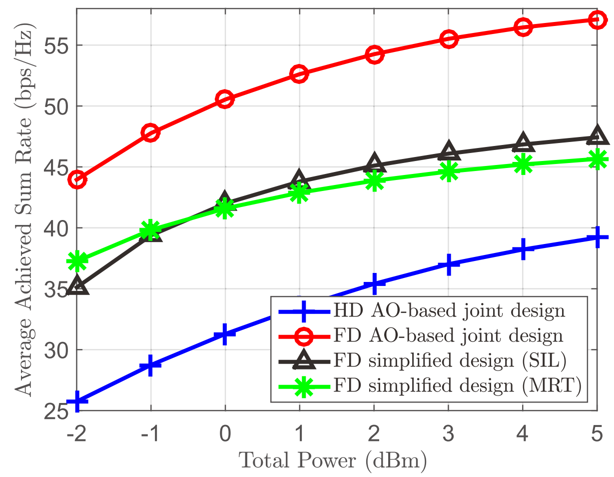

Fig. 8. The average achieved sum-rate versus the total transmission power P_{sum}.

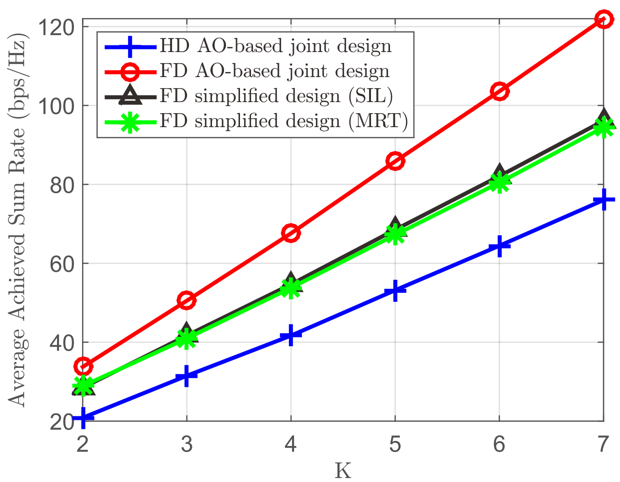

Fig. 9. The average achieved sum-rate versus the number of pairs K $(N = T = K + 1)$.

performance, followed by the MRT-based simplified algorithm, the SIL-based simplified algorithm and the HD AO-based algorithm.

Next, we show in Fig. 8 the maximum sum-rate achieved by the FD and HD schemes versus the total transmission power budget with $\varphi = 20/6$bps/Hz and $\psi = 0$dBm. We assume $P_{a,i} = P = 3$dBm for all $\{a, i\}$. Similar to Fig. 3, it is observed that the performance of the FD scheme is much better than that of the HD scheme. In particular, the achieved sum-rate of the AO-based and the simplified algorithms for the FD case is 20bps/Hz and 10bps/Hz higher than for the HD scheme, respectively. In the high sum power region, the performance of the SIL-based simplified algorithm is better than that of the MRT-based one since more power is available for transmission such that the interference is relatively severe and the SIL-based algorithm can better mitigate the excessive interference.

We show in Fig. 9 the influence of N, T and K on the maximum achieved sum-rate for all proposed algorithms with $\varphi = 2$bps/Hz, $\psi = 0$dBm and $P_{sum} = 0$dBm. Similar to Fig. 5, we set $N = T = K + 1$. It can be observed that when N, T and K increase, the achieved sum-rate is substantially

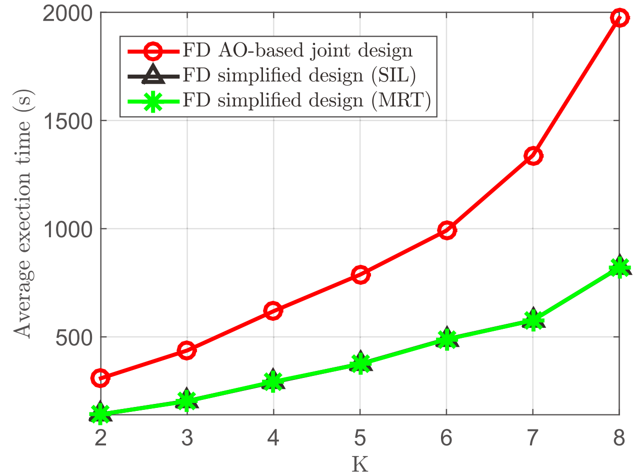

Fig. 10. The average execution time versus the number of pairs K $(N = T = K + 1)$ for the sum-rate maximization problem (8).

increased for all solutions. Again, the performance of the FD scheme is much better than that of the HD scheme. The two simplified algorithms exhibit similar performance.

Finally, in Fig. 10, we compare the average execution time of the AO-based algorithm and the simplified algorithms over 20 channel realizations. As we can see, the simplified algorithms are more time efficient than the AO-based algorithm, which is similar to the results in Fig. 6. However, in comparison to the results presented in Fig. 6 for the sum power minimization problem, the run times for the solutions of the sum-rate maximization problem significantly exceed those of the former problem. This can be explained by the fact that the subproblems in the sum-rate maximization problem are more complex that of the sum power minimization problem and thus more computational time is needed.

VI. Conclusion

In this work, we proposed joint transceiver design algorithms in an FD K-pair MIMO interference channel with SWIPT. Two important utility optimization problems were considered, i.e. the sum power minimization problem and the sum-rate maximization problem. In each case, AO-based iterative algorithms were proposed to successively optimize the transceiver coefficients, along with simplified algorithms based on the MRT and SIL beamformers. In each subproblem, we show that each group of the variables could be optimized by employing the CCCP method, the SDR technique or closed-form solutions. Simulation results were presented to verify that the proposed FD joint transceiver design algorithms outperform the corresponding HD algorithm. The computational advantages of the simplified algorithms were also demonstrated.

Appendix A
The Proof of Lemma 1

Suppose that $\{\mathbf{F}^*_{s,a,i}\}$ are the optimal solutions to (10) and $\text{Rank}(\mathbf{F}^*_{s,a,i}) \geq 1$, then according to [53], we have the

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following inequality about the rank property of (10):

$$2MK \leq \sum_{i=1}^{K}\sum_{a=1}^{2}\sum_{s=1}^{M}(\text{Rank}(\mathbf{F}_{s,a,i}^{*}))^2 \leq 2MK + K. \quad (46)$$

When $K = 2$, inequality (46) becomes

$$4M \leq \sum_{i=1}^{2}\sum_{a=1}^{2}\sum_{s=1}^{M}(\text{Rank}(\mathbf{F}_{s,a,i}^{*}))^2 \leq 4M + 2, \quad (47)$$

from which we can infer that $\text{Rank}(\mathbf{F}_{s,a,i}) = 1, \forall s, a, i$. If there exists one set of indices $\{\tilde{s}, \tilde{a}, \tilde{i}\}$ such that $\text{Rank}(\mathbf{F}_{\tilde{s},\tilde{a},\tilde{i}}) = 2$, then inequality (46) would become

$$4M \leq 4M - 1 + 4 \leq 4M + 2, \quad (48)$$

which is impossible. This completes the proof.

Appendix B
The Proof of Lemma 2

Firstly, we prove that (30) holds true for all $\{s, a, i\} \in \mathcal{L}$. Since $\{\mathbf{w}_{s,a,i}, p_{s,a,i}, \rho_i\}$ satisfy the SINR constraints of problem (26) with equality for all $\{s, a, i\} \in \mathcal{L}$, we have

$$\mathbf{w}_{s,1,i}^{H}\left(\frac{p_{s,2,i}}{\gamma_{s,1,i}}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}\widetilde{\mathbf{f}}_{s,2,i}^{H}\mathbf{H}_{i,i}^{(1,2)H} - \mathbf{R}_{s,1,i}\right)\mathbf{w}_{s,1,i} = 0. \quad (49)$$

By substituting the maximum SINR receiver (9) into (49), we have

$$p_{s,2,i}\widetilde{\mathbf{f}}_{s,2,i}^{H}\mathbf{H}_{i,i}^{(1,2)H}\mathbf{R}_{s,1,i}^{-1}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}\left(\frac{p_{s,2,i}}{\gamma_{s,1,i}}\widetilde{\mathbf{f}}_{s,2,i}^{H}\mathbf{H}_{i,i}^{(1,2)H}\mathbf{R}_{s,1,i}^{-1}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i} - 1\right) = 0 \quad (50)$$

Due to the fact that $\mathbf{R}_{s,1,i}$ is positive definite and $\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i} \neq \mathbf{0}$, we can infer that

$$\frac{p_{s,2,i}}{\gamma_{s,1,i}}\widetilde{\mathbf{f}}_{s,2,i}^{H}\mathbf{H}_{i,i}^{(1,2)H}\mathbf{R}_{s,1,i}^{-1}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i} = 1. \quad (51)$$

By employing the Schur complement [54] twice, it follows that $\mathbf{R}_{s,1,i} - \frac{p_{s,2,i}}{\gamma_{s,1,i}}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}\widetilde{\mathbf{f}}_{s,2,i}^{H}\mathbf{H}_{i,i}^{(1,2)H}$ is positive semidefinite. Therefore, by combining (49) we have the following expression for all $\{s, a, i\} \in \mathcal{L}$

$$\left(\mathbf{R}_{s,1,i} - \frac{p_{s,2,i}}{\gamma_{s,1,i}}\mathbf{H}_{i,i}^{(1,2)}\widetilde{\mathbf{f}}_{s,2,i}\widetilde{\mathbf{f}}_{s,2,i}^{H}\mathbf{H}_{i,i}^{(1,2)H}\right)^{\frac{1}{2}}\mathbf{w}_{s,1,i} = \mathbf{0}. \quad (52)$$

Similarly, we can prove (31) is true for all $\{s, a, i\} \in \mathcal{L}$. This completes the proof.

Appendix C
The Proof of Lemma 3

To proceed, we first define a compact form of the variables $\mathbf{F} = \{p_{s,a,i}, \rho_i\}_{i\in\mathcal{K},s\in\mathcal{M},a=\{1,2\}}$, $\mathbf{W} = \{\mathbf{w}_{s,a,i}\}_{i\in\mathcal{K},s\in\mathcal{M},a=\{1,2\}}$ and let $\text{SINR}_{s,a,i}(\mathbf{F}, \mathbf{W})$, $\text{EH}_i(\mathbf{F})$ denote the corresponding SINR and EH constraints in problem (26). These expressions will be used to make the dependency of the constraints on the design parameters explicit. Then, we define the following mappings:

$$\Theta_l(\mathbf{F}) \overset{\Delta}{=} \mathbf{w}_{s,a,i} = p_{s,b,i}\mathbf{R}_{s,a,i}^{-1}\bar{\mathbf{H}}_{i,i}^{(a,b)}\widetilde{\mathbf{f}}_{s,b,i}, \quad (53)$$

$$\begin{aligned}\mathbf{F} \in \Psi(\mathbf{W}) \Leftrightarrow \mathbf{F} \in \arg\min_{\mathbf{F}} &\sum_{s=1}^{M}\sum_{a=1}^{2}\sum_{i=1}^{K} p_{s,a,i}\|\widetilde{\mathbf{f}}_{s,a,i}\|^2\\ \text{s.t. } &\text{SINR}_{s,a,i}(\mathbf{F}, \mathbf{W}) \geq \gamma_{s,a,i},\\ &\text{EH}_i(\mathbf{F}) \geq \psi_k, \quad 0 \leq \rho_k \leq 1,\\ &i \in \mathcal{K}, \ s \in \mathcal{M}, \ a = \{1,2\}. \quad (54)\end{aligned}$$

Let $\Theta(\mathbf{F}) = \{\Theta_l(\mathbf{F})\}_{l=1}^{L}$ and let $\mathcal{F}(\mathbf{W})$ denote the feasible set for $\mathbf{F}$. Also, let $P\{\mathbf{F}, \mathbf{W}\}$ denote the value of the objective function of problem (26). We observe that the sequence $P\{\mathbf{F}^n, \mathbf{W}^n\}_{n=1}^{\infty}$ is bounded below and monotonically decreases, which also converges with n as the iteration number. Denote its limit as P^*. Since the objective function is coercive and $\{\mathbf{F}^n\}$ is bounded. Hence, the sequence $\{\mathbf{F}^n\}$ has at least one limit point $\{\mathbf{F}^*\}$. Let $\{\mathbf{F}^{n_j}, \}_{j=1}^{\infty}$ be the subsequence converging to $\{\mathbf{F}^*\}$. Since the mapping $\Theta(\cdot)$ is continuous, we must have

$$\lim_{j\to\infty}(\mathbf{F}^{n_j}, \mathbf{W}^{n_j}) = (\mathbf{F}^*, \mathbf{W}^*) \overset{\Delta}{=} (\mathbf{F}^*, \Theta(\mathbf{F}^*)). \quad (55)$$

First it can be shown that in the limit $\mathbf{F}^* \in \Psi(\mathbf{W}^*)$ holds. Next, we establish that $(\mathbf{F}^*, \mathbf{W}^*) = (\mathbf{F}^*, \Theta(\mathbf{F}^*))$ is a KKT solution of (26). This is mainly due to the fact that $\mathbf{F}^* \in \Psi(\mathbf{W}^*)$, thus $\mathbf{F}^*$ is a global optimal solution of the following problem:

$$\begin{aligned}\min_{\mathbf{F}} \ &\sum_{s=1}^{M}\sum_{a=1}^{2}\sum_{i=1}^{K} p_{s,a,i}\|\widetilde{\mathbf{f}}_{s,a,i}\|^2\\ \text{s.t. } &\text{SINR}_{s,a,i}(\mathbf{F}, \mathbf{W}^*) \geq \gamma_{s,a,i}, \ \text{EH}_i(\mathbf{F}) \geq \psi_k,\\ &0 \leq \rho_k \leq 1, i \in \mathcal{K}, s \in \mathcal{M}, a = \{1,2\}. \quad (56)\end{aligned}$$

Therefore, $\mathbf{F}^*$ must satisfy the optimality conditions of (56) with $\{\lambda_{s,a,i}^*\}$, $\{\mu_i^*\}$ as the associated Lagrangian multipliers of the SINR and EH constraints respectively. Similarly, due to the fact that $\mathbf{W}^* = \Theta(\mathbf{F}^*)$, we have that $\mathbf{W}^*$ must satisfy the first-order optimality conditions since if $\{s, a, i\} \in \mathcal{L}$, we have (30) and (31) according to Lemma 2 and if $\{s, a, i\} \notin \mathcal{L}$, $\lambda_{s,a,i}^* = 0$ must be true. Furthermore, the primal feasible conditions, the dual feasible conditions and the complementarity conditions must be satisfied due to the fact that $\mathbf{F}^* \in \Psi(\mathbf{W}^*)$.

In conclusion, we have that $(\mathbf{F}^*, \mathbf{W}^*) = (\mathbf{F}^*, \Theta(\mathbf{F}^*))$ along with the multipliers $\{\lambda_{s,a,i}^*\}$, $\{\mu_i^*\}$ satisfy the KKT conditions for problem (26). This result implies that $(\mathbf{F}^*, \mathbf{W}^*) = (\mathbf{F}^*, \Theta(\mathbf{F}^*))$ is a KKT solution to problem (26). This completes the proof.

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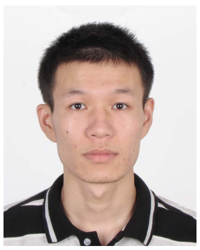

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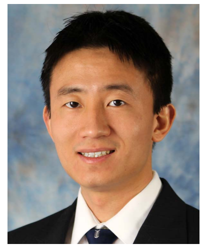

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