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Weighted Sum Transmit Power Minimization for Full-Duplex System With SWIPT and Self-Energy Recycling

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ABSTRACT This paper considers a full-duplex (FD) point-to-point system consisting of one multi-antenna FD access point (FD-AP) and one two-antenna FD mobile station (FD-MS). We adopt simultaneous wireless information and power transfer scheme and apply the self-energy recycling at FD-MS. In order to minimize the weighted sum transmit power, we jointly design the transmit beamforming vector of FD-AP, the receive power splitting ratio of FD-MS, and the transmit power value of FD-MS. Since the original problem is non-convex, we apply semidefinite relaxation and obtain a new convex problem. We further prove that both problems have exactly the same solutions. Finally, simulations are provided to verify our analysis, and the comparison with a half-duplex system demonstrates the significant performance gain from self-energy recycling.

INDEX TERMS Full-duplex, SWIPT, self-energy recycling, SDR, optimality.

I. INTRODUCTION

A promising technology called simultaneous wireless information and power transfer (SWIPT) has gained lots of attention since it can alleviate the energy-bottleneck of energy-constrained wireless networks without interrupting the data transmission [1]. Meanwhile, SWIPT is known as an alternative to conventional energy harvesting techniques which rely on the use of radio frequency signals and are expected to bring fundamental changes to the design of wireless communication networks [2].

In [3], SWIPT with a single-input single-output (SISO) additive white Gaussian noise (AWGN) channel was studied from an information-theoretic point. Later, [3] was extended to frequency-selective channels with AWGN and a SISO fading channel under co-channel interference in [4] and [5], respectively. However, a key drawback of [3]–[5] is that they assume the existing receiver circuits can decode information and harvest energy from the same received signal independently, which is not easy to implement in practice [1], [6]. As a result, two signal separation schemes, called time switching (TS) and power splitting (PS), were proposed in [6]. Based on [6], the authors of [1] and [7] investigated the joint transmit beamforming and receive PS ratio to minimize the transmit power at BS in a multiuser multiple-input single-output (MISO) broadcast system. Meanwhile, [8] revealed some fundamental tradeoffs when maximizing the efficiency in a three-node multiple-input multiple-output (MIMO) broadcasting system with SWIPT.

Recently, an interesting combination of SWIPT and full-duplex (FD) relay system was proposed in [9]–[11]. In [9], three different communication modes, as well as the corresponding throughput were presented for a dual-hop FD relaying system. Based on the work in [9], the authors further derived optimal and suboptimal solutions for the throughput maximization problem in the MIMO FD relay system. In [11], the authors proposed a novel approach to harvest energy at a full-duplex relay using the energy from the source and from itself [11].

Motivated by [11], we consider an FD point-to-point system consisting of a multi-antenna full-duplex access point (FD-AP) and a two-antenna full-duplex mobile station (FD-MS), where FD-MS applies PS scheme to receive both information and energy from FD-AP continuously at all time. In the meantime, part of the self-energy caused by the loop channel at FD-MS can be harvested by FD-MS itself, which is also known as self-energy recycling [11]. By this way, the harmful self-interference becomes beneficial since it provides an extra energy for FD-MS in addition to the energy sent from FD-AP. To our best knowledge, the proposed MIMO FD system with SWIPT and self-recycling is new and has not appeared in literature.

Under the SNR and the harvested energy constraints, we formulate an optimization problem to minimize the weighted sum transmit power of the FD system. The original problem is unfortunately non-convex and we then apply semidefinite relaxation (SDR) [12] to convert it into a convex problem. Since the SDR problem may not be equivalent to the original problem, we further prove the optimality of the SDR by showing the existence of rank-one solutions. Finally, Numerical results are provided to validate our analysis. Comparing to a half-duplex (HD) system without self-energy recycling, the proposed scheme could significantly reduce the weighted sum transmit power.

The rest of the paper is organized as follows. Section II presents our system model and the formulation of the optimization problem. Section III derives the solution via SDR and proves its optimality. The simulation results are provided in Section IV. Finally, the conclusions are made in Section V.

Notations: Vectors and matrices are boldface small and capital letters, respectively. The Hermitian, transpose, trace and rank of $\mathbf{A}$ are denoted by $\mathbf{A}^H$, $\mathbf{A}^T$, Tr($\mathbf{A}$) and Rank($\mathbf{A}$), respectively; $\mathbf{I}_n$ represents an $n \times n$ identity matrix; $\mathbf{A} \succeq 0$ and $\mathbf{A} \succ 0$ mean that $\mathbf{A}$ is positive semi-definite and positive definite, respectively; $\|\cdot\|$ denotes the Euclidean norm of a complex vector, while $|\cdot|$ denotes the absolute value of a complex scalar; The distribution of a circularly symmetric complex Gaussian (CSCG) random variable with zero mean and variance σ^2 is defined as $\mathcal{CN}(0, \sigma^2)$, and $\sim$ stands for 'distributed as'; Finally, $\mathbb{C}^{m\times n}$ denotes the space of $m \times n$ complex matrices.

II. SYSTEM MODEL AND PROBLEM FORMULATION

The considered system is shown in Fig. 1, where FD-MS uses the energy harvested from FD-AP and its own transmitted signal to communicate with FD-AP in a bidirectional way [13]. The proposed system involves two-way information flow between the FD-AP and the FD-MS, as well as one-way energy flow from the FD-AP to FD-MS. Specifically, FD-AP transmits the signal to FD-MS while receiving the signal from FD-MS simultaneously. FD-MS adopts PS scheme to split the signal from FD-AP into two signal streams. One is used for decoding information while the other is used for harvesting energy.

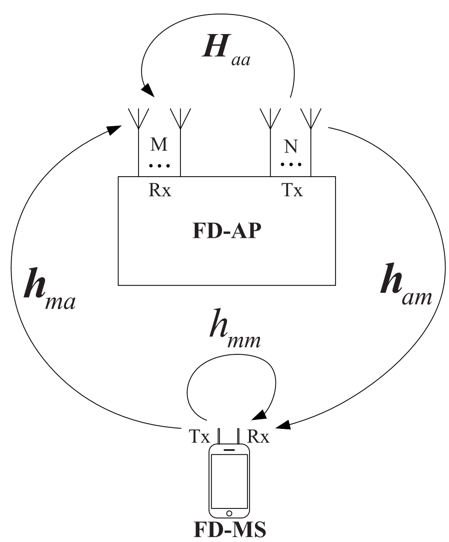

FIGURE 1. An FD point-to-point system with SWIPT, where FD-AP and FD-MS are equipped with $M + N$ and two antennas, respectively. The downlink and uplink transmissions are simultaneous.

Meanwhile, FD-AP is equipped with $M + N$ antennas, where M and N RF chains are used for receiving and transmitting information, respectively. Moreover, FD-MS is equipped with two antennas, of which one is used for both energy and information reception while the other is used for information transmission. All channels are assumed to be quasi-static flat-fading. The uplink channel from FD-MS to FD-AP is denoted as $\mathbf{h}_{ma} \in \mathbb{C}^{M\times 1}$, while the downlink channel from FD-AP to FD-MS is denoted as $\mathbf{h}_{am} \in \mathbb{C}^{N\times 1}$. The loop channels at FD-AP and FD-MS are defined as $\mathbf{H}_{aa} \in \mathbb{C}^{M\times N}$ and $h_{mm} \in \mathbb{C}^{1\times 1}$, respectively. In addition, the duration of one block is normalized for simplicity in this paper. Note that the acquisition of channel state information (CSI) in a bidirectional full-duplex (BFD) system [14], [15], which has already been studied in [14], is a separate topic from the discussed beamforming scheme. Therefore, as did in [11], we assume that CSI has already been obtained.

The uplink signal received by FD-AP is expressed as

$$\mathbf{r}_{ap} = \sqrt{p_{ms}}\mathbf{h}_{ma}v + \mathbf{H}_{aa}\boldsymbol{w}s + \mathbf{n}_1 \tag{1}$$

where $v \sim \mathcal{CN}(0, 1)$ is the transmit symbol at FD-MS, p_{ms} is the transmit power of FD-MS, and $\mathbf{n}_1 \sim \mathcal{CN}(0, \sigma_1^2\mathbf{I}_M)$ is the antenna noise at the receiver of FD-AP. Note that the second term on the right hand side (RHS) of (1) denotes the self-interference due to the loop channel. Moreover,

$$\mathbf{x} = \boldsymbol{w} \cdot s \tag{2}$$

is the transmit signal vector at FD-AP, where $s \sim \mathcal{CN}(0, 1)$ is the symbol from FD-AP and $\boldsymbol{w} \in \mathbb{C}^{N\times 1}$ stands for the transmit beamforming vector.

Note that FD-AP and FD-MS both perfectly know the CSI of the loop channels and the signals transmitted by themselves. Therefore, like some other FD works [16], [17], we assume perfect self-interference cancellation are implemented at each terminal by using digital cancellation directly.

Then, the achievable signal-to-noise ratio (SNR) at FD-AP after self-interference cancellation is given by

$$\mathrm{SNR}_{ap} = \frac{p_{ms}\|\mathbf{h}_{ma}\|^2}{\sigma_1^2}. \tag{3}$$

On the other hand, the downlink signal received by FD-MS is expressed as

$$r_{ms} = \mathbf{h}_{am}^H \boldsymbol{w} s + \sqrt{p_{ms}} h_{mm} v + n_2 \tag{4}$$

where $n_2 \sim \mathcal{CN}(0, \sigma_2^2)$ denotes the antenna noise at the receiver of FD-MS.

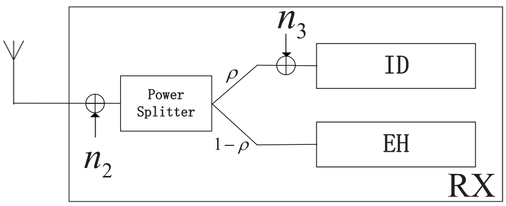

FIGURE 2. **The receiver of FD-MS applies PS scheme to decode information and harvest energy simultaneously.**

As shown in Fig. 2, FD-MS applies PS scheme to make a tradeoff between information decoding and energy harvesting [6]. Specifically, r_{ms} is split into two streams r_{ms}^{ID} and r_{ms}^{EH} by a power splitter [1], [2], where the former goes to the information decoder (ID) while the latter goes to the energy harvester (EH). Furthermore, the PS ratios for r_{ms}^{ID} and r_{ms}^{EH} are denoted as ρ and $1-\rho$, respectively. Consequently, the expressions of r_{ms}^{ID} and r_{ms}^{EH} are given by

$$\begin{aligned} r_{ms}^{ID} &= \sqrt{\rho} r_{ms} + n_3 \\ &= \sqrt{\rho}\mathbf{h}_{am}^H \boldsymbol{w} s + \sqrt{\rho p_{ms}} h_{mm} v + \sqrt{\rho} n_2 + n_3, \end{aligned} \tag{5}$$

$$\begin{aligned} r_{ms}^{EH} &= \sqrt{1-\rho} r_{ms} \\ &= \sqrt{1-\rho}\left(\mathbf{h}_{am}^H \boldsymbol{w} s + \sqrt{p_{ms}} h_{mm} v + n_2\right), \end{aligned} \tag{6}$$

where $n_3 \sim \mathcal{CN}(0, \sigma_3^2)$ is the additional noise introduced by ID at FD-MS [1].

With perfect self-interference cancellation, the remaining signal at ID is expressed as

$$r_{ms}^{ID-SIC} = \sqrt{\rho}\mathbf{h}_{am}^H \boldsymbol{w} s + \sqrt{\rho} n_2 + n_3. \tag{7}$$

Accordingly, the SNR of ID at FD-MS is given by

$$\mathrm{SNR}_{ms} = \frac{\rho|\mathbf{h}_{am}^H \boldsymbol{w}|^2}{\rho\sigma_2^2 + \sigma_3^2}. \tag{8}$$

For EH at FD-MS, we may ignore the power of n_2 in (6) [8]. Then the energy harvested by FD-MS can be expressed as

$$Q = \eta(1-\rho)\left(|\mathbf{h}_{am}^H \boldsymbol{w}|^2 + p_{ms}|h_{mm}|^2\right), \tag{9}$$

where $\eta \in (0, 1)$ denotes the energy conversion efficiency at EH.

In order to formulate a more general optimization problem, weighted sum transmit power of the system is considered to be the objective function. Besides, the purpose of including the transmit power of FD-AP in the objective function is to investigate how much power the self-recycling can save for the system, as did in [1]. To guarantee a continuous information exchange between FD-AP and FD-MS, the SNR at FD-AP and FD-MS should be higher than given thresholds, denoted by γ_{ap} and γ_{ms}, respectively. Note that SNR is an equivalent form of throughput since the block time is normalized. Meanwhile, the energy harvested by MS should be greater than a given target, denoted by ε, in order to prolong the lifetime of FD-MS. In addition, the transmit power of FD-MS should not be greater than the harvested energy threshold to keep FD-MS working persistently. With above considerations, the primal problem to minimize the sum weighted transmit power of the system is formulated as follows:

$$\begin{aligned} &P1: \\ &\min_{\boldsymbol{w},\, p_{ms},\, \rho} \quad c\cdot\|\boldsymbol{w}\|^2 + (1-c)\cdot p_{ms} \\ &\text{s.t.} \quad \frac{p_{ms}\|\mathbf{h}_{ma}\|^2}{\sigma_1^2} \geq \gamma_{ap}, && (10) \\ &\qquad \frac{\rho|\mathbf{h}_{am}^H \boldsymbol{w}|^2}{\rho\sigma_2^2 + \sigma_3^2} \geq \gamma_{ms}, && (11) \\ &\eta(1-\rho)\left(|\mathbf{h}_{am}^H \boldsymbol{w}|^2 + p_{ms}|h_{mm}|^2\right) \geq \varepsilon, && (12) \\ &\qquad 0 < p_{ms} \leq \varepsilon, && (13) \\ &\qquad 0 < \rho < 1, && (14) \end{aligned}$$

where $c \in (0, 1)$ and $1-c$ is defined as the weight factor of the transmit power of FD-AP and FD-MS, respectively.

Apparently, problem **P1** is non-convex due to the coupled beamforming vector $\boldsymbol{w}$ and PS ratio ρ, as well as the quadratic terms involving $\boldsymbol{w}$ in (11) and (12). Moreover, it is briefly discussed in Appendix A that problem **P1** is always feasible.

III. OPTIMAL SOLUTION

Denote $\mathbf{W} = \boldsymbol{w}\boldsymbol{w}^H \in \mathbb{C}^{N\times N}$ and $\mathbf{H}_{am} = \mathbf{h}_{am}\mathbf{h}_{am}^H \in \mathbb{C}^{N\times N}$, both being rank-one matrices. Since the rank-one constraint is non-convex and difficult to handle, we apply SDR to convert **P1** to a new problem **P1-SDR** by dropping the rank constraint:

$$\begin{aligned} &\mathbf{P1-SDR}: \\ &\min_{\mathbf{W},\, p_{ms},\, \rho} \quad c\cdot\mathrm{Tr}(\mathbf{W}) + (1-c)\cdot_{ms} \\ &\text{s.t.} \quad \frac{1}{\gamma_{ms}}\mathrm{Tr}(\mathbf{H}_{am}\mathbf{W}) \geq \sigma_2^2 + \frac{\sigma_3^2}{\rho}, && (15) \\ &\mathrm{Tr}(\mathbf{H}_{am}\mathbf{W}) + p_{ms}|h_{mm}|^2 \geq \frac{\varepsilon}{\eta(1-\rho)}, && (16) \\ &\qquad \frac{\sigma_1^2\gamma_{ap}}{\|\mathbf{h}_{ma}\|^2} \leq p_{ms} \leq \varepsilon, && (17) \\ &\qquad 0 < \rho < 1, && (18) \\ &\qquad \mathbf{W} \succeq \mathbf{0}. && (19) \end{aligned}$$

Obviously, **P1-SDR** is a convex problem that can be solved by the existing optimization software, e.g., CVX tools. Note that the complexity of the interior-point algorithm for solving **P1-SDR** is approximately $O(N^{2.5} + N^{3.5})$ according to [1] and [7]. Moreover, it is not difficult to know that the optimal value of **P1-SDR** provides a lower bound for that of **P1**. Nevertheless, if the optimal $\mathbf{W}$ for **P1-SDR**, denoted by $\mathbf{W}^*$, satisfies $\mathrm{Rank}(\mathbf{W}^*) = 1$, then **P1** and **P1-SDR** will be equivalent and the optimal solution of **P1** can be obtained from that of **P1-SDR**. Specifically, the optimal beamforming vector $\boldsymbol{w}^*$ for **P1** can be obtained from the eigenvector of $\mathbf{W}^*$ that corresponds to the non-zero eigenvalue. The optimal transmit power of FD-MS p_{ms}^* and PS ratio ρ^* of **P1** are the same as those of **P1-SDR**. In practice, since the problem is assumed to be solved at FD-AP, an extra control channel is needed to deliver the separated outputs and inputs of the above problem from one side to the other side.

Indeed, the optimality of the solution of **P1-SDR** can be proved by the following proposition.

Proposition 1: For problem **P1-SDR**, there are:

1) The constraints (15) and (16) in problem **P1-SDR** are satisfied with equality at the optimal point;

2) $\mathbf{W}^*$ satisfies $\mathrm{Rank}(\mathbf{W}^*) = 1$.

Proof: Since problem **P1-SDR** is convex and satisfies Slater's condition, its duality gap is zero [18]. Denote $\lambda \geq 0$, $\mu \geq 0$, $\alpha \geq 0$, $\beta \geq 0$ and $\mathbf{\Phi} \succeq 0$ as the dual variables corresponding to the constraints (15), (16), (17) and (19), respectively. The Lagrangian function of **P1-SDR** is expressed as

$$\begin{aligned}
&\mathcal{L}(\mathbf{W}, p_{ms}, \rho, \lambda, \mu, \alpha, \beta, \Phi) \\
&= c \cdot \mathrm{Tr}(\mathbf{W}) + (1-c) \cdot p_{ms} \\
&\quad -\lambda\left(\frac{1}{\gamma_{ms}}\mathrm{Tr}(\mathbf{H}_{am}\mathbf{W}) - \sigma_2^2 - \frac{\sigma_3^2}{\rho}\right) \\
&\quad -\mu\left(\mathrm{Tr}\left(\mathbf{H}_{am}\mathbf{W}\right) + p_{ms}|h_{mm}|^2 - \frac{\varepsilon}{\eta(1-\rho)}\right) \\
&\quad -\alpha\left(p_{ms} - \frac{\sigma_1^2\gamma_{ap}}{\|\mathbf{h}_{ma}\|^2}\right) + \beta(p_{ms} - \varepsilon) - \mathrm{Tr}\Big(\mathbf{\Phi}\mathbf{W}\Big).
\end{aligned} \tag{20}$$

Accordingly, the dual function of **P1-SDR** is given by

$$\begin{aligned}
&g\left(\lambda, \mu, \alpha, \beta, \Phi\right) \\
&= \min_{\mathbf{W}, p_{ms}, \rho}\left\{\mathrm{Tr}(\mathbf{A}\mathbf{W}) + b \cdot p_{ms} + \frac{\lambda\sigma_3^2}{\rho} + \frac{\mu\varepsilon}{\eta(1-\rho)} + k\right\}
\end{aligned}$$

where

$$\begin{cases}
\mathbf{A} = c\mathbf{I}_N - \left(\dfrac{\lambda}{\gamma_{ms}} + \mu\right)\mathbf{H}_{am} - \Phi, & (21) \\
b = 1 - \mu|h_{mm}|^2 + \beta - \alpha - c, & (22) \\
k = \lambda\sigma_2^2 + \dfrac{\alpha\sigma_1^2\gamma_{ap}}{\|\mathbf{h}_{ma}\|^2} - \varepsilon\beta & (23)
\end{cases}$$

Let λ^*, μ^*, α^* and Φ^* denote the optimal dual solution for problem **P1-SDR**. We then define

$$\mathbf{A}^* = c\mathbf{I}_N - \left(\tfrac{\lambda^*}{\gamma_{ms}} + \mu^*\right)\mathbf{H}_{am} - \Phi^*. \tag{24}$$

According to the K.K.T conditions [18], we obtain the following equations

$$\frac{\partial\mathcal{L}(\mathbf{W}, p_{ms}, \rho, \lambda, \mu, \alpha, \beta, \Phi)}{\partial\mathbf{W}} = \mathbf{0}, \tag{25}$$

$$\frac{\partial\mathcal{L}(\mathbf{W}, p_{ms}, \rho, \lambda, \mu, \alpha, \beta, \Phi)}{\partial\rho} = 0, \tag{26}$$

$$\lambda^*\left(\frac{1}{\gamma_{ms}}\mathrm{Tr}(\mathbf{H}_{am}\mathbf{W}^*) - \sigma_2^2 - \frac{\sigma_3^2}{\rho^*}\right) = 0, \tag{27}$$

$$\mu^*\left(\mathrm{Tr}\left(\mathbf{H}_{am}\mathbf{W}^*\right) + p_{ms}^*|h_{mm}|^2 - \frac{\varepsilon}{\eta(1-\rho^*)}\right) = 0, \tag{28}$$

$$\Phi^*\mathbf{W}^* = \mathbf{0}, \tag{29}$$

where only the necessary equations are listed. We can further derive (25) and (26) as

$$\begin{cases}
\mathbf{A}^* = \mathbf{0}, & (30) \\
\rho^* = \dfrac{\sqrt{\lambda^*\sigma_3^2}}{\sqrt{\lambda^*\sigma_3^2} + \sqrt{\frac{\mu^*\varepsilon}{\eta}}}. & (31)
\end{cases}$$

From (24) and (30), there is

$$\Phi^* = c\mathbf{I}_N - \left(\frac{\lambda^*}{\gamma_{ms}} + \mu^*\right)\mathbf{H}_{am}. \tag{32}$$

It is observed from (31) that if $\lambda^* = 0$ and $\mu^* > 0$, we have $\rho^* = 0$, which contradicts the constraint (18). Similarly, if $\lambda^* > 0$ and $\mu^* = 0$, we have $\rho^* = 1$, which also causes contradiction to (18). If both $\lambda^* = 0$ and $\mu^* = 0$, we obtain $\Phi^* = c\mathbf{I}_N$ from (32). Thus $\mathbf{W}^* = \mathbf{0}$ is derived according to (29), which contradicts (15) for any $\gamma_{ms} > 0$. Consequently, it follows that $\lambda^* > 0$ and $\mu^* > 0$. In the meantime, by considering the complementary slackness in (27) and (28), we know that

$$\begin{cases}
\dfrac{1}{\gamma_{ms}}\mathrm{Tr}(\mathbf{H}_{am}\mathbf{W}^*) - \sigma_2^2 - \dfrac{\sigma_3^2}{\rho^*} = 0, & (33) \\
\mathrm{Tr}\left(\mathbf{H}_{am}\mathbf{W}^*\right) + p_{ms}^*|h_{mm}|^2 - \dfrac{\varepsilon}{\eta(1-\rho^*)} = 0. & (34)
\end{cases}$$

Hence, the first part of Proposition 1 is proved.

According to (32), we obtain

$$\begin{aligned}
&\mathrm{Rank}(c\mathbf{I}_N) \\
&= \mathrm{Rank}\left(\Phi^* + \left(\frac{\lambda^*}{\gamma_{ms}} + \mu^*\right)\mathbf{H}_{am}\right) \\
&\leq \mathrm{Rank}\left(\Phi^*\right) + \mathrm{Rank}\left(\left(\frac{\lambda^*}{\gamma_{ms}} + \mu^*\right)\mathbf{H}_{am}\right).
\end{aligned} \tag{35}$$

Since $\lambda^* > 0$, $\mu^* > 0$ and $\mathbf{H}_{am} = \mathbf{h}_{am}\mathbf{h}_{am}^H$, there must be $\mathrm{Rank}\left(\left(\frac{\lambda^*}{\gamma_{ms}} + \mu^*\right)\mathbf{H}_{am}\right) = 1$. Therefore, $\mathrm{Rank}(\Phi^*) \geq N-1$ can be obtained from (35). Due to the complementary slackness in (29), we infer that $\mathbf{W}^*$ spans the null space of Φ^*. Therefore, $\mathrm{Rank}(\mathbf{W}^*) \leq 1$. Finally we know $\mathrm{Rank}(\mathbf{W}^*) = 1$, since $\mathrm{Rank}(\mathbf{W}^*) = 0$ contradicts the constraint in (15). ■

IV. SIMULATION RESULTS

In this section, we numerically examine the proposed studies via various examples. We take $\sigma_1^2 = \sigma_2^2 = -70$ dBm, $\sigma_3^2 = -50$ dBm and $\eta = 0.5$ in all simulations. Furthermore, the signal attenuation from FD-AP to FD-MS is set as 40 dBm corresponding to an identical distance of 5 meters [1]. Due to this short distance between FD-AP and FD-MS, the light-of-sight (LOS) signal is dominant. Thus, we use Rician fading to model the channel between FD-AP and FD-MS. Specifically, $\mathbf{h}_{am}$ is given by

$$\mathbf{h}_{am} = \sqrt{\frac{K_R}{1+K_R}}\mathbf{h}_{am}^{LOS} + \sqrt{\frac{1}{1+K_R}}\mathbf{h}_{am}^{NLOS} \tag{36}$$

where $\mathbf{h}_{am}^{LOS} \in \mathbb{C}^{N\times 1}$ denotes the LOS channel, $\mathbf{h}_{am}^{NLOS} \in \mathbb{C}^{N\times 1}$ denotes the Rayleigh fading channel and K_R is the Rician factor which is set to be 5 dB. The LOS channel is modeled as a far-field uniform linear antenna array model with $\mathbf{h}_{am}^{LOS} = 10^{-2}[1, e^{j\theta}, e^{j2\theta}, \ldots, e^{j(N-1)\theta}]^T$ and $\theta = 2\pi d \sin(\phi)/\lambda_{wave}$, where d is the spacing between successive antenna elements at FD-AP, λ_{wave} is the carrier wavelength, and ϕ is the direction of FD-MS to FD-AP. We set $d = \lambda_{wave}/2$ and $\phi = 60^\circ$. On the other hand, the non-light-of-sight (NLOS) channel is modeled as a Rayleigh fading channel where each element being a CSCG random variable with zero mean and covariance of -40 dB [1]. Since modeling $\mathbf{h}_{ma}$ is the same with $\mathbf{h}_{am}$, the description is omitted for brevity. Meanwhile, the loop channel is modeled as $h_{mm} = \sqrt{\beta_{mm}}$, where $\beta_{mm} = -15$ dB denotes the loop-link path loss [11].

In order to verify the optimality of the proposed solution, we demonstrate the weighted sum transmit power versus ρ, as well as the corresponding optimal ρ^* obtained by the proposed solution under different ε. We set $\gamma_{ap} = \gamma_{ms} = 17$ dB, $c = 0.1$ and $M = N = 4$ while ρ sweeps from -40 to -1 dB. Moreover, ε is set as 0, -2, -4 ,-6 and -8 dBm, respectively. As shown in Fig. 3, there exists one global minimum weighted sum transmit power as ρ increases under a fixed ε. Meanwhile, the optimal value of weighted sum transmit power corresponding to ρ^* obtained by the proposed solution matches with the minimum point in Fig. 3, which validates the optimality of the proposed solution.

Next, we compare the weighted sum transmit power of the FD system proposed in Section II with the HD system (We have briefly described the HD system in Appendix B) to show the benefits brought by the self-energy recycling. All the corresponding parameters of the two systems are set to be the same. Specifically, we take $c = 0.1$, $\gamma_{ap} = \gamma_{ms} = 11$, 17 and 23 dB respectively while the energy target ε sweeps from -10 dBm to -1 dBm. The weighted sum transmit powers versus ε with different SNR targets for both systems are shown in Fig. 4. With a given SNR target, it is observed that the weighted sum transmit power of the FD system is reduced by approximately 20 dB compared to that of the HD system. This significant transmit power reduction is caused by the self-energy recycling since the self-energy can be regarded as an extra energy for complementing the energy supplied by FD-AP. On the other hand, the optimal transmit power of both the systems becomes larger with the increasing SNR targets so as to satisfy the equality (33) which has been proved in Proposition 1.

Moreover, the impact of the number of transmit antennas at FD-AP is also investigated. We set $c = 0.1$, $\gamma_{ap} = \gamma_{ms} = 17$ dB while N varies from 4 to 32. Meanwhile, the energy target ε is set as -2, -4, -6 ,-8 and -10 dBm, respectively. As shown in Fig. 5, the weighted sum transmit power decreases with the increasing N under a given ε. Because the lager N is, the more concentrated beamforming is. Due to the increasing beamforming gain, the weighted sum power is reduced.

In addition, to illustrate the influence of beamforming, the proposed system is compared with a similar FD system where the FD-AP do not use beamforming to transmit the signal.

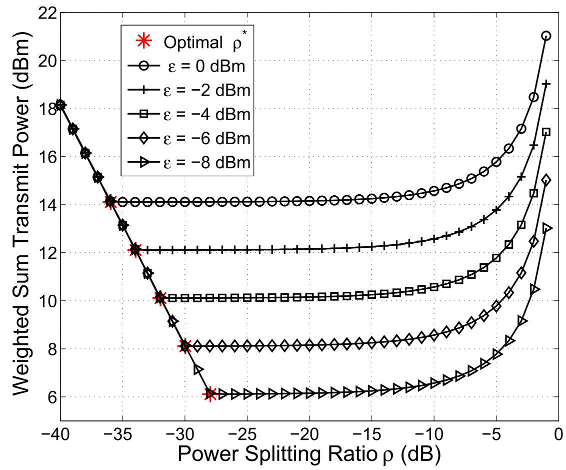

FIGURE 3. The weighted sum transmit power of the FD system versus ρ and optimal solutions obtained by solving P1-SDR under different energy targets.

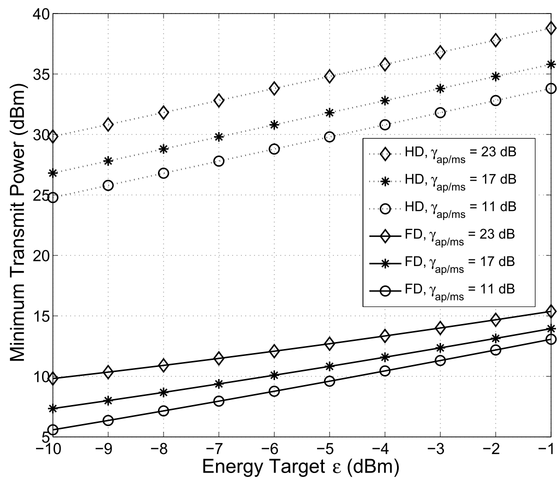

FIGURE 4. The weighted sum transmit power of the HD system and the FD system versus energy target ε with different SNR targets.

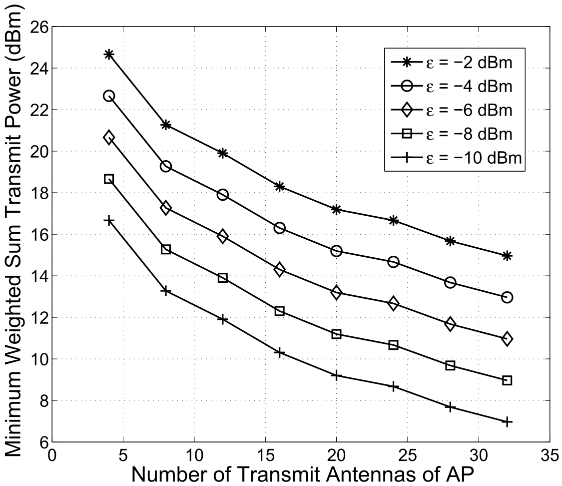

FIGURE 5. The weighted sum transmit power over N with different energy targets.

The weighted sum transmit power of the two systems versus ε under different N is shown in Fig. 6. It is observed that the weighted sum transmit power is reduced via beamforming since beamforming enables FD-AP focus the signal on FD-MS. On the other hand, due to beamforming gain increasing with N, the weighted sum power of the system with beamforming becomes smaller, while the weighted sum power of the system without beamforming becomes larger. Therefore, the weighted transmit power gap between the two systems becomes larger as N increases.

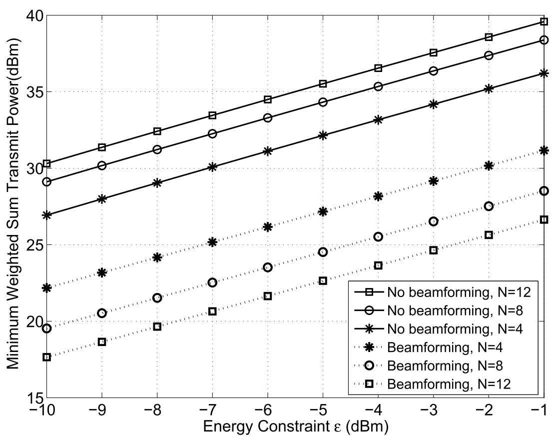

FIGURE 6. The impact of beamforming on weighted sum transmit power under different N.

Finally, the optimal transmit power of FD-AP and FD-MS versus the weight factor c under different ε are presented in Fig. 7. We take $\gamma_{ap} = \gamma_{ms} = 17$ dB, $M = N = 4$ while c changes from -40 to -5 dB with $\varepsilon = 10, 0$ and -10 dBm respectively. It can be inferred from Fig. 7 that the transmit power of FD-AP decreases with the rising c while the transmit power of FD-MS does the opposite. Because the increasing c causes that minimizing the transmit power of FD-AP takes

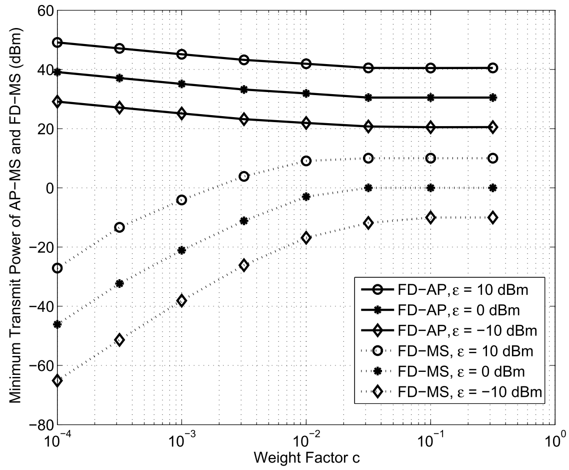

FIGURE 7. Transmit power of FD-AP and FD-MS over weight factor c under different ε.

the dominating place in the optimization problem. As a result of the decreasing transmit power of FD-AP, the power of FD-MS has to be increased to keep the equality (34) always hold. Since the transmit power of FD-MS is bounded by ε, it will increase until reaching the threshold. Therefore, if the weight factor is large enough, the optimal solution for p_{ms} will be given by

$$p^*_{ms} = \varepsilon. \tag{37}$$

Furthermore, the closed-form optimal solution for the power split ratio ρ can be easily obtained via substituting (37) into (33) and (34). Hence, we have

$$\rho^* = \frac{-B + \sqrt{B^2 - 4AC}}{2A}, \tag{38}$$

where $A = \gamma_{ms}{\sigma_2}^2 + \varepsilon|h_{mm}|^2$, $B = \varepsilon/\eta + \gamma_{ms}{\sigma_3}^2 - A$, $C = -\gamma_{ms}{\sigma_3}^2$. And it can be easily shown that $0 < \rho^* < 1$,which is omitted for brevity. With (37) and (38), the original problem **P1** can be significantly simplified. Unfortunately, the optimal closed-form of the beamforming vector w^* still is still not available due to the constraints (10) and (11). Even though the closed-form can not be obtained, the simplified problem can be solved in a much more efficient way.

V. CONCLUSION

In this paper, we proposed a new FD transmission structure with SWIPT and self-energy recycling. We jointly designed the optimal transmit beamforming vector of FD-AP, the receive PS ratio of FD-MS, and transmit power value of FD-MS to minimize the weighted sum transmit power of the whole system. Since the original problem is non-convex, SDR was applied to convert it into a new convex problem that serves as the lower bound of the original one. Most importantly, we strictly proved that the rank-one constraint is satisfied by the solutions of SDR problem, which makes them also the optimal solutions to the original problem.

Simulation results verify the optimality of the proposed solutions and reveal that self-energy recycling can dramatically reduce the weighted sum transmit power of the FD system.

APPENDIX A
THE FEASIBILITY OF PROBLEM P1

Obviously, **P1** is feasible if and only if the following problem is feasible.

$$P1 - feasibility:$$

$$\text{find} \quad \boldsymbol{w},\ p_{ms},\ \rho$$

$$\text{s.t.} \quad \frac{1}{\gamma_{ms}} |\mathbf{h}_{am}^{H}\boldsymbol{w}|^2 \geq \sigma_2^2 + \frac{\sigma_3^2}{\rho}, \tag{39}$$

$$|\mathbf{h}_{am}^{H}\boldsymbol{w}|^2 + p_{ms}|h_{mm}|^2 \geq \frac{\varepsilon}{\eta(1-\rho)}, \tag{40}$$

$$\frac{\sigma_1^2 \gamma_{ap}}{\|\mathbf{h}_{ma}\|^2} \leq p_{ms} \leq \varepsilon, \tag{41}$$

$$0 < \rho < 1. \tag{42}$$

For any given $\rho \in (0, 1)$, there always exists an appropriate $\boldsymbol{w}$ whose magnitude, i.e. $\|\boldsymbol{w}\|$, is large enough to meet the constraints (39) and (40) since the power of FD-AP is not limited. Obviously, a proper p_{ms} and ρ can always be found to meet (41) and (42), respectively .Thus, problem P1 is always feasible.

APPENDIX B
HALF-DUPLEX POINT-TO-POINT SYSTEM WITH SWIPT

We describe a HD point-to-point communication system with SWIPT for comparison. Assume HD-AP is equipped with $M+N$ antennas while HD-MS has one single antenna. Unlike the system in Fig. 1, the downlink and uplink data can not be transmitted simultaneously due to the HD operation. Thus, the self-interference does not exist so that the self-energy recycling is not enabled. In addition, the information transmission between HD-MS and HD-AP needs to be divided into two phases. In the first phase, HD-MS receives signal from HD-AP and applies PS scheme to decode information and harvest energy simultaneously. In the second phase, HD-MS sends the uplink data to HD-AP. The rest assumptions of the HD system is the same as the proposed FD system.

$$P2: \quad \min_{\boldsymbol{w},\ p_{ms},\ \rho} \quad c \cdot \|\boldsymbol{w}\|^2 + (1-c)\cdot p_{ms}$$

$$\text{s.t.} \quad \frac{p_{ms}\|\mathbf{h}_{ma}\|^2}{\sigma_1^2} \geq \gamma_{ap},$$

$$\frac{\rho|\mathbf{h}_{am}^{H}\boldsymbol{w}|^2}{\rho\sigma_2^2 + \sigma_3^2} \geq \gamma_{ms},$$

$$\eta(1-\rho)|\mathbf{h}_{am}^{H}\boldsymbol{w}|^2 \geq \varepsilon,$$

$$0 \leq p_{ms} \leq \varepsilon,$$

$$0 < \rho < 1.$$

Similarly, we could apply SDR to solve **P2** and obtain a new problem

$$P2 - SDR: \quad \min_{\mathbf{W},\ p_{ms},\ \rho} \quad c \cdot \mathrm{Tr}(\mathbf{W}) + (1-c)\cdot p_{ms}$$

$$\text{s.t.} \quad \frac{1}{\gamma_{ms}}\mathrm{Tr}(\mathbf{H}_{am}\mathbf{W}) \geq \sigma_2^2 + \frac{\sigma_3^2}{\rho},$$

$$\mathrm{Tr}(\mathbf{H}_{am}\mathbf{W}) \geq \frac{\varepsilon}{\eta(1-\rho)},$$

$$\frac{\sigma_1^2\gamma_{ap}}{\|\mathbf{h}_{ma}\|^2} \leq p_{ms} \leq \varepsilon,$$

$$0 < \rho < 1,$$

$$\mathbf{W} \succeq \mathbf{0}.$$

Apparently, problem **P2-SDR** is convex, and the proof of its equivalence to **P2** is omitted for brevity.

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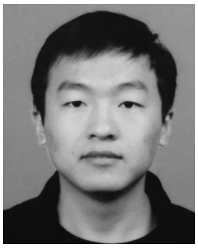

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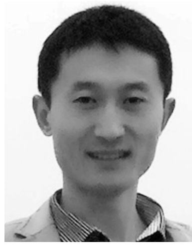

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