Secure Beamforming Design for SWIPT in MISO Full-Duplex Systems

Alexander A. Okandeji[1], Muhammad R. A. Khandaker[2], WONG Kai-Kit[2], ZHANG Yangyang[3], and ZHENG Zhongbin[4]

(1. Department of Electrical and Computer Engineering, Ighinedion University Okada, PMB 0006, Benin city, Edo State, Nigeria;
2. Department of Electronic and Electrical Engineering, University College London, WC1E 7JE, United Kingdom;
3. Kuang-Chi Institute of Advanced Technology, Shenzhen 518057, China;
4. East China Institute of Telecommunications, China Academy of Information and Communications Technology, Shanghai 200001, China)

◿ Abstract

This paper investigates the problem of bi-directional secure information exchange for a multiple-input single-output (MISO) broadcast channel in presence of potential and external eavesdroppers capable of decoding the confidential messages. Specifically, a multi-antenna base station (BS) simultaneously sends wireless information and power to a set of dual-antenna mobile stations (MSs) using power splitters (PSs) in the downlink and receives information in the uplink in full-duplex (FD) mode. We address the joint design of the receiver PS ratio and the transmit power at the MSs, the artificial noise covariance, and the beamforming matrix at the BS in order to guarantee the individual secrecy rate and energy harvesting constraints at each receiver, and the signal-to-interference plus noise ratio (SINR) at the BS and MSs. Using semidefinite relaxation (SDR) technique, we obtain solution to the problem with imperfect channel state information (CSI) of the self-interfering channels. Simulation results are presented to demonstrate the performance of our proposed scheme.

◿ Keywords

physical layer security; full-duplex; SWIPT; energy harvesting

1 Introduction

Since the continuous operation of wireless devices is mostly constrained by the limited capacity of their power sources, radio frequency (RF) energy harvesting (EH) has been identified as a promising technique to power future wireless devices. Technically, RF signals that carry information can also be used as a vehicle for transporting energy at the same time. For this reason, the study of simultaneous wireless information and power transfer (SWIPT) has attracted huge interest from industrial and academic communities [1]–[7]. Effectively, wireless data and energy access can be made available to mobile users through the application of SWIPT, which in turn gives mobile users great convenience. SWIPT technology has been practically deployed in wireless communication networks. For example, [8] considered the practical deployment of SWIPT in a wireless sensor network mounted at a remote and difficult-to-access location, and powered only by a battery with limited operation time. Recharging or replacing the battery may be inconvenient, and often attract huge cost. Consequently, electromagnetic energy emanating from numerous radio and television broadcasting in the environment may serve as an opportunistic as well as greener alternative for harvesting energy to power such devices.

SWIPT was first introduced in [9] to describe the trade-off between the rates at which energy and dependable information is transmitted over a noisy channel and thus, a capacity energy function was proposed. A key assumption in [9] was the possibility of the receiver capable of simultaneously decoding information and harvesting energy from the same received signal. However, this assumption does not hold in practice as practical circuits for harvesting energy from the received signal are not able to decode information directly. Consequently, to facilitate wireless information and power transfer at the receiver, the work done in [10] proposed two practical receiver architecture designs for SWIPT, namely time switching (TS) and power splitting (PS). Specifically, in the TS design architecture, the receiver switches over time to achieve information decoding and energy harvesting, while with PS, the receiver splits the receive signal into two streams of different power in order to decode information and harvest energy separately. It was shown in [10] that PS generally achieves better performance than TS as the TS receiver architecture scheme, in general, can be regarded as a special form of PS with only a binary split power ra-

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tio. In this paper, we focus on the PS design architecture.

Conventionally, half-duplex (HD) communication modes, in which the transmission and reception of radio signals occur over orthogonal time or frequency resources in order to avoid crosstalk, have been employed for wireless communications. Recent advances, however, suggest full-duplex (FD) communications that allow simultaneous transmission and reception of signals over the same radio channel in order to maximally use the limited available spectral resource [11], [12]. The key challenge to achieve FD communications is the presence of self-interference (SI). Thus, for wireless communication systems to communicate in FD mode, the generated SI must be significantly suppressed, if not cancelled completely, to receiver's noise floor [13]. Digital SI cancellation (SIC) for FD wireless systems was studied in [14]. Authors in [14] demonstrated the possibility of achieving SIC of 70 dB and 76 dB for antenna separation and digital cancellation at 20 cm and 40 cm spacing, respectively, between interfering antennas. It suffices to say therefore that radios can work in FD if SI can be significantly suppressed. Interestingly, the study of FD systems brings a new opportunity for SWIPT in FD systems [1], [7]. In particular, authors in [1] investigated the end-to-end sum-rate maximization approach for SWIPT in FD system while maintaining the energy harvesting threshold at each node by optimizing the receive power splitter and transmit power at each node. In contrast, [7] investigated the joint optimization of the transmit power at the source nodes and the two-way relay beamforming matrix for SWIPT with a FD MIMO amplify and forward (AF) relay employing PS, where the achievable sum-rate is maximized subject to energy harvesting and individual power constraints.

Wireless communication networks are susceptible to eavesdropping; however, SWIPT systems are more susceptible to eavesdropping due to the inherent characteristics of SWIPT and the open nature of wireless channel. The security issue in SWIPT systems is therefore a critical issue that needs to be addressed. Existing works on communication security have adopted the cryptographic technique. Although this technique may guarantee communication security to some degree in the application layer, the suitability of these technique on SWIPT systems is not guaranteed since it requires complex encryption/decryption algorithms and key distribution [15]. Alternatively, physical-layer (PHY) security has been identified to perfectly guarantee secure wireless information transmission against eavesdropping by using the physical properties of wireless channels such as fading and interference [16], [17]. In this context, artificial noise (AN), injected into the transmitted signal, is used to confuse the eavesdropper [18].

The fundamental study of the integration of PHY security and SWIPT was first considered in [15]. In particular, authors in [15] investigated the joint design of the beamforming vectors and power allocation to maximize the secrecy rate or the total harvested energy under a multi-user multiple-input single-output (MISO) SWIPT scenario. The work done in [19] proposed a robust secure transmission scheme which minimises the total transmit power in a multi-user MISO SWIPT system. In contrast, authors in [20] and [21] extended the study of PHY security into a multiple-input multiple-output (MIMO) broadcast SWIPT system. To maximise the secrecy rate in a MIMO SWIPT scenario, [20] investigated the secure transceiver beamforming design while ensuring efficient energy harvesting. On the other hand, authors in [21] considered the harvested energy maximization approach subject to secure information transmission by designing the transmit covariance.

In contrast to existing works, this paper considers the novel integration of SWIPT, FD technology and PHY security. In particular, this paper extends the secure HD MISO SWIPT communication scenario in [22] to the FD case, where the harvested energy at the MSs is utilized to send feedback information to the BS. We aim to minimize the end-to-end transmit power for SWIPT in FD MISO systems (**Fig. 1**) while satisfying the secrecy rate at each MSs, and the QoS requirements for each MS by optimizing jointly the receive PS ratio and the transmit power at the MSs, the beamforming matrix at the BS, and the AN covariance matrix. Specifically, we assume perfect CSI, which is accomplished from fine estimation by transmitting dedicated training symbols at the receiver, for the uplink and downlink channels for the MISO SWIPT system. In contrast, loop channels are assumed to be imperfect due to the fact that the distribution of SI channels are unknown. Also, SI channel measurement results obtained in [23] indicate that the SI channel has a multipath nature. Specifically, it was shown in [23] that multiple paths arising from SI channel measurements are known to have higher power relative to the line of sight (LOS) path. Consequently, there is the need to suppress SI with multiple path nature, through the use of an adaptive cancellation technique whose measurement can be used to cancel both the LOS path and the delayed version of the same, which is not the objective

BS: base station   MS: mobile station   EH: energy harvesting   ID: information decoder
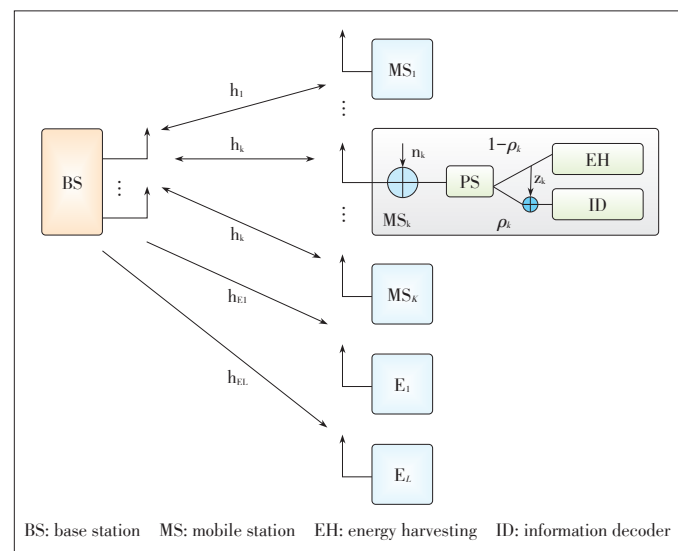
▲Figure 1. An FD MISO SWIPT system.

to be achieved in this work. Nonetheless, for convenience, SI channels can be modelled as Gaussian channels [7]. Accordingly, due to insufficient knowledge of the SI channel, a deterministic model for channel uncertainties is considered, where the magnitude of the estimation error as well as the SI power is bounded. Since the problem is non-convex, we propose an alternating optimization approach.

We denote scalars by non-bold letters and boldface lowercase letters are used to represent vectors, while boldface uppercase letters are used for matrices. For a square matrix $\mathbf{A}$, $\mathrm{Tr}(\mathbf{A})$, $\mathrm{Rank}(\mathbf{A})$, $\mathbf{A}^T$ and $\mathbf{A}^H$ denote its trace, rank, transpose and conjugate transpose, respectively, while $\mathbf{A} \geq 0$ means that $\mathbf{A}$ is a positive semidefinite matrix. Also, $\mathbf{I}_n$ denotes an $n \times n$ identity matrix, $\|\cdot\|$ returns the Euclidean norm. The distribution of a circularly symmetric complex Gaussian (CSCG) random vector with mean μ and covariance matrix $\mathbf{C}$ is denoted by $\mathcal{CN}(\mu, \mathbf{C})$. Finally, $\mathbb{C}^{m \times n}$ denotes the space of $m \times n$ complex matrices.

2 System Model and Problem Statement

2.1 System Model

In this paper, we consider the end-to-end transmit power minimization approach for secure SWIPT in a MISO FD system where the BS simultaneously transmits K independent confidential messages to K dual-antenna (one for transmission, one for reception) receivers in the downlink, in the presence of L external single-antenna eavesdroppers, and receives information in the uplink in FD mode. In addition to the external eavesdroppers, each receiver attempts to eavesdrop messages intended for other receivers. Hence we call them potential eavesdroppers. To ensure secure information transmission, AN is superimposed to the transmitted signal to confuse these potential as well as the external eavesdroppers. We denote the number of transmit and receive antennas at the BS, respectively, as N_t and N_r, and each MS uses an identical pair of transmitter and receiver antennas for signal transmission and reception. In the downlink transmission, the BS performs transmit beamforming to send information to the receivers while the MSs, using the harvested energy from the received signal, send feedback to the BS in the uplink with transmit power $P_{\mathrm{up},k}, k=1,...,K$. Therefore, the complex baseband signal transmitted at the BS is expressed as

$$x = \sum_{k=1}^{K} \mathbf{a}_k s_k + \mathbf{b}, \tag{1}$$

where $s_k \sim \mathcal{CN}(0, 1)$ denotes the confidential message intended for the kth receiver, $\mathbf{a}_k \in \mathbb{C}^{N_t \times 1}$ denotes the corresponding precoding vector. $\mathbf{b} \in \mathbb{C}^{N_t \times 1}$ is the AN vector modelled as $\mathbf{b} \sim \mathcal{CN}(0, \mathbf{V})$, where $\mathbf{V} = \mathbf{b}\mathbf{b}^H$ denotes the covariance matrix of

$\mathbf{b}$. For convenience, we use $\mathcal{K} = \{1, \dots, K\}$, $\mathcal{L} = \{1, \dots, L\}$ and $\mathcal{M}_k = \{1, \dots, k-1, k+1, \dots, K\}$ to denote the index set of receivers, external eavesdroppers, and receivers except the kth receiver, respectively. The received signal at the kth receiver and lth external eavesdropper can be expressed as

$$y_k = \mathbf{h}_{\mathrm{dl},k}^H \mathbf{x} + h_{\mathrm{SI},k} m_k + z_k, \forall k, \tag{2}$$

$$y_{e,l} = \mathbf{g}_l^H \mathbf{x} + n_{e,l}, \forall l, \tag{3}$$

where $\mathbf{h}_{\mathrm{dl},k} \in \mathbb{C}^{1 \times N_t}$ and $\mathbf{g}_l \in \mathbb{C}^{1 \times N_t}$ represent the channel vectors from the BS to the kth receiver, and from the BS to the lth external eavesdropper, respectively, and m_k denotes the information bearing symbol transmitted by MS_k. The antenna noises at the kth receiver and the lth external eavesdropper are modelled as $z_k \sim \mathcal{CN}(0, \sigma_k^2)$ and $n_{e,l} \sim \mathcal{CN}(0, \sigma_{e,l}^2)$. We assume each receiver adopts the PS scheme to coordinate the process of information decoding and energy harvesting. Specifically, a ρ portion of the received signal power is split to the information decoder (ID) while the remaining $(1-\rho)$ portion is split to the EH. Accordingly, the signal split to the ID of MS_k is given by

$$y_k^{\mathrm{ID}} = \sqrt{\rho_k} \left(\underbrace{\mathbf{h}_{\mathrm{dl},k}^H \mathbf{a}_k s_k}_{\text{desired signal}} + \underbrace{\sum_{j \neq k}^{K} \mathbf{h}_{\mathrm{dl},k}^H \mathbf{a}_j s_j}_{\text{interfering signal}} + \underbrace{\mathbf{h}_{\mathrm{dl},k}^H \mathbf{b}}_{\text{artificial noise}} + \underbrace{h_{\mathrm{SI},k} m_k}_{\text{self interference}} + z_k \right) + z_{a,k}, \forall k. \tag{4}$$

where $z_{a,k} \sim \mathcal{CN}(0, \sigma_{a,k}^2)$ denotes the additional processing noise introduced by the ID of the kth receiver. Consequently, the signal-to-interference plus noise ratio (SINR) at the ID of the kth receiver can be expressed as

$$\Gamma_k = \frac{\rho_k |\mathbf{h}_{\mathrm{dl},k}^H \mathbf{a}_k|^2}{\rho_k (\sum_{j \neq k} |\mathbf{h}_{\mathrm{dl},k}^H \mathbf{a}_j|^2 + \mathrm{Tr}(\mathbf{h}_{\mathrm{dl},k}^H \mathbf{h}_{\mathrm{dl},k} \mathbf{V}) + |h_{\mathrm{SI},k}|^2 P_{\mathrm{up},k} + \sigma_k^2) + \sigma_{a,k}^2}. \tag{5}$$

Also, the signal received at the EH of the kth receiver can be written as

$$y_k^{\mathrm{EH}} = \sqrt{1-\rho_k} \left(\sum_{j=1}^{K} \mathbf{h}_{\mathrm{dl},k}^H \mathbf{a}_j s_j + \mathbf{h}_{\mathrm{dl},k}^H \mathbf{b} + h_{\mathrm{SI},k} m_k + z_k \right), \forall k. \tag{6}$$

Then, the energy harvesting rate (i.e., power) at the kth EH can be expressed as [22]

$$Q_k = \eta_k (1-\rho_k) \left(\sum_{j=1}^{K} |\mathbf{h}_{\mathrm{dl},k}^H \mathbf{a}_j|^2 + \mathbf{h}_{\mathrm{dl},k}^H \mathbf{V} \mathbf{h}_{\mathrm{dl},k} + P_{\mathrm{up},k} |h_{\mathrm{SI},k}|^2 + \sigma_k^2 \right), \tag{7}$$

where η_k is the energy conversion efficiency which for notational simplicity is assumed to be unity throughout this paper.

Special Topic ◀

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Meanwhile, the received signal at the BS can be written as

$$y^{\mathrm{BS}} = \underbrace{\sum_{k=1}^{K} \mathbf{h}_{\mathrm{ul},k} m_k}_{\text{desired signal}} + \underbrace{\sum_{j=1}^{K} \mathbf{H}_{\mathrm{SI,BS}} \mathbf{a}_j s_j}_{\text{self interference}} + \mathbf{n}_{\mathrm{BS}}, \tag{8}$$

where $\mathbf{h}_{\mathrm{ul}}$ is the channel vector from the MS to the transmit antennas at the BS and $\mathbf{n}_{\mathrm{BS}}$ denotes the additive white Gaussian noise (AWGN) vector at the BS modelled as $\mathbf{n}_{\mathrm{BS}} \sim \mathcal{CN}(0, \sigma_{BS}^2 \mathbf{I})$. To equalize the uplink signal from MS_k, we assume that the BS applies a receive beamformer $\mathbf{w}_k$ expressed as given in [24]

$$s_k^{\mathrm{UL}} = \mathbf{w}_k^H \mathbf{h}_{\mathrm{ul},k} m_k + \mathbf{w}_k^H \sum_{j \neq k}^{K} \mathbf{h}_{\mathrm{ul},j} m_j + \mathbf{w}_k^H \sum_{j=1}^{K} \mathbf{H}_{\mathrm{SI,BS}} \mathbf{a}_j + \mathbf{w}_k^H \mathbf{n}_{\mathrm{BS}}. \tag{9}$$

Therefore, the SINR at the BS for the kth user's signal can be written as [24]

$$\Gamma_k^{\mathrm{BS}} = \frac{P_{\mathrm{up},k} \left| \mathbf{w}_k^H \mathbf{h}_{\mathrm{ul},k} \right|^2}{\sum_{j \neq k}^{K} P_{\mathrm{up},j} \left| \mathbf{w}_k^H \mathbf{h}_{\mathrm{ul},j} \right|^2 + \sum_{j=1}^{K} \left| \mathbf{w}_k^H \mathbf{H}_{\mathrm{SI,BS}} \mathbf{a}_j \right|^2 + \sigma_{BS}^2 \|\mathbf{w}_k\|^2}. \tag{10}$$

As in [22], we also assume that the legitimate receivers themselves may attempt to overhear the signals transmitted to other receivers and we want to guarantee secrecy against any such attempt. As such, we denote the received SINR which corresponds to the signal transmitted for the kth receiver at the mth potential eavesdropper and the lth external eavesdropper as

$$\Gamma_{m,k} = \frac{\left| \mathbf{h}_{\mathrm{dl},m}^H \mathbf{a}_k \right|^2}{\sum_{j \neq k} \left| \mathbf{h}_{\mathrm{dl},m}^H \mathbf{a}_j \right|^2 + \mathbf{h}_{\mathrm{dl},m}^H \mathbf{V} \mathbf{h}_{\mathrm{dl},m} + P_{\mathrm{up},m} \left| h_{\mathrm{SI},m} \right|^2 + \sigma_m^2}, \forall k, \forall m, \tag{11}$$

and

$$\Gamma_l^k = \frac{\left| \mathbf{g}_l^H \mathbf{a}_k \right|^2}{\sum_{j \neq k} \left| \mathbf{g}_l^H \mathbf{a}_j \right|^2 + \mathbf{g}_l^H \mathbf{V} \mathbf{g}_l + \sigma_{e,l}^2}, \forall k, \forall l. \tag{12}$$

In this work, the communication security is considered from the viewpoint of individual security [25], [26]. Consequently, the achievable individual secrecy rate at the kth receiver is given as [22]

$$R_k = \left[\log_2(1 + \Gamma_k) - \log_2 \left(1 + \max_{\forall m, \forall l} \left\{ \Gamma_{m,k}, \Gamma_l^k \right\} \right) \right]^+, \tag{13}$$

where $[x]^+ = \max(0, x)$.

2.2 Modelling the SI

Due to the insufficient knowledge of the SI channel, it is impossible to completely eliminate residual SI (RSI). For this reason, we consider a deterministic model for imperfect SI channels. In particular, the actual SI channels $h_{\mathrm{SI},k}$ and $\mathbf{H}_{\mathrm{SI,BS}}$ are assumed to respectively lie in the neighbourhood of the estimated channels $\hat{h}_{\mathrm{SI},k}$ and $\hat{\mathbf{H}}_{\mathrm{SI,BS}}$ that are available at the corresponding nodes. Therefore, the actual channels due to imperfect channel estimates can be modelled as

$$h_{\mathrm{SI},k} = \hat{h}_{\mathrm{SI},k} + \triangle h_{\mathrm{SI},k}, \tag{14a}$$

$$\mathbf{H}_{\mathrm{SI,BS}} = \hat{\mathbf{H}}_{\mathrm{SI,BS}} + \triangle \mathbf{H}_{\mathrm{SI,BS}}, \tag{14b}$$

where $\triangle h_{\mathrm{SI},k}$ and $\triangle \mathbf{H}_{\mathrm{SI,BS}}$ represent the channel uncertainties assumed to be bounded as

$$\left| \triangle h_{\mathrm{SI},k} \right| = \left| h_{\mathrm{SI},k} - \hat{h}_{\mathrm{SI},k} \right| \leqslant \epsilon_1, \tag{15a}$$

$$\left\| \triangle \mathbf{H}_{\mathrm{SI,BS}} \right\| = \left\| \mathbf{H}_{\mathrm{SI,BS}} - \hat{\mathbf{H}}_{\mathrm{SI,BS}} \right\| \leqslant \epsilon_2, \tag{15b}$$

for some $\epsilon_1, \epsilon_2 \geqslant 0$. It is worth noting that the bounding values $\{\epsilon_k\}$ are dependent on the accuracy of the CSI estimates. Consequently, to efficiently define the worst-case SI level, we modify (14a) and (14b) using the triangle inequality and the Cauchy-Schwarz inequality, respectively [27]. It follows from (14a) and (14b) that

$$\left| h_{\mathrm{SI},k} \right|^2 = \left| \left(\hat{h}_{\mathrm{SI},k} + \triangle h_{\mathrm{SI},k} \right) \right|^2 \leqslant \left(\left| \hat{h}_{\mathrm{SI},k} \right| + \left| \triangle h_{\mathrm{SI},k} \right| \right)^2 \leqslant$$
$$\left| \hat{h}_{\mathrm{SI},k} \right|^2 + \epsilon_1^2 + 2\epsilon_1 \left| \hat{h}_{\mathrm{SI},k} \right|, \tag{16a}$$

$$\left\| \mathbf{H}_{\mathrm{SI,BS}} \mathbf{a}_k \right\|^2 \leqslant \left\| \mathbf{H}_{\mathrm{SI,BS}} \right\|^2 \|\mathbf{a}_k\|^2 =$$
$$\left\| \hat{\mathbf{H}}_{\mathrm{SI,BS}} + \triangle \mathbf{H}_{\mathrm{SI,BS}} \right\|^2 \|\mathbf{a}_k\|^2 \leqslant$$
$$\left(\left\| \hat{\mathbf{H}}_{\mathrm{SI,BS}} \right\| + \left\| \triangle \mathbf{H}_{\mathrm{SI,BS}} \right\| \right)^2 \|\mathbf{a}_k\|^2 \leqslant$$
$$\left(\left\| \hat{\mathbf{H}}_{\mathrm{SI,BS}} \right\|^2 + \epsilon_2^2 + 2 \left\| \hat{\mathbf{H}}_{\mathrm{SI,BS}} \right\| \epsilon_2 \right) \|\mathbf{a}_k\|^2. \tag{16b}$$

Note that in the absence of statistical information about the error, we use ϵ_k to represent the minimal knowledge of the upper-bound of the channel error which is sufficient enough to describe the error. Thus from (16a) and (16b), we obtain

$$\max_{\left| \triangle h_{\mathrm{SI},k} \right| \leqslant \epsilon_1} \left| h_{\mathrm{SI},k} \right|^2 \leqslant \left| \hat{h}_{\mathrm{SI},k} \right|^2 + \epsilon_1^2 + 2\epsilon_1 \left| \hat{h}_{\mathrm{SI},k} \right|, \tag{17a}$$

$$\max_{\left\| \triangle \mathbf{H}_{\mathrm{SI,BS}} \mathbf{a}_k \right\| \leqslant \epsilon_2} \left\| \mathbf{H}_{\mathrm{SI,BS}} \mathbf{a}_k \right\|^2 \leqslant \left(\left\| \hat{\mathbf{H}}_{\mathrm{SI,BS}} \right\|^2 + \epsilon_2^2 + 2 \left\| \hat{\mathbf{H}}_{\mathrm{SI,BS}} \right\| \epsilon_2 \right) \|\mathbf{a}_k\|^2. \tag{17b}$$

In addition, it holds that

$$\left| \left(\hat{h}_{\mathrm{SI},k} + \triangle h_{\mathrm{SI},k} \right) \right|^2 \geqslant \left(\left| \hat{h}_{\mathrm{SI},k} \right| - \left| \triangle h_{\mathrm{SI},k} \right| \right)^2 \geqslant \left| \hat{h}_{\mathrm{SI},k} \right|^2 + \epsilon_1^2 - 2 \left| \hat{h}_{\mathrm{SI},k} \right|^2 \epsilon_1. \tag{18}$$

Here, we assume, in general, that $\left| \hat{h}_{\mathrm{SI}} \right| \geqslant \left| \triangle h_{\mathrm{SI}} \right|$ implies that the error $\left| \triangle h_{\mathrm{SI}} \right|$ is sufficiently small relative to the estimate or the estimate is not meaningful. Consequently,

$$\min_{\left| \triangle h_{\mathrm{SI},k} \right| \leqslant \epsilon_1} \left| h_{\mathrm{SI},k} \right|^2 \geqslant \left| \hat{h}_{\mathrm{SI},k} \right|^2 + \epsilon_1^2 - 2 \left| \hat{h}_{\mathrm{SI},k} \right| \epsilon_1. \tag{19}$$

2.3 Problem Statement

In this work, we consider the case where each MS_k is char-

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acterised by strict quality of service (QoS) requirement. In particular, to ensure a continuous information transfer, the QoS constraints require that the SINR of the downlink channel at all times should be above a given threshold denoted by $\gamma_k^{\mathrm{DL}}, \forall k$. Furthermore, it is required that the harvested energy by each MS must be above a certain useful level specified by a prescribed threshold denoted by $Qk, \forall k$. In addition, for the uplink channel, we also assume that each MS is expected to send feedback to the BS and a strict QoS is required such that the SINR of the uplink channel is expected to be higher than a given threshold denoted as $\gamma_k^{\mathrm{UL}}, k = 1, ..., K$.

FD communication is feasible subject to the effective cancellation of SI. As an increase in transmit power causes a corresponding increase in the SI, it is important for communication nodes to transmit at optimality. To this end, our aim is to minimize the total transmit power while maintaining the achievable secrecy rate and energy harvesting constraints at each receiver as well as the non - zero uplink and downlink SINR ($\gamma^{\mathrm{UL}} > 0; \gamma^{\mathrm{DL}} > 0$). We achieve this by jointly designing the transmit beamforming vector ($\mathbf{a}$) at the BS, the AN covariance ($\mathbf{V}$), the uplink transmit power ($P_{\mathrm{up},k}$) and the receive PS ratio (ρ_k) at MS_k, $k = 1, ..., K$. Thus we formulate the optimization problem as

$$\min_{\mathbf{a}_k, \rho_k, \mathbf{w}_k, \mathbf{V}, P_{up,k}} \sum_{k=1}^{K}\left(\|\mathbf{a}_k\|^2 + P_{up,k}\right) + \mathrm{Tr}(\mathbf{V}) \quad s.t.$$

$$R_k \geq r_k, Q_k \geq \bar{Q}_k, \Gamma_k^{\mathrm{BS}} \geq \gamma^{\mathrm{UL}}, \forall k,$$

$$\mathbf{V} \geq 0, 0 < \|\mathbf{a}_k\|^2 \leq P_{\max}, \Gamma_k \geq \gamma^{\mathrm{DL}}, \forall k, \quad (20)$$

$$0 < \rho_k < 1, 0 < P_{up,k} \leq \min(\bar{Q}_k, P_{\max}), \forall k,$$

where R_k and $\bar{Q}_k$, respectively, denote the required secrecy rate and the energy harvesting constraints at each receiver. Problem (20) can explicitly be written as given in (21). Problem (21) is a non-convex problem due to the coupled optimization variables in the constraints, and hence it is very difficult to solve in closed form. Thus, we solve (21) in an alternating manner.

$$\min_{\mathbf{a}_k, \rho_k, \mathbf{w}_k, \mathbf{V}, P_{up,k}} \sum_{k=1}^{K}\left(\|\mathbf{a}_k\|^2\right) + \mathrm{Tr}(\mathbf{V}) + \sum_{k=1}^{K} P_{up,k}$$

$$s.t.$$

$$\frac{\left|\mathbf{g}_l^H \mathbf{a}_k\right|^2}{\sum_{j \neq k}\left|\mathbf{g}_l^H \mathbf{a}_j\right|^2 + \mathbf{g}_l^H \mathbf{V} \mathbf{g}_l + \sigma_l^2} \leq \gamma_e^k, \forall k, \forall l,$$

$$\frac{\rho_k\left|\mathbf{h}_{dl,k}^H \mathbf{a}_k\right|^2}{\rho_k\left(\sum_{j \neq k}\left|\mathbf{h}_{dl,k}^H \mathbf{a}_j\right|^2 + \mathbf{h}_{dl,k}^H \mathbf{V} \mathbf{h}_{dl,k} + \left|h_{\mathrm{SI},k}\right|^2 P_{up,k} + \sigma_k^2\right) + \sigma_{a,k}^2} \geq \gamma_k^{\mathrm{DL}}, \forall k,$$

$$\frac{\left|\mathbf{h}_{dl,m}^H \mathbf{a}_k\right|^2}{\sum_{j \neq k}\left|\mathbf{h}_{dl,m}^H \mathbf{a}_j\right|^2 + \mathbf{h}_{dl,m}^H \mathbf{V} \mathbf{h}_{dl,m} + \left|h_{\mathrm{SI},k}\right|^2 P_{up,k} + \sigma_m^2} \leq \gamma_e^k, \forall k, \forall m, \quad (21)$$

$$\eta_k(1 - \rho_k)\left(\sum_{j=1}^{K}\left|\mathbf{h}_{dl,k}^H \mathbf{a}_j\right|^2 + \mathbf{h}_{dl,k}^H \mathbf{V} \mathbf{h}_{dl,k} + \left|h_{\mathrm{SI},k}\right|^2 P_{up,k} + \sigma_k^2\right) \geq \bar{Q}_k,$$

$$\frac{P_{up,k}\left|\mathbf{w}_k^H \mathbf{h}_{ul,k}\right|^2}{\sum_{j \neq k}^{K} P_{up,j}\left|\mathbf{w}_k^H \mathbf{h}_{ul,j}\right|^2 + \sum_{j=1}^{K}\left\|\mathbf{H}_{\mathrm{SI,BS}} \mathbf{a}_j\right\|^2 \left\|\mathbf{w}_k\right\|^2 + \sigma_{\mathrm{BS}}^2 \|\mathbf{w}_k\|^2} \geq \gamma^{\mathrm{UL}}, \forall k,$$

$$\mathbf{V} \geq 0, 0 < \|\mathbf{a}_k\|^2 \leq P_{\max},$$

$$0 < \rho_k < 1, 0 < P_{up,k} \leq \min(\bar{Q}_k, P_{\max}).$$

It is worth noting that the upper bound of the SI at the BS and the MS_k is obtained when the source node transmits at maximum available power [24], i.e., $P_{\mathrm{up},k} = \|\mathbf{a}_k\|^2 = P_{\max}$. Consequently, we denote the upper-bounds of the SI power at the BS and MS_k as $\bar{E}$ and $\bar{G}$, respectively. Thus, the optimization problem in (21) can be upper bounded as given in (22).

$$\min_{\mathbf{a}_k, \rho_k, \mathbf{w}_k, \mathbf{V}, P_{up,k}} \sum_{k=1}^{K}\|\mathbf{a}_k\|^2 + \mathrm{Tr}(\mathbf{V}) + \sum_{k=1}^{K} P_{up,k}$$

$$s.t.$$

$$\frac{\left|\mathbf{g}_l^H \mathbf{a}_k\right|^2}{\sum_{j \neq k}\left|\mathbf{g}_l^H \mathbf{a}_j\right|^2 + \mathbf{g}_l^H \mathbf{V} \mathbf{g}_l + \sigma_l^2} \leq \gamma_e^k, \forall k, \forall l,$$

$$\frac{\rho_k\left|\mathbf{h}_{dl,k}^H \mathbf{a}_k\right|^2}{\rho_k\left(\sum_{j \neq k}\left|\mathbf{h}_{dl,k}^H \mathbf{a}_j\right|^2 + \mathbf{h}_{dl,k}^H \mathbf{V} \mathbf{h}_{dl,k} + \left(\left|\hat{h}_{\mathrm{SI},k}\right|^2 + \epsilon_1^2 + 2\epsilon_1\left|\hat{h}_{\mathrm{SI},k}\right|\right) P_{\max} + \sigma_k^2\right) + \sigma_{a,k}^2} \geq \gamma_k^{\mathrm{DL}}, \forall k,$$

$$\frac{\left|\mathbf{h}_{dl,m}^H \mathbf{a}_k\right|^2}{\sum_{j \neq k}\left|\mathbf{h}_{dl,m}^H \mathbf{a}_j\right|^2 + \mathbf{h}_{dl,m}^H \mathbf{V} \mathbf{h}_{dl,m} + \left(\left|\hat{h}_{\mathrm{SI},k}\right|^2 + \epsilon_1^2 + 2\epsilon_1\left|\hat{h}_{\mathrm{SI},k}\right|\right) P_{\max} + \sigma_m^2} \leq \gamma_e^k, \forall k, \forall m, \quad (22)$$

$$\eta_k(1 - \rho_k)\left(\sum_{j=1}^{K}\left|\mathbf{h}_{dl,k}^H \mathbf{a}_j\right|^2 + \mathbf{h}_{dl,k}^H \mathbf{V} \mathbf{h}_{dl,k} + \left(\left|\hat{h}_{\mathrm{SI},k}\right|^2 + \epsilon_1^2 - 2\epsilon_1\left|\hat{h}_{\mathrm{SI},k}\right|\right) P_{\max} + \sigma_k^2\right) \geq \bar{Q}_k,$$

$$\frac{P_{up,k}\left|\mathbf{w}_k^H \mathbf{h}_{ul,k}\right|^2}{\sum_{j \neq k}^{K} P_{up,j}\left|\mathbf{w}_k^H \mathbf{h}_{ul,j}\right|^2 + \left(\left\|\hat{\mathbf{H}}_{\mathrm{SI,BS}}\right\|^2 + \epsilon_2^2 + 2\left\|\hat{\mathbf{H}}_{\mathrm{SI,BS}}\right\|^2 \epsilon_2\right) K P_{\max} \left\|\mathbf{w}_k\right\|^2 + \sigma_{\mathrm{BS}}^2 \|\mathbf{w}_k\|^2} \geq \gamma_k^{\mathrm{UL}}, \forall k,$$

$$\mathbf{V} \geq 0, 0 < \|\mathbf{a}_k\|^2 \leq P_{\max},$$

$$0 < \rho_k < 1, 0 < P_{up,k} \leq \min(\bar{Q}_k, P_{\max}).$$

To solve problem (22), first we observe that the QoS uplink constraint does not have the PS coefficient and this is because the BS does not need to harvest energy. As a consequence, problem (22) can be decomposed into two sub-problems given in (23) and (24),

$$\min_{\mathbf{w}_k, P_{up,k}} \sum_{k=1}^{K} P_{up,k}$$

$$s.t.$$

$$\frac{P_{up,k}\left|\mathbf{w}_k^H \mathbf{h}_{ul,k}\right|^2}{\sum_{j \neq k}^{K} P_{up,j}\left|\mathbf{w}_k^H \mathbf{h}_{ul,j}\right|^2 + \bar{E}\left\|\mathbf{w}_k\right\|^2 + \sigma_{\mathrm{BS}}^2 \|\mathbf{w}_k\|^2} \geq \gamma_k^{\mathrm{UL}}, \forall k, \quad (23)$$

$$0 < P_{up,k} \leq \min(\bar{Q}_k, P_{\max}),$$

where $\bar{E} \triangleq \left(\left\|\hat{\mathbf{H}}_{\mathrm{SI,BS}}\right\|^2 + \epsilon_2^2 + 2\left\|\hat{\mathbf{H}}_{\mathrm{SI,BS}}\right\|^2 \epsilon_2\right) K P_{\max}$,

$\bar{G}_k \triangleq \left(\left|\hat{h}_{\mathrm{SI},k}\right|^2 + \epsilon_1^2 + 2\epsilon_1\left|\hat{h}_{\mathrm{SI},k}\right|\right) P_{\max}$, and

$\hat{G}_k \triangleq \left(\left|\hat{h}_{\mathrm{SI},k}\right|^2 + \epsilon_1^2 - 2\epsilon_1\left|\hat{h}_{\mathrm{SI},k}\right|\right) P_{\max}$ are the SI power associated with the energy harvesting constraints.

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$$\min_{\mathbf{a}_k,\rho_k,\mathbf{V},}\sum_{k=1}^{K}\|\mathbf{a}_k\|^2+\mathrm{Tr}(\mathbf{V})\quad s.t.$$

$$\frac{\left|\mathbf{g}_l^H\mathbf{a}_k\right|^2}{\sum_{j\neq k}\left|\mathbf{g}_l^H\mathbf{a}_j\right|^2+\mathbf{g}_l^H\mathbf{V}\mathbf{g}_l+\sigma_l^2}\leq\gamma_e^k,\forall k,\forall l,$$

$$\frac{\rho_k\left|\mathbf{h}_{\mathrm{dl},k}^H\mathbf{a}_k\right|^2}{\rho_k\left(\sum_{j\neq k}\left|\mathbf{h}_{\mathrm{dl},k}^H\mathbf{a}_j\right|^2+\mathbf{h}_{\mathrm{dl},k}^H\mathbf{V}\mathbf{h}_{\mathrm{dl},k}+\bar{G}+\sigma_k^2\right)+\sigma_{a,k}^2}\geq\gamma_k^{\mathrm{DL}},\forall k,$$

(24)

$$\frac{\left|\mathbf{h}_{\mathrm{dl},m}^H\mathbf{a}_k\right|^2}{\sum_{j\neq k}\left|\mathbf{h}_{\mathrm{dl},m}^H\mathbf{a}_j\right|^2+\mathbf{h}_{\mathrm{dl},m}^H\mathbf{V}\mathbf{h}_{\mathrm{dl},m}+\bar{G}+\sigma_m^2}\leq\gamma_e^k,\forall k,\forall m,$$

$$\eta_k(1-\rho_k)\left(\sum_{j=1}^{K}\left|\mathbf{h}_{\mathrm{dl},k}^H\mathbf{a}_j\right|^2+\mathbf{h}_{\mathrm{dl},k}^H\mathbf{V}\mathbf{h}_{\mathrm{dl},k}+\hat{G}+\sigma_k^2\right)\geq\bar{Q}_k,$$

$$\mathbf{V}\geq0,0<\rho_k<1,0<P_{\mathrm{up},k}\leq\min(\bar{Q}_k,P_{\max})$$

$$0<\|\mathbf{a}_k\|^2\leq P_{\max}.$$

Note that problem (23) corresponds to optimizing the variables involved in the uplink while problem (24) corresponds to those in the downlink. We then apply semidefinite relaxation (SDR) to the sub-problems as discussed below.

3 Proposed Solution

Clearly, even with the fixed set of $\gamma_e^k,\gamma_k^{\mathrm{DL}},\gamma_k^{\mathrm{UL}},\forall k$, problem (22) is non-convex. As a result of the possibility of decomposing problem (22) into two sub-problems, namely, problems (23) and (24), we will solve problem (23) analytically and use SDR technique to solve (24).

We first proceed by solving problem (23) to determine the optimal value of $P_{\mathrm{up},k}^*$ and $\mathbf{w}_k^*$.

The optimal receiver ($\mathbf{w}_k^*$) can be defined as the Wiener filter [28]

$$\mathbf{w}_k^*=\left(\sum_{j=1}^{K}P_{\mathrm{up},j}\mathbf{h}_{\mathrm{ul},j}\mathbf{h}_{\mathrm{ul},j}^H+\left[\sigma_j^2+\sum_{j=1}^{K}\|\mathbf{a}_j\|^2\right]\mathbf{I}\right)^{-1}\times\sqrt{P_{\mathrm{up},j}}\mathbf{h}_{\mathrm{ul},j}.$$

(25)

For given $\mathbf{w}_k^*$, the problem (23) can be reformulated as

$$\min_{P_{\mathrm{up},k}}\sum_{k=1}^{K}P_{\mathrm{up},k}$$

(26a)

$s.t.$

$$\frac{P_{\mathrm{up},k}\left|\mathbf{w}_k^H\mathbf{h}_{\mathrm{ul},k}\right|^2}{\sum_{j\neq k}^{K}P_{\mathrm{up},j}\left|\mathbf{w}_k^H\mathbf{h}_{\mathrm{ul},j}\right|^2+\bar{E}\|\mathbf{w}_k\|^2+\sigma_{\mathrm{BS}}^2\|\mathbf{w}_k\|^2}\geq\gamma_k^{\mathrm{UL}},$$

(26b)

$$0<P_{\mathrm{up},k}\leq\min(\bar{Q}_k,P_{\max}),\forall k.$$

(26c)

Upon investigation, the optimal $P_{\mathrm{up},k}^*$ is the minimum $P_{\mathrm{up},k}$

which satisfies (26b) to equality. Consequently, optimal $P_{\mathrm{up},k}^*$ is given by

$$P_{\mathrm{up}}^*=\frac{\gamma_k^{\mathrm{UL}}\left(\bar{E}\|\mathbf{w}_k\|^2+\sigma_{\mathrm{BS}}^2\|\mathbf{w}_k\|^2\right)}{\left|\mathbf{w}_k^H\mathbf{h}_{\mathrm{ul},k}\right|^2-\gamma_k^{\mathrm{UL}}\left(\sum_{j\neq k}^{K}\left|\mathbf{w}_k^H\mathbf{h}_{\mathrm{ul},j}\right|^2\right)}.$$

(27)

Then, by defining $\mathbf{A}_k=\mathbf{a}_k\mathbf{a}_k^H$, $\mathbf{H}_{\mathrm{dl},k}=\mathbf{h}_{\mathrm{dl},k}^H\mathbf{h}_{\mathrm{dl},k}$ and $\mathbf{G}_l=\mathbf{g}_l^H\mathbf{g}_l$, problem (24) can be equivalently rewritten as

$$\min_{\mathbf{A}_k,\mathbf{V},\rho_k\in(0,1)}\sum_{k=1}^{K}\mathrm{Tr}(\mathbf{A}_k)+\mathrm{Tr}(\mathbf{V})\quad s.t.$$

$$\frac{\mathrm{Tr}(\mathbf{G}_l\mathbf{A}_k)}{\sum_{j\neq k}\mathrm{Tr}(\mathbf{G}_l\mathbf{A}_j)+\mathrm{Tr}(\mathbf{G}_l\mathbf{V})+\sigma_l^2}\leq\gamma_e^k,\forall k,\forall l,$$

$$\frac{\rho_k\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{A}_k)}{\rho_k\left(\sum_{j\neq k}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{A}_j)+\mathrm{Tr}(\mathbf{H}_k\mathbf{V})+\bar{G}_k+\sigma_k^2\right)+\sigma_{a,k}^2}\geq\gamma_k^{\mathrm{DL}},\forall k,$$

(28)

$$\frac{\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},m}\mathbf{A}_k)}{\sum_{j\neq k}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},m}\mathbf{A}_j)+\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},m}\mathbf{V})+\bar{G}+\sigma_m^2}\leq\gamma_e^k,\forall k,\forall m,$$

$$\eta_k(1-\rho_k)\left(\sum_{j=1}^{K}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{A}_j)+\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{V})+\hat{G}_k+\sigma_k^2\right)\geq\bar{Q}_k,$$

$$\mathbf{V}\geq0,0<\|\mathbf{a}_k\|^2\leq P_{\max},0<P_{\mathrm{up},k}\leq\min(\bar{Q}_k,P_{\max}),$$

$$\mathrm{rank}(\mathbf{A}_k)=1,\forall k.$$

Note that problem (28) is still non-convex due to the non-convex rank-one constraints and the coupled optimization variables $\mathbf{A}_k$ and ρ_k. Applying SDR technique, after dropping the rank-one constraint, problem (28) can be reformulated into the following problem:

$$\min_{\mathbf{A}_k,\mathbf{V},\rho_k}\sum_{k=1}^{K}\mathrm{Tr}(\mathbf{A}_k)+\mathrm{Tr}(\mathbf{V})\quad s.t.$$

$$\frac{1}{\gamma_e^k}\mathrm{Tr}(\mathbf{G}_l\mathbf{A}_k)-\sum_{j\neq k}^{K}\mathrm{Tr}(\mathbf{G}_l\mathbf{A}_j)-\mathrm{Tr}(\mathbf{G}_l\mathbf{V})\leq\sigma_l^2,\forall k,\forall l,$$

$$\frac{1}{\gamma_k^{\mathrm{DL}}}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{A}_k)-\sum_{j\neq k}^{k}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{A}_j)-\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{V})-\bar{G}_k\geq\sigma_k^2+\frac{\sigma_{a,k}^2}{\rho_k},\forall k,$$

(29)

$$\frac{1}{\gamma_e^k}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},m}\mathbf{A}_k)-\sum_{j\neq k}^{k}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},m}\mathbf{A}_j)-\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},m}\mathbf{V})-\bar{G}_k\leq\sigma_m^2,\forall k,\forall m,$$

$$\sum_{j=1}^{K}\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{A}_j)+\mathrm{Tr}(\mathbf{H}_{\mathrm{dl},k}\mathbf{V})+\hat{G}_k\geq\frac{\bar{Q}_k}{\eta_k(1-\rho_k)}-\sigma_k^2,\forall k,$$

$$\mathbf{V}\geq0,\mathbf{A}_k\geq0,0<\rho_k\leq1,0<P_{\mathrm{up},k}\leq\min(\bar{Q}_k,P_{\max}).$$

Clearly, from (29), both $1/\rho_k$ and $1/(1-\rho_k)$ are convex functions over ρ_k, so that problem (29) is convex and can be solved using a disciplined convex programming toolbox such as CVX [29]. Now, let $\mathbf{A}_k^*$, $\forall k$ denotes the optimal solution to problem (29). If each $\mathbf{A}_k^*$ is rank-one, it follows that the optimal solution of problem (24) can be obtained through eigenval-

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ue decomposition of $\mathbf{A}_k^*$. Interestingly, as shown in Proposition 1, problem (24) always yields rank-one optimal solutions for $\mathbf{A}_k^*, \forall k$.

Proposition 1: There always exists an optimal solution A_k^* for problem (29) such that rank $(\mathbf{A}_k^*) = 1, \forall k$.

To summarise, Proposition 1 implies that the relaxation is tight, and thus the globally optimal solution of problem (24) can be obtained.

4 Numerical Examples

In this section, numerical examples are provided. In particular, the performance of the proposed joint beamforming design (Joint Opt in **Fig. 2**) for secure SWIPT in MISO FD systems is evaluated through simulations. We simulate a flat Rayleigh fading environment in which the channel fading coefficients are characterized as complex Gaussian numbers with zero mean and are independent and identically distributed, and we assume there are $K = 2$ MSs and all MSs have the same set of parameters, i.e., $\sigma_k^2 = \sigma^2$, $\delta_k^2 = \delta^2$, $\bar{Q}_k = Q$, $\gamma_k^{DL} = \gamma^{DL}$, $\gamma_k^{UL} = \gamma^{UL}$, and $y_E = -15$ dB. We also assume that 70% of the SI power has been cancelled using existing SI cancellation techniques [14]. All simulations are averaged over 1000 independent channel realizations.

In Fig. 2, we investigate the achievable secrecy sum-rate for secure SWIPT in MISO FD system versus the uplink SINR target for all MSs for fixed values of the downlink SINR ($\gamma^{DL} = -5$ dB), the harvested power threshold, and the eavesdropper's SINR ($y_E = -15$ dB). Fig. 2 shows the performance comparison between the joint optimization and the fixed ρ optimization. As can be observed, an increase in the uplink SINR corresponds to a decrease in the secrecy sum-rate due to increase in transmission power which invariably increases the generated SI. However, the joint optimization scheme outperforms the fixed ρ optimization. Also, as the harvested power threshold increases, the secrecy sum-rate decreases, but the joint optimization scheme outperforms the fixed ρ optimization scheme.

In **Fig. 3**, we further investigate the achievable secrecy sum-rate for secure

SWIPT in MISO FD system versus the harvested power threshold for given values of the uplink and downlink SINR. As can be observed, as the harvested power threshold increases, the secrecy sum rate increases. Technically, an increase in the harvested power allows more power to be directed to the ER of each MS, thus less power for SI, which invariably increases the secrecy sum-rate.

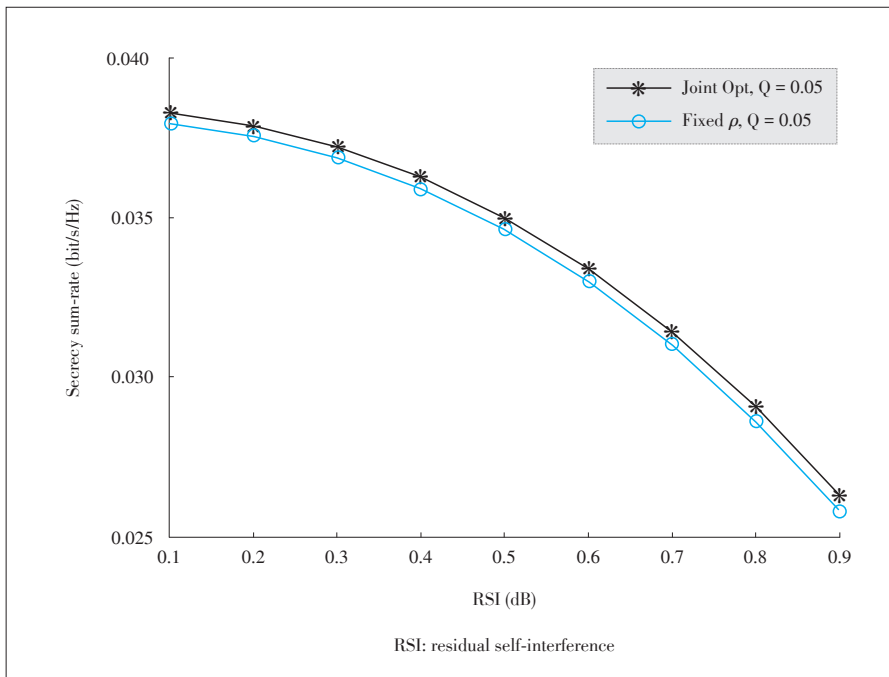In **Fig. 4**, the achievable secrecy sum-rate for secure
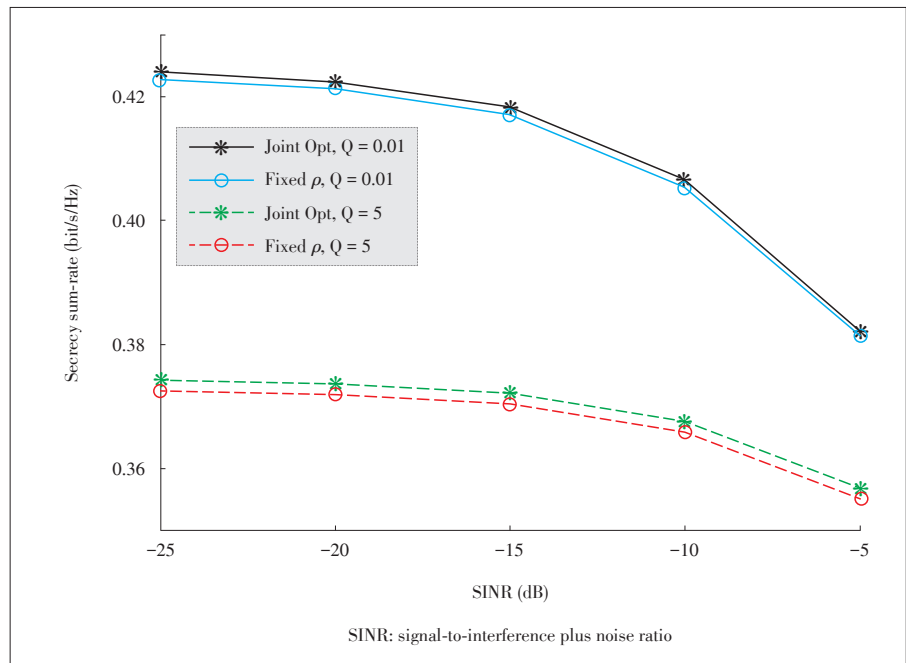
▲Figure 2. Secrecy sum-rate versus SINR γ^{UL}.
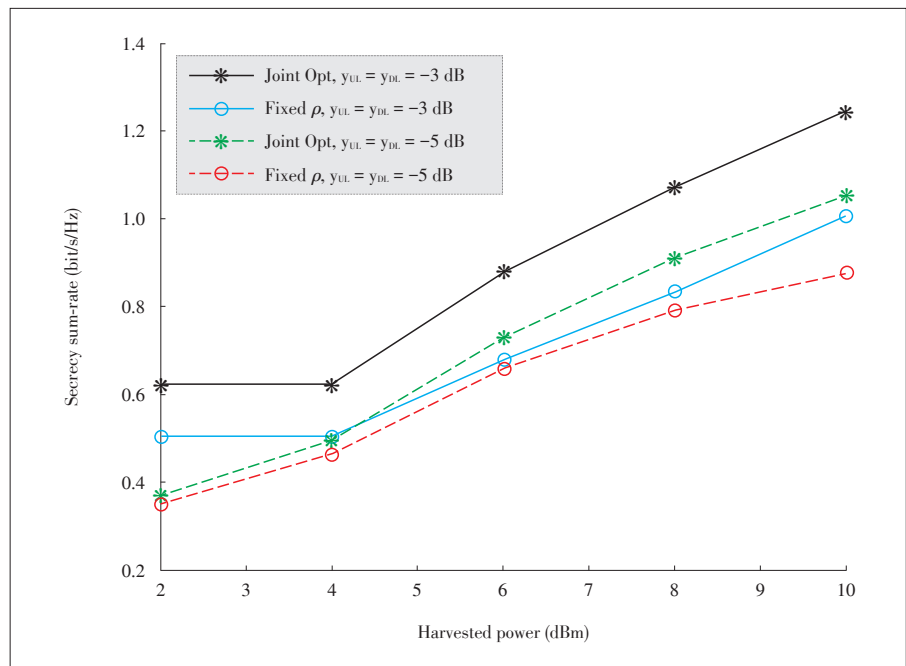
▲Figure 3. Secrecy sum-rate versus harvested power.

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▲Figure 4. Secrecy sum-rate versus RSI.

SWIPT in MISO FD system versus the residual SI is investigated. In particular, we investigate the performance comparison of the proposed system for both joint optimization, and optimization with a fixed ρ. Clearly, an increase in the RSI results in a corresponding decrease in the achievable secrecy rate. However, the joint optimization scheme outperforms the optimization scheme with fixed ρ. Thus, a need for joint optimization is justified.

5 Conclusions

In this paper, we investigated the secure beamforming design for SWIPT in MISO FD system. To reduce SI and achieve secure FD transmission, the end-to-end sum transmit power has been minimized subject to the given SINR and harvested power constraints for each MS. In particular, to achieve optimal system performance, using SDR technique, we jointly optimized the transmit beamforming vector at the BS, and the received PS ratio and the transmit power at the MSs, and the AN covariance matrix. We show the need for joint optimization of system parameters through the numerical results.

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Biographies

Alexander A. Okandeji (alexander.okandeji@iuokada.edu.ng) received the B.Sc. degree (First class honours) from Olabisi Onabanjo University, Ogun State, Nigeria in 2010, M.Sc. degree from Loughborough University, UK in 2013, and Ph.D. degree from the Department of Electronic and Electrical Engineering, University College London, UK in 2017. His Ph.D. was funded by the Federal Government of Nigeria under the presidential special scholarship scheme for innovation and development (PRESSID) scheme. Dr. Okandeji presently lectures at the Department of Electrical and Computer Engineering, Igbinedion University, Nigeria where he teaches control systems engineering, electromagnetic field and wave theory, communication systems, telecommunication principle, and reliability and maintainability of systems. His research interests include full-duplex radio, MIMO, energy harvesting wireless communications, physical layer security, digital signal processing, and cloud computing.

Muhammad R. A. Khandaker (m.khandaker@ucl.ac.uk) received his Ph.D. degree in electrical and computer engineering from Curtin University, Australia in 2013. He has held a number of academic positions in Bangladesh. Since 2013, he has been a post-doctoral researcher with the Department of Electronic and Electrical Engineering, University College London, UK. He received the Curtin International Postgraduate Research Scholarship for his Ph.D. study in 2009. He received the Best Paper Award at the 16th IEEE Asia-Pacific Conference on Communications, Auckland, New Zealand in 2010. He regularly serves in the technical program committees of IEEE conferences, including Globecom, ICC, and VTC. He is currently serving as an associate editor as well as the lead guest editor of the *EURASIP Journal on Wireless Communications and Networking* Special Issue on Heterogeneous Cloud Radio Access Networks. He also served as the managing guest editor of the *Physical Communication* (Elsevier) Special Issue on Self-Optimizing Cognitive Radio Technologies.

WONG Kai-Kit (kai-kit.wong@ucl.ac.uk) received the B.Eng., M.Phil., and Ph.D. degrees from The Hong Kong University of Science and Technology, China in 1996, 1998, and 2001, respectively, all in electrical and electronic engineering. He is currently a professor of wireless communications with the Department of Electronic and Electrical Engineering, University College London, UK. He is a fellow of IET. He is a senior editor of the *IEEE Communications Letters* and the *IEEE Wireless Communications Letters*.

ZHANG Yangyang (yangyang.zhang@kuang-chi.org) received the B.S. and M.S. degrees in electronics and information engineering from Northeastern University, China in 2002 and 2004 respectively, and the Ph.D. degree in electrical engineering from the University of Oxford, UK in 2008. He is currently with the Shenzhen Key Laboratory of Artificial Microstructure Design, Guangdong Key Laboratory of Meta-RF Microwave Radio Frequency and Kuang-Chi Institute of Advanced Technology, China. From 2008 to 2010, he was a postdoctoral research fellow with the University College London, UK. He is currently the Executive Vice President with the Kuang-Chi Institute of Advanced Technology, China. His research interests are mainly focused on metamaterial-based future wireless communication system, such as MIMO communication system, metamaterial-based RF devices, and metamaterial-based spatial modulation technology. He received over 20 honors from various national and international competitions and published around 40 papers in various journals and conferences.

ZHENG Zhongbin (ben@ecit.org.cn) received the bachelor's and master's degrees in information and communications engineering from Beijing University of Posts and Telecommunications, China in 2002 and 2005, respectively. He is currently the Vice Director of the China Academy of Information and Communications Technology and the East China Institute of Telecommunications. He was also the former head of the Technology Department for the East China Institute of the Ministry of Industry and Information Technology, China. He is very active in research, with a number of international paper publications as well as patents and draft standards.

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