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Secure Beamforming for IRS-Assisted Nonlinear SWIPT Systems with Full-Duplex User

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***Abstract*—Physical layer security of wireless communication network with intelligent reflecting surface (IRS) is investigated in this letter. At the access point (AP) transmission power and user nonlinear energy harvest constraints, maximum secrecy rate of full-duplex (FD) user which employs simultaneous wireless information and power transfer (SWIPT) technology is achieved by jointly optimizing transmit beamformer of AP, IRS phase shift, and artificial noise power of FD user. Due to its undesirable properties, such as coupling of multiple optimization variables and non-convex objective function, original optimization problem is hard to be handled. We solved this problem by an alternating optimization algorithm consisting of semi-definite relaxation, Charnes-Cooper transform and continuous phase shifting discretization techniques to obtain a feasible suboptimal solution. Simulation results demonstrate that although may only guarantee to convergence a local optimal, proposed algorithm still achieves the best security rate comparing to the benchmark algorithms with the computational efficiency sacrificing.**

***Index Terms*—Intelligent reflecting surface, physical layer security, SWIPT, full-duplex user, artificial noise**

I. Introduction

GREEN, cost-effective and security are the basic requirements of the next generation wireless communication technology [1]. Simultaneous wireless information and power transfer (SWIPT), which can harvests both energy and information from radio frequency signals to extend the lifetime of devices, is considered as the crucial technologies to addressing those basic requirements [2]. Meanwhile, intelligent reflecting surface (IRS) technology, which can independently changes the phase of the incident signal at the each element of IRS without energy consumption, is also introduced in the wireless communication network to replace the relay for energy conservation recently [3].

With the spread using IRS and SWIPT together in communication network, its physical layer security (PLS) has attracted the wide research attention. Using artificial noise (AN) generated by relay to confuse eavesdropper may be an optional scheme to improve PLS in traditional relay-assisted network. However, due to its passive property, IRS is unable to generate AN, thus the PLS scheme rely on relay generating AN directly extend in IRS-assisted networks are unsuitable. To circumvent this problem, base station or friendly jammer is introduced to generate AN [2], however the AN generated by base station or friendly jammer will interfere with legitimate users (UE) too [4], thus the signal-interference-noise-ratio (SINR) of the UE will be reduced, lead decoding complexity of UE to increasing.

Using full duplex cooperative jamming (FD-CJ) ,namely, FD UE employing SWIPT to receive signal and to harvest energy meanwhile generating AN to confuse eavesdropper,may be an optional scheme to handle above problem since the UE can employ self-interference cancellation (SIC) to cancel the interference of themselves. Thus, the SINR of the UE can be enhanced, resulting in the PLS of wireless communication network improving [5].

For instance, X. Xuan demonstrates that employing FD-CJ scheme can reduce the secrecy outage probability of the Peer-to-Peer SWIPT network [6]. W. Li shows that using FD-CJ scheme can effectively improve the security rate of the relay-assisted network [7]. Since employing linear EH models and using relay to forwarding signal, their works can not extend to practical scenarios in which the nonlinear energy harvest (EH) models of SWIPT and the IRS with discrete phase shift are employed [8][9].

The PLS of IRS-assisted SWIPT communication network is mainly concerned in this letter. Specifically, FD UE first employs SWIPT technology with nonlinear EH model to receive signal and to harvest energy and then uses the FD-CJ scheme to achieve the maximum security rate of the wireless communication network on the constraints including IRS phase-shift, AN power of FD user and transmitting power of AP. Since the original FD-CJ scheme has the undesirable properties of coupling including multiple optimization variables and nonconvex objective functions, obtaining its optimal solution is impractical, an iterative alternating optimization algorithm consisting mainly of semidefinite relaxation (SDR), Charnes-Coopers transform and continuous phase shift discretization is proposed here to obtain the suboptimal solution of the original problem. Simulations show the effectiveness and superiority of the proposed algorithm.

Notations: Column vectors and matrices are denoted by boldface lower case letters and capital letters, respectively. The Hermitian conjugate transpose, rank and trace of a matrix $\boldsymbol{A}$ are denoted by $\boldsymbol{A}^H$, $\mathrm{Rank}\left(\boldsymbol{A}\right)$ and $\mathrm{Tr}\left(\boldsymbol{A}\right)$, respectively. The Euclidean norm of a vector and the modulus of a complex scalar are denoted by $\|\bullet\|$ and $|\bullet|$, respectively. $\mathrm{diag}\left(\boldsymbol{c}\right)$ is the diagonal matrix with the vector $\boldsymbol{c}$ on the diagonal. $\mathbb{E}\left(\bullet\right)$ denotes the expectation operation.

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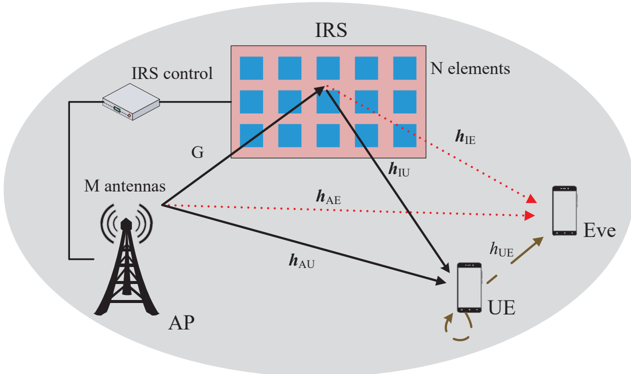

Fig. 1. IRS-assisted SWIPT secure communication network

II. System model and problem formulation

An IRS-assisted wireless communication network is shown in Fig.1, in which FD UE employs SWIPT to receive signal transmitted by AP and to harvest energy as well, moreover, transmits AN to confuse eavesdropper (Eve) at the same time. We can find that the wireless communication network is composed of one access point (AP) with a M-antennas uniform linear array (ULA) , one IRS with a N-elements uniform rectangular array (URA), one UE with a dual separate antenna and one eavesdropper (Eve) with a single antenna. The channel between AP and IRS, AP and UE, AP and Eve, IRS and UE, IRS and Eve as well as UE and Eve can be expressed as $\boldsymbol{G} \in \mathbb{C}^{M\times N}$, $\boldsymbol{h}_{AU} \in \mathbb{C}^{M}$, $\boldsymbol{h}_{AE} \in \mathbb{C}^{M}$, $\boldsymbol{h}_{IU} \in \mathbb{C}^{N}$, $\boldsymbol{h}_{IE} \in \mathbb{C}^{N}$ and $h_{UE} \in \mathbb{C}$, respectively. Suppose Eve is the legitimate UE of previous time slot, thus channel between AP and Eve may be obtained perfectly [10]. Moreover, some assumptions including frequency-flat fading of all channels, all channel state information (CSI) between AP and UE fully known by AP [1], the signal of reflected by IRS two more times negligible are satisfied [11].

In data communication, the FD UE divides the received signal into two parts according to power splitting SWIPT protocol; one is used for information decoding (ID), which can be expressed as $y_{ID} = \sqrt{\rho} y_U + n_{ID}$, where $n_{ID} \sim \mathcal{CN}\left(0, \sigma_{ID}^2\right)$ is the complex additive white Gaussian noise (AWGN) introduced by the ID circuit; the other is used to EH and then harvested energy which can be expressed as $y_{EH} = \sqrt{1-\rho} y_U$ is used to transmit AN, in which y_U described in (1) denotes the signal received by UE after SIC.

$$y_U = \left(\boldsymbol{h}_{IU}^H \boldsymbol{\Theta} \boldsymbol{G} + \boldsymbol{h}_{AU}^H\right) \boldsymbol{w} s + n_{SI} + n_U \tag{1}$$

where $\boldsymbol{\Theta} = \operatorname{diag}\left(v_1, ..., v_n\right)$ represents the IRS phase-shift matrix, $v_i = \beta_i e^{j\theta_i}$, $\beta \in [0,1]$ is the amplitude reflection coefficient and be set to 1 for simplicity. θ_i is uniformly quantized in the interval $[0, 2\pi)$, i.e., $\theta_i \in \Phi \triangleq \left\{0, 2\pi/2^L, ..., 2\pi\left(L-1\right)/2^L\right\}$, L represents the control bits of phase shifts for each IRS element, $L=\infty$ means continuous phase shift. $\boldsymbol{w} \in \mathbb{C}^M$ denotes the transmit beamforming vector at AP, s denotes the transmitted signal of AP, which satisfies $\mathbb{E}\left(\|s\|^2\right) = 1$. $n_{SI} \sim \mathcal{CN}\left(0, \sigma_{SI}^2\right)$ represents the residual self-interference, $n_U \sim \mathcal{CN}\left(0, \sigma_U^2\right)$ represents the complex AWGN received by UE.

[1]While perfect CSI in practical scenarios may be difficult to be achieved, the results of this paper can be served as the upper bound of the FD-CJ scheme.

The splitted power at UE can by expresed as $E_U = (1-\rho)\,\mathbb{E}\left\{|y_U|^2\right\}$. Similar as [9], the nonlinear EH model with the sigmoidal function is introduced here. Thus, the total harvested energy at the UE can be modeled as

$$\phi_U^{NL} = \frac{\psi_U^{NL} - E_M \Omega}{1-\Omega}, \Omega = \frac{1}{1+\exp\left(a_U b_U\right)}, \tag{2a}$$

$$\psi_U^{NL} = \frac{E_M}{1+\exp\left(-a_U\left(E_U - b_U\right)\right)}, \tag{2b}$$

where ψ_U^{NL} is the traditional logistic function and Ω is a constant to guarantee a zero input/output response. In addition, a_U and b_U are constants related to the circuit characteristics. E_M is a constant which represents the maximum energy that can be harvested by the user.

Due to the additional energy consumption related to transmitting AN, the constraint $\sigma_{AN}^2 \leqslant \phi_U^{NL}$ need to be satisfied.

Note that since Ω is a constant and irrelevants with solving the optimization problem, so ψ_U^{NL} is directly employed to describe the harvested power at UE. Hence, the inverse function of (2b) can be written as

$$E_U\left(\psi_U^{NL}\right) = b_U - \frac{1}{a_U}\ln\left(\frac{E_M}{\psi_U^{NL}} - 1\right). \tag{3}$$

Supposing harvested power at UE is entirely utilized to generate AN, namely $\sigma_{AN}^2 = \phi_U^{NL}$, energy consumption constraint can be transformed into

$$E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\left|\left(\boldsymbol{h}_{IU}^H \boldsymbol{\Theta} \boldsymbol{G} + \boldsymbol{h}_{AU}^H\right)\boldsymbol{w}\right|^2 \tag{4}$$

in which $(1-\rho)\left|\left(\boldsymbol{h}_{IU}^H \boldsymbol{\Theta} \boldsymbol{G} + \boldsymbol{h}_{AU}^H\right)\boldsymbol{w}\right|^2$ means the energy which send to the UE by AP beamformer and IRS reflection. Another, the signal received by the Eve is given by

$$y_E = \left(\boldsymbol{h}_{IE}^H \boldsymbol{\Theta} \boldsymbol{G} + \boldsymbol{h}_{AE}^H\right) \boldsymbol{w} s + h_{UE} s_{AN} + n_E, \tag{5}$$

where $s_{AN} \sim \mathcal{CN}\left(0, \sigma_{AN}^2\right)$ represents the AN signal transmitted by UE, $n_E \sim \mathcal{CN}\left(0, \sigma_E^2\right)$ represents the complex AWGN received by Eve.

So far, the SINR of UE and Eve can be defined as (6) and (7), respectively.

$$SINR_U = \frac{\left|\left(\boldsymbol{h}_{IU}^H \boldsymbol{\Theta} \boldsymbol{G} + \boldsymbol{h}_{AU}^H\right)\boldsymbol{w}\right|^2}{\sigma_{SI}^2 + \sigma_U^2 + \sigma_{ID}^2/\rho}, \tag{6}$$

$$SINR_E = \frac{\left|\left(\boldsymbol{h}_{IE}^H \boldsymbol{\Theta} \boldsymbol{G} + \boldsymbol{h}_{AE}^H\right)\boldsymbol{w}\right|^2}{\left|h_{UE}\right|^2 \sigma_{AN}^2 + \sigma_E^2}. \tag{7}$$

Similar to [4], the security rate is defined as

$$R_{Sec} = \log_2\left(1 + SINR_U\right) - \log_2\left(1 + SINR_E\right). \tag{8}$$

To ensure secure communications in IRS-assisted wireless communication network, our proposed scheme of maximize

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the security rate on the constraints including AP transmit power, IRS shift phase and power of AN is described as (9)

$$\max_{\boldsymbol{w},\boldsymbol{\Theta},\sigma_{AN}^2} \quad R_{Sec} \tag{9a}$$

$$\text{s.t.} \quad \|\boldsymbol{w}\|^2 \leqslant P, \tag{9b}$$

$$\theta \in \Phi \triangleq \left\{0, 2\pi/2^L, ..., 2\pi\left(L-1\right)/2^L\right\} \tag{9c}$$

$$E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\left|\left(\boldsymbol{h}_{IU}^H\boldsymbol{\Theta}\boldsymbol{G}+\boldsymbol{h}_{AU}^H\right)\boldsymbol{w}\right|^2, \tag{9d}$$

where P denotes the threshold of AP transmission power.

Obviously, the (9) is a mixed-integer non-convex optimization problem with the coupling of multiple optimized variables. We propose an iterative algorithm based on alternating optimization (AO) to tackle this problem in the following section.

III. The Optimization algorithm

A. Optimizing AP Beamforming Vector with given $\boldsymbol{\Theta}$ and σ_{AN}^2

Given $\boldsymbol{\Theta}$ and σ_{AN}^2, the problem (9) can be equivalently recasted as the following fractional programming problem

$$\begin{aligned} \max_{\boldsymbol{w}} \quad & \frac{\sigma_{AU}^2+\left|\boldsymbol{g}_{AU}^H\boldsymbol{w}\right|^2}{\sigma_{AE}^2+\left|\boldsymbol{g}_{AE}^H\boldsymbol{w}\right|^2} \\ \text{s.t.} \quad & \|\boldsymbol{w}\|^2 \leqslant P \\ & E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\left|\boldsymbol{g}_{AU}^H\boldsymbol{w}\right|^2, \end{aligned} \tag{10}$$

where $\boldsymbol{g}_{AU}^H = \boldsymbol{h}_{IU}^H\boldsymbol{\Theta}\boldsymbol{G}+\boldsymbol{h}_{AU}^H$, $\boldsymbol{g}_{AE}^H = \boldsymbol{h}_{IE}^H\boldsymbol{\Theta}\boldsymbol{G}+\boldsymbol{h}_{AE}^H$, $\sigma_{AU}^2 = \sigma_{SI}^2+\sigma_U^2+\sigma_{ID}^2/\rho$ and $\sigma_{AE}^2 = |h_{UE}|^2\sigma_{AN}^2+\sigma_E^2$. Although to be a single vector optimization, problem (10) is still hard to be solved due to the fractional and quadratic forms of the objective function. We define $\bar{\boldsymbol{W}} = \boldsymbol{w}\boldsymbol{w}^H$ to transform problem into

$$\begin{aligned} \max_{\bar{W},\geqslant 0} \quad & \frac{\sigma_{AU}^2+\mathrm{Tr}\left(\boldsymbol{F}_{AU}\bar{\boldsymbol{W}}\right)}{\sigma_{AE}^2+\mathrm{Tr}\left(\boldsymbol{F}_{AE}\bar{\boldsymbol{W}}\right)} \\ \text{s.t.} \quad & \mathrm{Tr}\left(\bar{\boldsymbol{W}}\right) \leqslant P \\ & \bar{\boldsymbol{W}} \succeq 0, \mathrm{Rank}\left(\bar{\boldsymbol{W}}\right) = 1 \\ & E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\,\mathrm{Tr}\left(\boldsymbol{F}_{AU}\bar{\boldsymbol{W}}\right), \end{aligned} \tag{11}$$

where $\boldsymbol{F}_{AE} = \boldsymbol{g}_{AE}\boldsymbol{g}_{AE}^H$, $\boldsymbol{F}_{AU} = \boldsymbol{g}_{AU}\boldsymbol{g}_{AU}^H$.

Although we reformulate the quadratic form of the problem (10) into the problem (11), the fractional form still exists in the objective function of problem (11), thus difficult to handle it too. We use the Charnes-Cooper transformation and define $\mu = \left(Tr\left(\boldsymbol{F}_{AE}\tilde{\boldsymbol{W}}\right)+\sigma_{AE}^2\right)^{-1}$ and $\boldsymbol{W} = \mu\bar{\boldsymbol{W}}$, rewrite (11) in the following form

$$\begin{aligned} \max_{\boldsymbol{W},\mu\geqslant 0} \quad & \mu\sigma_{AU}^2+\mathrm{Tr}\left(\boldsymbol{F}_{AU}\boldsymbol{W}\right) \\ \text{s.t.} \quad & \mathrm{Tr}\left(\boldsymbol{W}\right) \leqslant \mu P \\ & \boldsymbol{W} \succeq 0, \mathrm{Rank}\left(\boldsymbol{W}\right) = 1 \\ & \mu\sigma_{AE}^2+\mathrm{Tr}\left(\boldsymbol{F}_{AE}\boldsymbol{W}\right) = 1 \\ & \mu E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\,\mathrm{Tr}\left(\boldsymbol{F}_{AU}\boldsymbol{W}\right), \end{aligned} \tag{12}$$

Relaxing the constraint $\mathrm{Rank}\left(\boldsymbol{W}\right) = 1$, (12) can be solved efficiently by the interior point method (e.g. CVX [12]). Once relaxed solution $\boldsymbol{W}^*$ of (12) is obtained, the beamforming vector $\boldsymbol{w}^*$ can be obtained by eigenvalue decomposition of $\boldsymbol{W}^*$ or Gaussian randomization technique [13].

B. Optimizing IRS Phase-shift with given $\boldsymbol{w}$ and σ_{AN}^2

Given $\boldsymbol{w}$ and σ_{AN}^2, the original problem can be reformulated as a discrete optimization problem. We can relax discrete constraints of optimized variable to obtain continuous solution firstly, and then discretize this continuous solution to achieve the solution of the original problem.

According to the problem (9), with given $\boldsymbol{w}$ and σ_{AN}^2, the relax continuous version of original problem can be described as.

$$\max_{\boldsymbol{q}} \quad \frac{\boldsymbol{q}^H\boldsymbol{H}_U\boldsymbol{q}+g_u+1}{\boldsymbol{q}^H\boldsymbol{H}_E\boldsymbol{q}+g_E+1} \tag{13a}$$

$$\text{s.t.} \quad |q_n| = 1, n \in N+1 \tag{13b}$$

$$E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\,\sigma_{AU}^2\left(\boldsymbol{q}^H\boldsymbol{H}_U\boldsymbol{q}+g_U\right), \tag{13c}$$

where $\boldsymbol{q} = [\mathrm{diag}\left(\boldsymbol{\Theta}\right);1]$, $g_U = \boldsymbol{h}_{AU}^H\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{h}_{AU}/\sigma_{AU}^2$, $g_E = \boldsymbol{h}_{AE}^H\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{h}_{AE}/\sigma_{AE}^2$, $\boldsymbol{H}_U$ and $\boldsymbol{H}_E$ are described in (14) and (15), respectively.

It is clearly that objective function and constraint in (13) are still non-convexity. By defining $\boldsymbol{Q} = \tau\boldsymbol{q}\boldsymbol{q}^H$, $\tau = \left(\mathrm{Tr}\left(\boldsymbol{H}_E\boldsymbol{Q}\right)+g_E+1\right)^{-1}$ and using Charnes-Cooper transformation again, The problem (13) can be re-expressed as

$$\begin{aligned} \max_{\boldsymbol{Q},\tau\geqslant 0} \quad & \mathrm{Tr}\left(\boldsymbol{H}_U\boldsymbol{Q}\right)+\tau\left(g_U+1\right) \\ \text{s.t.} \quad & \boldsymbol{Q} \succeq 0, \mathrm{Rank}\left(\boldsymbol{Q}\right) = 1 \\ & \mathrm{Tr}\left(\boldsymbol{E}_n\boldsymbol{Q}\right) = \tau, \forall N, \\ & \mathrm{Tr}\left(\boldsymbol{H}_E\boldsymbol{Q}\right)+\tau\left(g_E+1\right) = 1 \\ & \tau E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\,\sigma_{AU}^2\left(\mathrm{Tr}\left(\boldsymbol{H}_U\boldsymbol{Q}\right)\right), \end{aligned} \tag{16}$$

where $\boldsymbol{E}_n$ is the zero matrix with the n-th diagonal element 1. Obviously,the problem (16) can be solved efficiently by relaxing the constraint $\mathrm{Rank}\left(\boldsymbol{Q}\right) = 1$. Also using the eigenvalue decomposition or Gaussian randomization technique, we obtain the optimal solution of continuous phase shift $\boldsymbol{\theta}^{C*}$, the optimal discrete phase shift $\boldsymbol{\theta}^{P*}$ can be obtained by solving the following problem

$$\theta_i^{P*} = \arg\min_{\theta\in\Phi} \quad \left|\theta-\theta_i^C\right|, \forall N, \tag{17}$$

where θ_i^{P*} and θ_i^C represent the i-th element of $\boldsymbol{\theta}^{P*}$ and $\boldsymbol{\theta}^{C*}$, respectively.

C. Optimizing AN transmit power with given $\boldsymbol{w}$ and $\boldsymbol{\Theta}$

Given $\boldsymbol{w}$ and $\boldsymbol{\Theta}$, the problem (9) can be equivalently reformulated into

$$\max_{\sigma_{AN}^2} \quad \log_2\left(1+D_1\right)-\log_2\left(1+\frac{D_2}{D_3+\sigma_{AN}^2}\right) \tag{18a}$$

$$\text{s.t.} \quad E_U\left(\sigma_{AN}^2\right) \leqslant (1-\rho)\left|\boldsymbol{g}_{AU}^H\boldsymbol{w}\right|^2, \tag{18b}$$

where $D_1 = \left|\boldsymbol{g}_{AU}^H\boldsymbol{w}\right|^2/\sigma_{AU}^2$, $D_2 = \left|\boldsymbol{g}_{AE}^H\boldsymbol{w}\right|^2/|h_{UE}|^2$ and $D_3 = \sigma_E^2/|h_{UE}|^2$.

Set $y = \log_2\left(1+D_1\right)-\log_2\left(1+\frac{D_2}{D_3+\sigma_{AN}^2}\right)$ according to objection function of (18a), the first-order derivative of y with respect to σ_{AN}^2 can be described as $\frac{\mathrm{d}y}{\mathrm{d}\sigma_{AN}^2} = D_2\left(2\ln 2\left[\left(D_3+\sigma_{AN}^2\right)^2+D_2\left(D_3+\sigma_{AN}^2\right)\right]\right)^{-1}$, and it is easy to verify that $\frac{\mathrm{d}y}{\mathrm{d}\sigma_{AN}^2} > 0$ always holds, which indicates that y is monotonically increasing. Due to constraint (18b), the optimal AN transmit power can be achieved at $E_U\left(\sigma_{AN}^2\right) =$

$$\boldsymbol{H}_U = \frac{1}{\sigma_{AU}^2}\begin{bmatrix} \operatorname{diag}\left(\boldsymbol{h}_{IU}^H\right)\boldsymbol{G}\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{G}^H\operatorname{diag}\left(\boldsymbol{h}_{IU}^H\right)^H & \operatorname{diag}\left(\boldsymbol{h}_{IU}^H\right)\boldsymbol{G}\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{h}_{AU} \\ \boldsymbol{h}_{AU}^H\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{G}^H\operatorname{diag}\left(\boldsymbol{h}_{IU}^H\right)^H & 0 \end{bmatrix}, \tag{14}$$

$$\boldsymbol{H}_E = \frac{1}{\sigma_{AE}^2}\begin{bmatrix} \operatorname{diag}\left(\boldsymbol{h}_{IE}^H\right)\boldsymbol{G}\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{G}^H\operatorname{diag}\left(\boldsymbol{h}_{IE}\right)^H & \operatorname{diag}\left(\boldsymbol{h}_{IE}^H\right)\boldsymbol{G}\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{h}_{AE} \\ \boldsymbol{h}_{AE}^H\boldsymbol{w}\boldsymbol{w}^H\boldsymbol{G}^H\operatorname{diag}\left(\boldsymbol{h}_{IE}^H\right)^H & 0 \end{bmatrix}. \tag{15}$$

$(1-\rho)\left|\boldsymbol{g}_{AU}^H\boldsymbol{w}\right|^2$ and the optimal of (18) can be expresses as (19)

$$\sigma_{AN}^{2\ *} = \frac{E_M}{1+\exp\left(a_U\left(b_U - \left|\boldsymbol{g}_{AU}^H\boldsymbol{w}\right|^2\right)\right)}, \tag{19}$$

Details of the proposed algorithm is summarized in Algorithm 1.

Algorithm 1 Proposed algorithm for solving (9)

1: **Initialization :** Set $\boldsymbol{w}^0 = \boldsymbol{w}^{MRT}$, $\boldsymbol{q}^0 = \mathbf{1}$, $\sigma_{AN}^{2\ 0} = 0$dBm, $R_{Sec}^0 = 0$, $\varepsilon = 10^{-3}$ and $k = 0$
2: **Repeat :**
3: $k = k + 1$
4: Solve problem (12) and return $\boldsymbol{W}^k$, apply EVD or Gaussia randomization to obtain $\boldsymbol{w}^k$
5: Solve problem (16) and return $\boldsymbol{Q}^k$, get discrete phase shift $\boldsymbol{\theta}^{P*}$ by (17)
6: Get $\sigma_{AN}^{2\ k}$ by (19)
7: Get secrecy rate R_{Sec}^k by (8)
8: **Until :** $\left|R_{Sec}^k - R_{Sec}^{k-1}\right| \leqslant \varepsilon$

D. Complexity Analysis

It can be seen from Algorithm 1 that the computational complexity of our algorithm mainly stems from the solving (12) and (16).

According to the [9], the complexity of a SDP problem including j SDP constraints each with an $k \times k$ positive semi-definite matrix can be expressed as $\mathcal{O}\left(\sqrt{k}\log\left(1/\varepsilon\right)\left(jk^3+j^2k^2+j^3\right)\right)$, where ε is accuracy threshold. Thus the complexity of solving (12) can be approximately expressed as $\mathcal{O}\left(\log\left(1/\varepsilon\right)\left(M^{3.5}\right)\right)$; Similarly, the complexity of solving (16) can be approximately expressed as $\mathcal{O}\left(\log\left(1/\varepsilon\right)\left(N^4\sqrt{N+1}\right)\right)$. To this end, the complexity of the whole iterative algorithm 1 can be simplified as $\mathcal{O}\left(\ln\left(1/\varepsilon\right)\left(M^{3.5}+N^4\sqrt{N+1}+N\right)\right)$.

IV. Simulation Results and Analysis

In this section, we perform simulation experiments to evaluate the effectiveness of the proposed algorithm. The simulation scenario is as depicted in Fig. 1, and the simulation parameters refer to Table I. In addition, assuming that all channels experience the Rician fading, i.e.

$$\boldsymbol{h} = \sqrt{C_0(d)^{-\xi}}\left(\sqrt{\frac{\kappa}{\kappa+1}}\boldsymbol{h}^{\mathrm{Los}} + \sqrt{\frac{1}{\kappa+1}}\boldsymbol{h}^{\mathrm{NLos}}\right). \tag{20}$$

TABLE I
Parameters of Simulation

Parameters	Notations	Typical Values
M	Number of AP antennas	8
N	Number of IRS elements	32
C_0	Fade coefficient	10^{-3}
ξ	Path loss exponent	2.5
κ	Rician fading factor	10
σ_U^2	Noise power of UE	-80dBm
σ_E^2	Noise power of Eve	-80dBm
d_{AI}	Distance between AP and IRS	25m
d_{AU}	Distance between AP and UE	30m
d_{AE}	Distance between AP and Eve	40m
d_{IU}	Distance between IRS and UE	25m
d_{IE}	Distance between IRS and Eve	28m

We evaluate the performance of proposed algorithm by comparing with the following benchmark algorithm.

A. Random phase shift, namely in the step 5 of algorithm 1, the optimal discrete phase shift $\boldsymbol{\theta}^{P*}$ is obtained by random generation instead of solving (16).

B. Maximum ratio transmission (MRT), namely, in the step 4 of algorithm 1, the optimal beamforming vector $\boldsymbol{w}$ is obtained by MRT algorithm rather than by solving (12).

C. Without IRS. IRS is not employed in the proposed algorithm, namely, the step 5 of algorithm 1 is skipped.

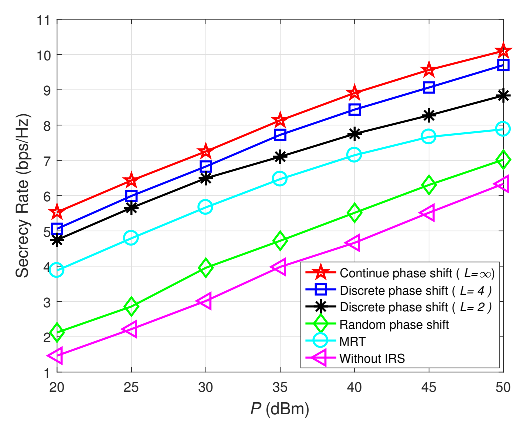

Fig. 2. Secrecy rate versus P

Security rate of algorithms versus the transmit power threshold of AP is plotted in Fig. 2. It can be seen that employing IRS can enhances the secrecy rate and the proposed algorithm achieves a higher secrecy rate than that of random phase shift algorithm. In addition, the continuous phase shift algorithm achieves the highest security rate in all algorithm and security rate of the discrete phase shift algorithm trend to be increase with L. It is also clearly that continuous phase discretization

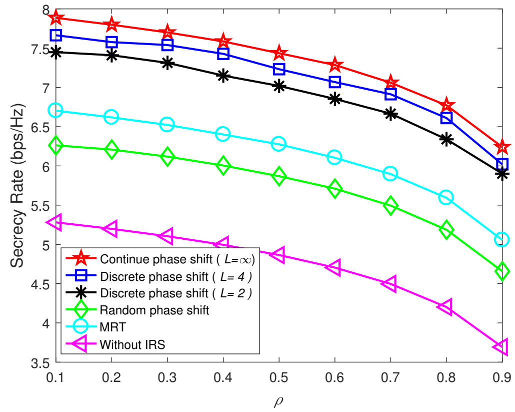

Fig. 3. Secrecy rate versus ρ

means inevitable performance loss and this loss will increase with discrete step. Due to nonlinear EH model employed, growth rate of energy harvested by UE trends to reducing with respect to transmitting power of AP increasing. Thus all algorithm suffer growth rate of security rate decreasing with transmitting power of AP increasing too. Moreover, since MRT algorithm employs the CSI of the UE and ignores the CSI of the Eve, the signal gain obtained by the Eve will gradually increase with the transmitting power of AP, which means SINR of Eve enhancing and leads growth rate of security rate of MRT algorithm lower than that of others algorithm.

In Fig. 3, we plot the security rate versus the factor ρ. As expected, the security rate of all algorithms decreases with the ρ increasing, since increasing ρ means the power using to generate AN to be decreased, lead security rate of wireless communication network to decreasing. Similar as Fig. 2, the continuous phase shift algorithm achieves the highest security rate in all algorithms and security rate of the discrete phase shift algorithm trend to be increase with L. Although there is a performance gap compared with the continuous phase shift algorithm, the proposed discrete phase shift algorithm is more easily easily realized by digital system, thus it is more suitable in the practical application.

TABLE II
COMPUTATIONAL COMPLEXITY

Algorithm	complexity
Random phase shift	$\mathcal{O}\left(\ln\left(1/\varepsilon\right)\left(M^{3.5}+N\right)\right)$
MRT	$\mathcal{O}\left(\ln\left(1/\varepsilon\right)\left(N^{4}\sqrt{N+1}\right)\right)$
Without IRS	$\mathcal{O}\left(\ln\left(1/\varepsilon\right)\left(M^{3.5}\right)\right)$
Continue phase shift	$\mathcal{O}\left(\ln\left(1/\varepsilon\right)\left(M^{3.5}+N^{4}\sqrt{N+1}\right)\right)$
Discrete phase shift	$\mathcal{O}\left(\ln\left(1/\varepsilon\right)\left(M^{3.5}+N^{4}\sqrt{N+1}+N\right)\right)$

The computational complexity of the above five algorithms can be seen in Table II. As expected, proposed method achieves the best security rate with the computational efficiency sacrificing. In addition, although the proposed algorithm experience the slightly higher complexity compared to the continuous phase shift algorithm, as mentioned in the above, it may easy to be fulfilled in current digital communication system.

V. Conclusion

In this letter, a FD-CJ scheme with the nonlinear EH model of SWIPT and discrete phase shift of IRS is proposed to improve PLS in IRS-assisted wireless communication network. Moreover, an AO-based iterative algorithm is also introduced to decompose original non-convex optimization scheme into three sub-problems. Thus, the AP beamformer, IRS phase shift matrix and AN transmit power can be sequential obtained by iterative processing. Simulations verify the effectiveness and superiority of the proposed algorithm. Notably, our algorithm obtains a suboptimal solution of the original problem since employing SDR technique. In our future work, we will explore the performance gap between the optimal of original problem and the suboptimal solution achieved by proposed algorithm.

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