Enhanced Secure Wireless Information and Power Transfer via Intelligent Reflecting Surface

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***Abstract*— In this letter, an intelligent reflecting surface (IRS)-aided secure wireless information and power transfer system is studied. To maximize the harvested power of energy harvesting receiver (EHR), we optimize the secure transmit beamforming at the access point (AP) and phase shifts at the IRS subject to the secrecy rate (SR) and the reflecting phase shifts at the IRS constraints. Due to the non-convexity of optimization problem and coupled optimization variables, we convert the optimization problem into a semidefinite relaxation (SDR) problem and a sub-optimal solution is obtained. To reduce the high-complexity of the proposed SDR method, a low-complexity alternating optimization (LC-AO) algorithm is proposed. Simulation results show that the harvested power of the proposed SDR and LC-AO methods approximately double that of the existing method without IRS with the same SR. In particular, the proposed LC-AO achieves almost the same performance as the proposed SDR but with a much lower complexity.**

***Index Terms*— Intelligent reflecting surface, secure transmit beamforming, secrecy rate, phase shifts, harvested power.**

I. Introduction

SUSTAINABLE green, cost-effective and secure techniques are basic requirements for beyond fifth-generation (B5G) and sixth-generation (6G) communication systems [1], [2]. As a high energy-efficient tool, intelligent reflecting surface (IRS), which may adjust the phase shifts automatically via a large number of low-cost, passive and reflecting units, has attracted wide research attention from both industry and academia [3]. Besides, IRSs are capable of enriching multipath scattering and boosting communication links as wave propagation controllers [4].

There have been some innovative studies on the IRS-assisted wireless communication systems by jointly optimizing the beamforming vector and the phase shifts at the IRS [5]–[8]. In a single-user multiple-input single-output (MISO) scenario [5], semidefinite relaxation (SDR) and Gaussian randomization algorithms were proposed to obtain a sub-optimal solution to maximize the total received signal power. Similarly, gradient descent and sequential fractional programming were applied to maximize the energy efficiency in multi-user MISO scenario [6]. Particularly, continuous and discrete phase shifts were exploited in [7], and the authors proposed Lagrangian dual transform to decouple the coupled optimization variables. The authors in [8] proposed a deep-learning method to optimize the transmit beamforming and phase shifts at the IRS for multi-user MISO system. On the other hand, simultaneous wireless information and power transfer (SWIPT) could enhance the energy efficiency and solve energy-limited issues of wireless networks to some extent [9]. However, due to the severe path loss, wireless power transfer was only suitable for short distance transmission, hence the range of energy harvesting receivers (EHR) is limited. The introduction of the IRS in the vicinity of EHR could deal with the above issue for the reason that the IRS provides additional communication links to EHR and compensate for the severe power loss over long distance. To enhance the harvested power of EHR, the SWIPT wireless network assisted by an IRS has been researched. In [10], the authors made an investigation of the maximization problem of the weighted sum power for the IRS-aided SWIPT system, and proved that it was not necessary to send a dedicated energy beamformer to EHR. Pan *et al* extended the case to MIMO broadcasting channels with the purpose of maximizing the weighted sum-rate [11].

However, owing to the broadcast nature of wireless channels, security is a crucial matter in wireless communication system [12]–[15]. Secure wireless communication assisted by IRS has been investigated in [16], [17] to enhance the achievable secrecy rate (SR) of the legitimate user. Note that the aforementioned studies do not consider the security of the IRS-aided SWIPT system. Although the IRS-aided SWIPT system could enhance the power transfer efficiency and the received signal power to the legitimate receiver, the power received at eavesdroppers (EVEs) will be also improved. Therefore, how to improve the secrecy performance

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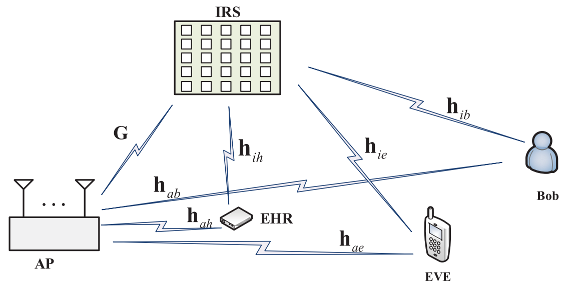

Fig. 1. An IRS-aided secure SWIPT wireless network.

of the IRS-assisted SWIPT system is an important issue. Consequently, we apply the physical layer security techniques to enhance the security performance of the considered system. Our main contributions are as follows:

1) A secure SWIPT system model assisted by an IRS is proposed, in which a multi-antenna access point (AP) is to serve an EHR and an information receiver with the help of an IRS in the presence of an EVE. By jointly optimizing the secure transmit beamforming and the phase shifts at the IRS, the harvested power maximization problem (HPMP) is established for an IRS-aided secure MISO-SWIPT system.
2) Due to the non-convex SR constraint, the uni-modular phase shifts constraint and the coupled optimization variables, thus the HPMP is non-convex and hard to solve directly. To address this issue, the problem is solved via alternating optimization (AO) algorithms. Firstly, the problem is converted into a linear optimization problem by applying the trace function. Secondly, SDR-based AO and Gaussian randomization methods are adopted to get a sub-optimal solution.
3) However, the above proposed SDR method has a high complexity. It is usually run by interior point method with hard hardware implementation. Furthermore, wireless communication has high demands of real-time transmission and requires a quick response in practice. To reduce its computational complexity, a low-complexity alternating optimization (LC-AO) algorithm is proposed. By utilizing the first-order Taylor approximation and inequality transformation, a sub-optimal solution is also obtained. Moreover, the phase shifts of the IRS are computed in semi-closed-forms per iteration. Simulation results show that our proposed SDR and LC-AO algorithms can harvest higher power than conventional scheme without IRS.

Notations: Lowercase letters represent scalars. Boldface uppercase and lowercase letters stand for matrices and vectors, respectively. $|\cdot|$ and $\|\cdot\|$ denote the modulus of a scalar and Euclidean norm of a vector, respectively. Signs $\Re(\cdot)$ and $\arg(\cdot)$ represent the real part and the phase of a complex number. $(\cdot)^H$, $(\cdot)^*$, $\mathbb{E}$ and $\mathrm{tr}(\cdot)$ represent conjugate transpose, conjugate, expectation and trace of a matrix, respectively.

II. System Model

Fig. 1 sketches a downlink MISO system with an IRS for SWIPT. In Fig. 1, there are an AP with M transmit antennas, an IRS with N reflecting units, an information receiver denoted as Bob, and an EHR in the presence of an EVE. All receivers are equipped with a single antenna. Note that the minimum requirement for EHR such as the low-power sensors is -10 dBm, which is much higher than that for Bob (-60 dBm) [9]. Therefore, EHR should be placed close to the AP. Besides, as shown in Fig. 1, an IRS is deployed in the vicinity of EHR for compensating for the associated power loss and collecting more energy. The transmit signal from AP can be expressed as

$$\mathbf{x} = \mathbf{w}s, \tag{1}$$

where $\mathbf{w} \in \mathbb{C}^{M\times 1}$ denotes the transmit beamforming vector, which forces the confidential message (CM) to the desired direction, and $s \sim \mathcal{CN}(0,1)$ is the CM for the Bob. Suppose that P_s is the total transmission power constraint. Thus, we have $\mathbb{E}\left\{\mathbf{x}^H\mathbf{x}\right\} = \|\mathbf{w}\|^2 \le P_s$.

In this letter, we assume a quasi-static fading environment. The baseband equivalent channel responses from the AP to the IRS, from the AP to Bob, from the AP to EHR, from the AP to EVE, from the IRS to Bob, from the IRS to the EHR, from the IRS to EVE are denoted by $\mathbf{G} \in \mathbb{C}^{N\times M}$, $\mathbf{h}_{ab}^H \in \mathbb{C}^{1\times M}$, $\mathbf{h}_{ah}^H \in \mathbb{C}^{1\times M}$, $\mathbf{h}_{ae}^H \in \mathbb{C}^{1\times M}$, $\mathbf{h}_{ib}^H \in \mathbb{C}^{1\times N}$, $\mathbf{h}_{ih}^H \in \mathbb{C}^{1\times N}$, $\mathbf{h}_{ie}^H \in \mathbb{C}^{1\times N}$, respectively. The diagonal reflection-coefficient matrix of the IRS is denoted as $\boldsymbol{\Theta} = \mathrm{diag}(\beta_1 e^{j\theta_1}, \cdots, \beta_n e^{j\theta_N})$, where $\theta_n \in [0, 2\pi)$ and $\beta_n \in (0,1], \forall n$ [5]. θ_n and β_n are the phase shift and amplitude reflection-coefficient of the nth unit, respectively. In this letter, $\beta_n = 1$. The received signal at Bob can be written as

$$y_b(\mathbf{w}, \boldsymbol{\Theta}) = (\mathbf{h}_{ib}^H \boldsymbol{\Theta} \mathbf{G} + \mathbf{h}_{ab}^H)\mathbf{w}s + n_b, \tag{2}$$

where $n_b \sim \mathcal{CN}(0, \sigma_b^2)$ is the complex additive white Gaussian noise (AWGN). The received signals at EVE and at EHR are

$$y_e(\mathbf{w}, \boldsymbol{\Theta}) = (\mathbf{h}_{ie}^H \boldsymbol{\Theta} \mathbf{G} + \mathbf{h}_{ae}^H)\mathbf{w}s + n_e, \tag{3}$$

$$y_r(\mathbf{w}, \boldsymbol{\Theta}) = (\mathbf{h}_{ih}^H \boldsymbol{\Theta} \mathbf{G} + \mathbf{h}_{ah}^H)\mathbf{w}s + n_h, \tag{4}$$

respectively, where n_e and n_h are the complex AWGN variables, which follow the distribution $n_e \sim \mathcal{CN}(0, \sigma_e^2)$ and $n_h \sim \mathcal{CN}(0, \sigma_h^2)$. Moreover, we assume that $\sigma_b^2 = \sigma_e^2 = \sigma_h^2 = \sigma^2$. According to (2) and (3), the achievable transmission rate at Bob and EVE can be expressed as [16]

$$R_b(\mathbf{w}, \boldsymbol{\Theta}) = \log_2\left(1 + \frac{|(\mathbf{h}_{ib}^H \boldsymbol{\Theta} \mathbf{G} + \mathbf{h}_{ab}^H)\mathbf{w}|^2}{\sigma^2}\right) \tag{5}$$

and

$$R_e(\mathbf{w}, \boldsymbol{\Theta}) = \log_2\left(1 + \frac{|(\mathbf{h}_{ie}^H \boldsymbol{\Theta} \mathbf{G} + \mathbf{h}_{ae}^H)\mathbf{w}|^2}{\sigma^2}\right), \tag{6}$$

respectively. The achievable SR is defined by [18]

$$R_s(\mathbf{w}, \boldsymbol{\Theta}) = \max\left\{0, R_b(\mathbf{w}, \boldsymbol{\Theta}) - R_e(\mathbf{w}, \boldsymbol{\Theta})\right\}. \tag{7}$$

In addition, because of the broadcast nature of wireless channels, the energy is carried by information beam. The harvested power at EHR is [9]

$$E_r(\mathbf{w}, \boldsymbol{\Theta}) = \zeta(|(\mathbf{h}_{ih}^H \boldsymbol{\Theta} \mathbf{G} + \mathbf{h}_{ah}^H)\mathbf{w}|^2), \tag{8}$$

where ζ denotes the efficiency of power harvesting.

III. Problem Formulation and Proposed Solution

In this section, we maximize the harvested power at EHR by jointly optimizing the secure transmit beamforming vector and phase shifts at the IRS to ensure that the achieved SR is greater than a predefined threshold. Moreover, similar to [16], the CSIs of all the receivers' channels are assumed to be available at the AP and the IRS. The results in this letter can be considered as the upper bound of performance [19]. Then the optimization problem can be mathematically cast as

$$\text{(P1)}: \max_{\mathbf{w},\boldsymbol{\Theta}} E_r(\mathbf{w},\boldsymbol{\Theta}) \tag{9a}$$
$$\text{s. t. } R_s(\mathbf{w},\boldsymbol{\Theta}) \geq r_0 \tag{9b}$$
$$\|\mathbf{w}\|^2 \leq P_s \tag{9c}$$
$$|e^{j\theta_n}| = 1, \quad \forall n = 1\cdots N, \tag{9d}$$

where $r_0 > 0$ denotes the minimum SR, $P_s > 0$ refers to the prescribed power budget at AP, and θ_n is the phase shift at the IRS. It can be observed that problem (P1) is non-convex because the objective function and the constraints are non-convex; moreover, optimization variables $\mathbf{w}$ and $\boldsymbol{\Theta}$ are coupled. It is particularly noted that the objective function is convex with respect to $\mathbf{w}$ and $\boldsymbol{\Theta}$, respectively. Thus we address the corresponding optimization problem by applying the alternating and iterative manner in the following. Note that the feasibility of problem (P1) is related to r_0. The problem (P1) is feasible when r_0 is reasonably set.

By defining $\mathbf{u} = [\mathbf{e}^{j\theta_1},\cdots,\mathbf{e}^{j\theta_N}]^H$, $\mathbf{v} = [\mathbf{u};1]$, $\mathbf{H}_r = [\text{diag}\{\mathbf{h}_{ih}^H\}\mathbf{G};\mathbf{h}_{ah}^H]$, $\mathbf{H}_b = [\text{diag}\{\mathbf{h}_{ib}^H\}\mathbf{G};\mathbf{h}_{ab}^H]$, and $\mathbf{H}_e = [\text{diag}\{\mathbf{h}_{ie}^H\}\mathbf{G};\mathbf{h}_{ae}^H]$, problem (P1) is equivalent to

$$\text{(P2)}: \max_{\mathbf{w},\mathbf{v}} |\mathbf{v}^H\mathbf{H}_r\mathbf{w}|^2 \tag{10a}$$
$$\text{s. t. } |\mathbf{v}^H\mathbf{H}_b\mathbf{w}|^2+\sigma^2 \geq 2^{r_0}(|\mathbf{v}^H\mathbf{H}_e\mathbf{w}|^2+\sigma^2) \tag{10b}$$
$$\|\mathbf{w}\|^2 \leq P_s \tag{10c}$$
$$|\mathbf{v}_n| = 1, \quad \forall n = 1\cdots N \ \mathbf{v}_{N+1} = 1. \tag{10d}$$

A. Proposed SDR-Based Alternating Optimization Method

In this subsection, we present a near optimal solution to problem (P2). For brevity, we rewrite (10a) and (10b) as $f(\mathbf{W},\mathbf{V}) \triangleq \text{tr}(\mathbf{H}_r^H\mathbf{V}\mathbf{H}_r\mathbf{W})$ and

$$\text{tr}(\mathbf{H}_b^H\mathbf{V}\mathbf{H}_b\mathbf{W})+\sigma^2 \geq 2^{r_0}(\text{tr}(\mathbf{H}_e^H\mathbf{V}\mathbf{H}_e\mathbf{W})+\sigma^2), \tag{11}$$

where $\mathbf{V} = \mathbf{v}\mathbf{v}^H$ and $\mathbf{W} = \mathbf{w}\mathbf{w}^H$. After dropping the rank-one constraints (i.e., rank$(\mathbf{W}) = 1$, rank$(\mathbf{V}) = 1$), the SDR of problem (P2) is

$$\text{(P3)}: \max_{\mathbf{W},\mathbf{V}} f(\mathbf{W},\mathbf{V}) \tag{12a}$$
$$\text{s. t. } \text{tr}(\mathbf{W}) \leq P_s \tag{12b}$$
$$\text{tr}(\mathbf{E}_n\mathbf{V}) = 1, \quad \forall n = 1\cdots N+1 \tag{12c}$$
$$\mathbf{W} \succeq 0, \mathbf{V} \succeq 0, (11), \tag{12d}$$

where $\mathbf{E}_n \in \mathbb{R}^{(N+1)\times(N+1)}$, which means that the value of the element on position (n,n) is 1 and 0 otherwise. Due to the fact that $\mathbf{W}$ and $\mathbf{V}$ are coupled and problem (P3) is non-convex, it is difficult to solve this kind of non-convex problems directly. However, problem (P3) could be decomposed into two subproblems and solved by applying AO algorithm. By alternately fixing $\mathbf{V}$ and $\mathbf{W}$, (P3) is reduced to two standard semidefinite programs (SDP), which can be solved by CVX directly. Making use of the AO algorithm, we obtain the solution to problem (P3). Considering that rank-one constraints are relaxed in problem (P3), the solutions to (P3) cannot be guaranteed to be rank-one. To recover the rank-one solution, we apply the standard Gaussian randomization method and obtain a high-quality sub-optimal solution of problem (P2) [5]. Different from [5], the Gaussian randomization method is used only once after we obtain the solution of problem (P3). This not only reduces the computational complexity but also ensures the convergence of the SDR-based AO algorithm. The objective function value of problem (P3) of the proposed SDR-based AO method increases in each iteration. And the optimal value of (P3) has an upper bound due to SR constraint. Therefore, the convergence of the proposed SDR-based AO algorithm is guaranteed.

B. Proposed Low-Complexity Alternating Optimization Method

In subsection III-A, we proposed the SDR-based AO Algorithm to obtain the information beamforming matrix $\mathbf{W}$ and the phase shifts matrix $\mathbf{V}$ of the IRS. However, it has a high computational complexity (i.e., $\mathcal{O}(M^{6.5}+N^{6.5})$, according to (23)). To reduce the computational complexity, a low-complexity AO Algorithm is proposed in what follows.

By fixing $\mathbf{v}$, problem (P2) is reduced to

$$\text{(P4.1)}: \max_{\mathbf{w}} |\mathbf{v}^H\mathbf{H}_r\mathbf{w}|^2 \tag{13a}$$
$$\text{s. t. } |\mathbf{v}^H\mathbf{H}_b\mathbf{w}|^2+\sigma^2 \geq 2^{r_0}(|\mathbf{v}^H\mathbf{H}_e\mathbf{w}|^2+\sigma^2) \tag{13b}$$
$$\|\mathbf{w}\|^2 \leq P_s. \tag{13c}$$

Note that problem (P4.1) is still non-convex but the objective function (13a) is convex. This motivates us to apply successive convex approximation method. The first-order Taylor expansion of $\mathbf{x}^H\mathbf{A}\mathbf{x}$ at point $\tilde{\mathbf{x}}$ is $\mathbf{x}^H\mathbf{A}\mathbf{x} \geq 2\Re\{\mathbf{x}^H\mathbf{A}\tilde{\mathbf{x}}\} - \tilde{\mathbf{x}}^H\mathbf{A}\tilde{\mathbf{x}}$. Therefore, problem (P4.1) can be further written as

$$\text{(P4.1}'\text{)}: \max_{\mathbf{w}} 2\Re\{\mathbf{w}^H\mathbf{H}_{rv}\tilde{\mathbf{w}}\} - \tilde{\mathbf{w}}^H\mathbf{H}_{rv}\tilde{\mathbf{w}} \tag{14a}$$
$$\text{s. t. } 2^{r_0}(\mathbf{w}^H\mathbf{H}_{ev}\mathbf{w}+\sigma^2) \leq 2\Re\{\mathbf{w}^H\mathbf{H}_{bv}\tilde{\mathbf{w}}\} - \tilde{\mathbf{w}}^H\mathbf{H}_{bv}\tilde{\mathbf{w}}+\sigma^2 \tag{14b}$$
$$\mathbf{w}^H\mathbf{w} \leq P_s, \tag{14c}$$

where $\mathbf{H}_{iv} = \mathbf{H}_i^H\mathbf{v}\mathbf{v}^H\mathbf{H}_i$, i takes r,e,b respectively. $\tilde{\mathbf{w}}$ is the transmit beamforming vector of previous iteration. Subproblem (P4.1$'$) can be optimally solved by using the existing software such as CVX [20].

Secondly, for any fixed $\mathbf{w}$, problem (P2) is simplified as

$$\text{(P4.2A)}: \max_{\mathbf{v}} |\mathbf{v}^H\mathbf{H}_r\mathbf{w}|^2 \tag{15a}$$
$$\text{s. t. } |\mathbf{v}^H\mathbf{H}_b\mathbf{w}|^2+\sigma^2 \geq 2^{r_0}(|\mathbf{v}^H\mathbf{H}_e\mathbf{w}|^2+\sigma^2) \tag{15b}$$
$$|\mathbf{v}_n| = 1, \quad \forall n = 1\cdots N \ \mathbf{v}_{N+1} = 1, \tag{15c}$$

where $\mathbf{v} = [\mathbf{u};1]$. By separating out the constant term of vector $\mathbf{v}$, problem (P4.2A) can be equivalently rewritten as

$$\text{(P4.2)}: \max_{\mathbf{u}} |\mathbf{u}^H\mathbf{a}+\alpha|^2 \tag{16a}$$
$$\text{s. t. } |\mathbf{u}^H\mathbf{b}+\beta|^2+\sigma^2 \geq 2^{r_0}(|\mathbf{u}^H\mathbf{c}+\gamma|^2+\sigma^2) \tag{16b}$$
$$|\mathbf{u}_n| = 1, \quad \forall n = 1\cdots N, \tag{16c}$$

where $\mathbf{a} = \text{diag}\{\mathbf{h}_{ih}^H\}\mathbf{G}\mathbf{w}$, $\alpha = \mathbf{h}_{ah}^H\mathbf{w}$, $\mathbf{b} = \text{diag}\{\mathbf{h}_{ib}^H\}\mathbf{G}\mathbf{w}$, $\beta = \mathbf{h}_{ab}^H\mathbf{w}$, $\mathbf{c} = \text{diag}\{\mathbf{h}_{ie}^H\}\mathbf{G}\mathbf{w}$ and $\gamma = \mathbf{h}_{ae}^H\mathbf{w}$. By applying the first-order Taylor expansion, (16a) can be expressed as

$$|\mathbf{u}^H\mathbf{a}+\alpha|^2 \geq 2\Re\{\mathbf{u}^H\mathbf{d}\}+c_1, \tag{17}$$

where $\mathbf{d}=\mathbf{a}\mathbf{a}^H\tilde{\mathbf{u}}+\mathbf{a}\alpha^*$, $c_1=\alpha\alpha^*-\tilde{\mathbf{u}}^H\mathbf{a}\mathbf{a}^H\tilde{\mathbf{u}}$ and $\tilde{\mathbf{u}}$ is the phase shifts vector of previous iteration. Similarly, (16b) can be expressed as

$$\mathbf{u}^H\mathbf{A}\mathbf{u}+2\Re\{\mathbf{u}^H(2^{r_0}\mathbf{c}\gamma^*-\mathbf{b}\beta^*)\} \leq \beta\beta^*+\sigma^2-2^{r_0}(\gamma\gamma^*+\sigma^2). \tag{18}$$

Furthermore $\mathbf{u}^H\mathbf{A}\mathbf{u}$ can be rewritten as [17]

$$\mathbf{u}^H\mathbf{A}\mathbf{u}\leq\mathbf{u}^H\mathbf{M}\mathbf{u}+2\Re\{\mathbf{u}^H(\mathbf{A}-\mathbf{M})\tilde{\mathbf{u}}\}+\tilde{\mathbf{u}}^H(\mathbf{M}-\mathbf{A})\tilde{\mathbf{u}}, \tag{19}$$

where $\mathbf{A}=2^{r_0}\mathbf{c}\mathbf{c}^H-\mathbf{b}\mathbf{b}^H$, $\mathbf{M}\succeq\mathbf{A}$. In this letter, we set $\mathbf{M}=\lambda_{max}(\mathbf{A})\mathbf{I}_N$ and $\mathbf{u}^H\mathbf{u}=N$, thus $\mathbf{u}^H\mathbf{M}\mathbf{u}=N\lambda_{max}(\mathbf{A})$. $\lambda_{max}(\mathbf{A})$ denotes the largest eigenvalue of matrix $\mathbf{A}$. By substituting (19) into (18), (16b) can be rewritten as

$$2\Re\{\mathbf{u}^H[(\mathbf{M}-\mathbf{A})\tilde{\mathbf{u}}+(\mathbf{b}\beta^*-2^{r_0}\mathbf{c}\gamma^*)]\}\geq c_2, \tag{20}$$

where $c_2=N\lambda_{max}(\mathbf{A})+\tilde{\mathbf{u}}^H(\mathbf{M}-\mathbf{A})\tilde{\mathbf{u}}+2^{r_0}(\gamma\gamma^*+\sigma^2)-\beta\beta^*-\sigma^2$. However, problem (P4.2) is still non-convex due to constraint (16c). It is worth noting that there always exists a non-negative μ such that (P4.2) can be formulated into the following equivalent problem.

$$(\text{P4.2}'):\ \max_{\mathbf{u}}\ 2\Re\{\mathbf{u}^H\mathbf{d}\}+2\mu\Re\{\mathbf{u}^H\mathbf{f}\}\quad \text{s. t. } (16c), \tag{21}$$

where $\mathbf{f}=(\mathbf{M}-\mathbf{A})\tilde{\mathbf{u}}+(\mathbf{b}\beta^*-2^{r_0}\mathbf{c}\gamma^*)$. When phase-shift vector $\mathbf{u}$ is equal to $\mathbf{d}+\mu\mathbf{f}$, the objective value is maximized. Therefore, the optimal solution to problem (P4.2$'$) is

$$\mathbf{u}(\mu)=e^{j\arg(\mathbf{d}+\mu\mathbf{f})}, \tag{22}$$

where μ is unknown. Substituting $\mathbf{u}(\mu)$ into constraint (20), μ can be obtained according to the complementary slackness condition $\mu[2\Re(\mathbf{u}(\mu)\mathbf{f})-c_2]=0$ [11]. To get the μ, there are two cases: (1) When $\mu=0$, then the solution of problem (P4.2) is $\mathbf{u}(0)=e^{j\arg(\mathbf{d})}$. And this solution needs to satisfy constraint (20); (2) When $\mu>0$, the complementary slackness condition holds if and only if $2\Re(\mathbf{u}(\mu)\mathbf{f})=c_2$. In this case, μ can be obtained by using bisection method since $2\Re(\mathbf{u}(\mu)\mathbf{f})$ is a monotone increasing function of μ. The result is denoted as μ', therefore the optimal solution $\mathbf{u}(\mu')$ is obtained.

Because the objective value of problem (P2) with a finite upper-bound is non-decreasing after each iteration, the proposed LC-AO method is guaranteed to converge.

C. Complexity Analysis

In this section, we will calculate the complexities of the two proposed methods and make a comparison. The total complexity of the proposed SDR-based AO algorithm without Gaussian randomization is [21]

$$\mathcal{O}\{D[\sqrt{(2+M)}\big(M^2(2+M^3)+M^4(2+M^2)+M^6\big) +\sqrt{(2N+2)}\big(N^2(N^3+N+2)+N^4(N^2+N+2)+N^6\big)]\}, \tag{23}$$

where D denotes the number of alternating iterations. The complexity of the proposed LC-AO algorithm is

$$\mathcal{O}\{L[L_1\big(2(M(5^2+(M+1)^2)+M^3)\big) +\big(L_2N^3\log_2((\lambda_{max}-\lambda_{min})/\varepsilon)\big)]\}, \tag{24}$$

where L denotes the maximum number of alternating iterations. L_1 and L_2 denote the iterative numbers of subproblems (P4.1) and (P4.2), respectively. λ_{max}, λ_{min} and ε are the upper-bound, lower-bound, and the accuracy of bisection method, respectively. $\log_2((\lambda_{max}-\lambda_{min})/\varepsilon)$ is the maximum number of bisection search. Obviously, although the LC-AO algorithm has two-level iteration, the highest order of computational complexity is M^3 and N^3 FLOPS compared to $M^{6.5}$ and $N^{6.5}$ FLOPS of the SDR-based AO algorithm. Therefore, the computational complexity of the LC-AO algorithm is much lower than that of the SDR-based AO algorithm, especially in massive IRS or massive MIMO scenario.

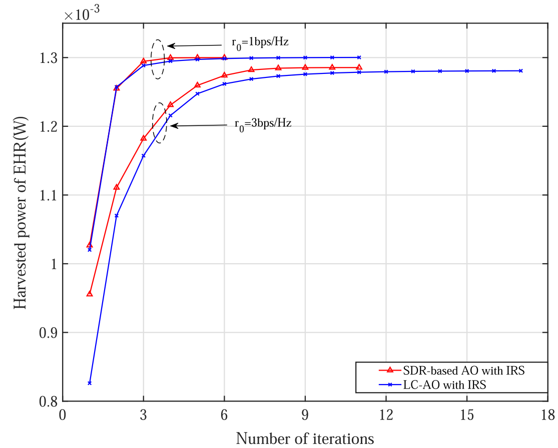

Fig. 2. Harvested power of EHR versus the number of iterations.

IV. Simulation and Discussion

In this section, we evaluate the performance of the proposed methods by numerical simulations. Two benchmark schemes are used: 1) Random phase shifts, which means that θ_n $(n=1\cdots N)$ is randomly chosen from $[0,2\pi)$. 2) Without IRS, i.e., $\mathbf{\Theta}=\mathbf{0}$. In our simulation, it is assumed that the channels of IRS-Bob/EHR/EVE and AP-Bob/EHR/EVE experience Rayleigh fading. The AP-IRS channel is assumed to be line-of-sight (LoS) links. In addition, we consider a worse scenario that the AP/IRS is closer to EVE than to Bob, which is also applicable for the scenario when Bob is close to AP/IRS. The reference path loss is 30 dB per 1 m. The path loss exponents of AP-IRS and IRS-Bob/EHR/EVE are set to be 2, whereas those of AP-Bob/EHR/EVE are 3. This is because the IRS is usually deployed to avoid blocking the signal from the AP. The distances (i,e., d_{i-j}) between node i and node j $(i,j\in\{\text{AP, IRS, EHR, Bob, EVE}\})$ are set as follows: $d_{\text{AP}-\text{IRS}}=8$ m, $d_{\text{IRS}-\text{EHR}}=3$ m, $d_{\text{AP}-\text{EHR}}=6$ m $d_{\text{AP}-\text{Bob}}=d_{\text{IRS}-\text{Bob}}=220$ m, $d_{\text{AP}-\text{EVE}}=d_{\text{IRS}-\text{EVE}}=85$ m. Other simulation parameters setting are $\sigma^2=-70$ dBm, $\zeta=0.5$ and $P_s=15$ W.

Fig. 2 demonstrates the convergence of the proposed methods when r_0 is set to be 1 bps/Hz and 3 bps/Hz, respectively. It is seen that the two proposed AO methods could converge rapidly to the power ceils within about ten iterations. This verifies the feasibility of the algorithm. After convergence, the two proposed schemes achieve excellent harvested power improvements over initial phases.

Fig. 3 illustrates the harvested power of EHR versus SR threshold r_0 for $N=50$. From this figure, it is observed

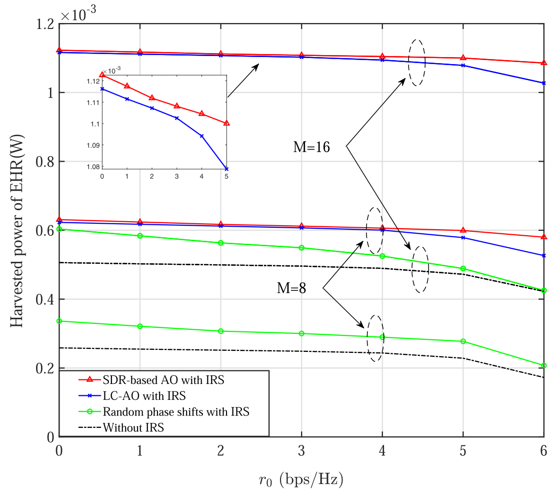

Fig. 3. Harvested power of EHR versus r_0.

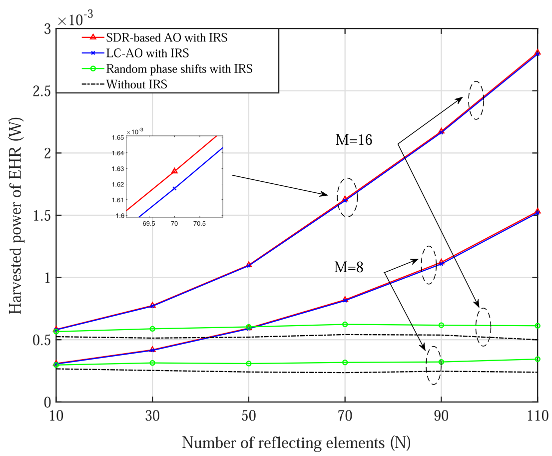

Fig. 4. Harvested power of EHR versus N.

that the harvested power of the proposed schemes decreases as r_0 increases. Compared with no IRS case, the two proposed methods with the aid of IRS approximately double the harvested power. This is because the IRS provides a new degree of freedom and diversity gain to enhance the harvested power of EHR by optimizing the phase shifts at IRS. Moreover, the proposed two methods perform much better than conventional method without IRS and random-phase-shifts method with IRS. Additionally, increasing the number of antennas at the AP accordingly improves the harvested power.

Fig. 4 plots the harvested power of EHR versus N with r_0 being 1 bps/Hz. Obviously, the proposed methods are still better than the existing methods. As N increases, the harvested power at EHR increases gradually. The main reason is that with more reflecting elements, more new degrees of freedom are achieved by the IRS.

V. Conclusion

In this letter, we have presented an investigation of secure transmit beamforming and phase shifts at the IRS in a secure IRS-assisted MISO-SWIPT network to maximize the harvested power at EHR. Two alternating iterative algorithms SDR and LC-AO were proposed to address the non-convex optimization problem. With a much lower-complexity, the proposed LC-AO method can achieve almost the same performance as the proposed SDR method. More importantly, with the help of IRS, the proposed two methods approximately double the harvested power compared to existing methods.

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