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# AN-aided beamforming for IRS-assisted SWIPT systems

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This paper studies artificial noise (AN)-aided beamforming design in an intelligent reflecting surface (IRS)-assisted system empowered by simultaneous wireless information and power transfer (SWIPT) technique. Multiple power splitting (PS) single-antenna receivers simultaneously receive information and energy from a multi-antenna base station (BS). Although all users are legitimate, in each transmission interval only one receiver is authorized to receive information and the others are only allowed to harvest power which are considered as unauthorized receivers (URs). To prevent information decoding by URs, AN signal is transmitted from the BS. We adopt a non-linear model for energy harvesting. In the optimization problem, we minimize the total transmit power, and for this purpose, we utilize an alternating optimization (AO) algorithm. For the non-convex rank-one constraint for IRS phase shifts, we utilize a sequential rank-one constraint relaxation (SROCR) algorithm. In addition to single antenna URs scenario, we investigate multi-antenna URs scenario and evaluate their performance. Simulation results validate the effectiveness of using IRS.



## 1. Introduction

Recently, intelligent reflecting surface (IRS) has been introduced as a promising technique to enhance the performance of the future generations wireless communication networks [1]. The reflecting elements of the IRS can reflect signals toward desired directions. By optimizing the phase shifts of the IRS, we can strengthen the received signal at the receivers and manipulate wireless propagation channels to have a better performance [2,3]. On the other hand, simultaneous wireless information and power transfer (SWIPT) technology is another promising technology that has been proposed for low-power devices and power-constrained systems in recent years [4]. IRS-aided wireless communication has been extensively studied recently [5–7]. For instance, [5] investigated sum-rate maximization of users by jointly optimizing the transmit beamforming at the base station (BS) and phase shifts of the IRS. In [8], the authors studied positioning of users with the help of angle of arrival (AOA) information and deploying aerial IRS. In addition, the research on SWIPT technology has been widely studied in recent years [9–11]. Moreover, some recent works proposed using IRS in SWIPT systems with separate information decoding (ID) and energy harvesting (EH) receivers [12,13]. Specifically, in [12], the authors proposed a penalty-based approach to minimize the total transmit power at the transmitter. Furthermore, like other wireless networks, there is always physical layer security concerns in SWIPT systems due to the wireless propagation nature. As we increase the transmit power to satisfy quality of service (QOS) constraints for desired receivers, the signal is more susceptible to being eavesdropped by other illegitimate receivers. To this end, transmitting artificial noise (AN) with the original signal is a solution to tackle this issue [14,15]. In [16], the authors investigated secrecy energy efficiency maximization problem with using AN. In [17–19], total transmit power minimization problem was investigated for an AN-aided multiple-input single-output (MISO) SWIPT system with linear energy harvesting (LEH) receivers. Physical layer security in IRS-assisted systems has been studied recently [20–22]. Specifically, the authors in [20] studied the use of AN in IRS system and analyzed its performance. It verified that using AN is beneficial in IRS system, and it provides a more secrecy rate than no-IRS system. The authors in [23] studied maximizing the minimum harvested energy in a multiple-input multiple-output (MIMO) SWIPT systems, and a two-layer method was introduced to solve the optimization problem. In addition, The authors in [24] investigated a secrecy rate maximization problem in a millimeter wave (mmWave) SWIPT system with non-linear energy harvesting (NLEH) model under perfect and imperfect channel state information (CSI) condition. Moreover, The combination of secure SWIPT system and IRS has been studied recently [25–28]. For example, [25] studied energy efficiency maximization in a secure IRS SWIPT system with separate ID and EH receivers. The authors proposed NLEH model for EH receivers and Dinkelbach's method was used for optimizing the beamforming vector and the AN covariance matrix. The authors in [26,27] studied the application

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of IRS for secrecy rate maximization in a SWIPT system with MIMO channel. Specifically, in [27], a penalty-based approach was proposed for optimizing the unit modulus constraint in the problem. Furthermore, the authors in [29] investigated a robust and secure SWIPT system, aiming to maximize the information rate of the information users.

In this paper, we investigate secure communication in an IRS-assisted SWIPT system. In particular, we propose an effective method to minimize the total transmit power while ensuring secure beamforming for the authorized receiver (AR) in presence of unauthorized receivers (URs) with the help of AN transmission. Most of the existing works such as [9,12,13,21,25] considered separate ID and EH receivers, which is not a proper type of receivers to deploy in internet of things (IoT) SWIPT systems. Moreover, there is a low number of works that studied NLEH model for receivers, and most of them such as [4,9,13,15–19,26] assumed LEH model, which is far away from reality. In addition, there are a small number of literatures that investigated physical layer security in IRS-assisted SWIPT systems (e.g. [25–29]), and the major concentration was in secrecy rate maximization problem. To the best of our knowledge, in this area, [28] is the only work that studied transmit power minimization problem, which has a major difference to our work. Specifically, the receivers in the aforementioned work are separate ID and EH receivers, whereas we use co-located receivers in our work. Furthermore, [28] studied terahertz technology and the AN transmission is not considered in the system. Moreover, the optimization problem, the approach and algorithms for its solution are significantly different from our work. The main contributions of this paper are summarized as follows:

- In this work, we propose IRS-aided SWIPT systems with physical layer security concern, where the AN-transmission is adopted to combat information leakage to URs. Our goal is to minimize the total transmit power under QoS constraints. Unlike most existing works on SWIPT that focused on single antenna receivers, we extend our system model to a multi-antenna scenario.
- The formulated optimization problem is non-convex and hard to solve. To tackle this, first, we obtain a globally optimal solution for the active beamforming problem in the first phase, and then in the second phase, we propose a novel sequential rank-one constraint relaxation (SROCR) technique to attain a rank-one solution for the passive beamforming problem. Then we employ an efficient alternating optimization (AO) algorithm to jointly optimize active and passive beamforming phase efficiently.
- For the energy harvesting at the receivers, we adopt non-linear model which is practical and more realistic compared to the LEH model. Furthermore, we propose a power splitting (PS) architecture for receivers and optimize PS ratio efficiently, which is more practical than other receivers and more complex than separate receivers. Accordingly, the receivers can decode information and harvest energy simultaneously in an efficient way, especially in low-powered devices.
- When the CSI of URs is unknown, we adopt isotropic-AN design, where the AN is generated and transmitted with isotropic pattern without interfering with AR's channel. Accordingly, the transmitter should allocate more power for generating AN to prevent information decoding of AR by URs.
- Simulation results show that deploying IRS can effectively lower total transmit power in comparison with no-IRS case. Furthermore, utilizing CSI of URs has superior performance than isotropic-AN design. Moreover, numerical results demonstrate that the decrease in transmit power does not lead to secrecy rate deterioration.

The rest of this paper is organized as follows. In Section 2, the system model and NLEH model are introduced. Section 3 provides the problem formulation and its solution. Section 4 provides computational complexity. In Section 5, the multi-antenna extension is investigated. The simulation results are provided in Section 6. Finally, conclusions are given in Section 7.

## 2. System model

We consider an IRS-assisted downlink SWIPT system, as in Fig. 1, consisting a base station (BS) with $N_T$ antennas, an IRS with $M$ reflecting elements and $K+1$ hybrid single-antenna receivers. All the receivers are legitimate users, but there is only one AR in each transmission interval which is allowed to decode information and harvest energy simultaneously. At the same time, the other $K$ receivers are URs in that interval and only allowed to harvest energy. To ensure a secure communication for the AR, AN signal is transmitted from BS. The results of our proposed scheme in this paper serve as the upper bounds for the cases with CSI errors. The received signals at the AR and at the $k$'th UR are respectively expressed as follows

$$y^A = (\mathbf{h}_r^H \Theta \mathbf{G} + \mathbf{h}_d^H)\mathbf{x} + n^A, \tag{1}$$

$$y_k^U = (\mathbf{g}_{r,k}^H \Theta \mathbf{G} + \mathbf{g}_{d,k}^H)\mathbf{x} + n_k^U, \tag{2}$$

where $\mathbf{G} \in \mathbb{C}^{M\times N_T}$ is the channel matrix from the BS to the IRS and $\mathbf{g}_{r,k}^H \in \mathbb{C}^{1\times M}$, $\mathbf{g}_{d,k}^H \in \mathbb{C}^{1\times N_T}$, $\mathbf{h}_r^H \in \mathbb{C}^{1\times M}$, and $\mathbf{h}_d^H \in \mathbb{C}^{1\times N_T}$ denote channel vector from the IRS to $k$'th UR, from the BS to the $k$'th UR, from the IRS to the AR, and from the BS to the AR, respectively. The phase shift matrix of the IRS elements is denoted by $\Theta = \mathrm{diag}(\exp(j\boldsymbol{\theta}))$, where $\boldsymbol{\theta} = [\theta_1, \ldots, \theta_M]^{\mathrm{T}}$, and $\theta_m \in [0, 2\pi]$ denotes the phase shift of the $m$'th reflecting elements of the IRS. The additive white Gaussian noise at the AR and at the $k$'th UR are denoted by $n^A$ and $n_k^U$, respectively both with zero mean and variance $\sigma_n^2$. The transmitted signal vector from the BS is denoted by $\mathbf{x} \in \mathbb{C}^{N_T}$ and is expressed as

$$\mathbf{x} = \mathbf{w}s + \mathbf{z}, \tag{3}$$

where $\mathbf{w}$ and $s \sim \mathcal{CN}(0,1)$ denote beamforming vector and the information bearing signal for AR, respectively. In addition, $\mathbf{z} \in \mathbb{C}^{N_T}$ denotes the AN signal vector. In our proposed scheme, the receivers employ PS technique. Thus the received signal in each user is split into two streams with a power splitting factor $\rho$. The received signal portion at the AR for information decoding can be written as

$$y^{ID} = \sqrt{\rho}(\mathbf{h}^H\mathbf{x} + n^A) + n_s, \tag{4}$$

where $n_s \sim \mathcal{CN}(0, \sigma_s^2)$ is the processing noise in the receivers and $\mathbf{h}^H = \mathbf{h}_r^H \Theta \mathbf{G} + \mathbf{h}_d^H$. We consider a non-linear model for energy harvesting at the receivers and the harvested energy $E^{\mathrm{H}}$ is expressed as [30]

$$E^{\mathrm{H}} = \frac{\frac{\Lambda}{1+e^{-a(E^{\mathrm{in}}-b)}} - \frac{\Lambda}{1+e^{ab}}}{1 - \frac{1}{1+e^{ab}}}, \tag{5}$$

where $a$, $b$ and $\Lambda$ are constant parameters [30]. For the receivers with NLEH model, the input power is denoted by $E^{\mathrm{in}}$, while the output power from NLEH circuit is denoted by $E^{\mathrm{H}}$. The achievable secrecy rate is given by

$$\mathrm{R}^{sec} = [\log_2(1+\Gamma^A) - \max_{k\in\{1,\ldots,K\}} \log_2(1+\Gamma_k^U)]^+, \tag{6}$$

where $\Gamma^A$ and $\Gamma_k^U$ are the signal-to-interference-plus-noise ratio (SINR) at the AR and $k$'th UR, respectively, and can be expressed as

$$\Gamma^A = \frac{\rho|\mathbf{h}^H\mathbf{w}|^2}{\rho\mathrm{Tr}(\mathbf{h}\mathbf{h}^H\mathbf{Z}) + \rho\sigma_n^2 + \sigma_s^2}, \tag{7}$$

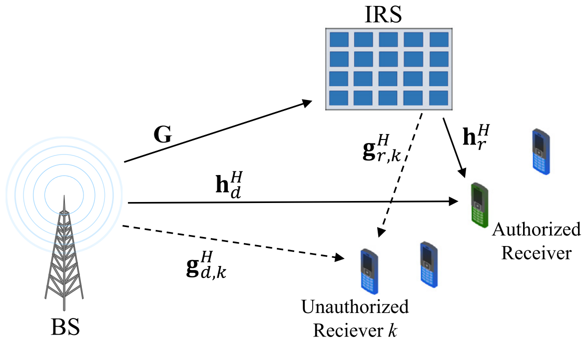


**Fig. 1.** An IRS-assisted SWIPT system.

$$\Gamma_k^U = \frac{|\mathbf{g}_k^H \mathbf{w}|^2}{\mathrm{Tr}(\mathbf{g}_k \mathbf{g}_k^H \mathbf{Z}) + \sigma_n^2 + \sigma_s^2}, \tag{8}$$

where $\mathbf{g}_k^H = \mathbf{g}_{r,k}^H \Theta \mathbf{G} + \mathbf{g}_{d,k}^H$. We assume the power splitting ratio is equal to one for URs, which means all URs are trying to decode information signal.

## 3. Problem formulation

In this problem, our objective is to minimize total transmit power by jointly optimizing the beamforming vector at the BS, AN covariance matrix at the BS, PS ratio and the phase shift matrix $\Theta$ at the IRS while ensuring the security at the AR. The optimization problem is expressed by

$$\min_{\mathbf{w},\mathbf{Z},\rho,\boldsymbol{\theta}} \quad \|\mathbf{w}\|^2 + \mathrm{Tr}(\mathbf{Z}) \tag{9a}$$

$$\text{s.t.} \quad \frac{\rho |\mathbf{h}^H \mathbf{w}|^2}{\rho \mathrm{Tr}(\mathbf{h}\mathbf{h}^H \mathbf{Z}) + \rho \sigma_n^2 + \sigma_s^2} \geq \gamma, \tag{9b}$$

$$\frac{|\mathbf{g}_k^H \mathbf{w}|^2}{\mathrm{Tr}(\mathbf{g}_k \mathbf{g}_k^H \mathbf{Z}) + \sigma_n^2 + \sigma_s^2} \leq \gamma_{e,k}, \quad \forall k, \tag{9c}$$

$$E^{\mathrm{H}}(E^{\mathrm{r}}) \geq E_{\min}, \tag{9d}$$

$$E^{\mathrm{H}}(E_{e,k}^{\mathrm{r}}) \geq E_{\min_{e,k}}, \quad \forall k, \tag{9e}$$

$$0 < \rho < 1, \tag{9f}$$

$$0 \leq \theta_m \leq 2\pi, \quad \forall m, \tag{9g}$$

$$\mathbf{Z} \succeq 0, \tag{9h}$$

where $\gamma$ is the minimum required SINR for AR and $\gamma_{e,k}$ is the SINR threshold for the $k$'th UR, which the SINR for the URs should be lower than this threshold to prevent them from decoding the information. Thus when SINR constraints for both types of receivers holds, secure communication is guaranteed [14,25], and $R^{sec} \geq 0$ holds for the communication system. For the NLEH model, we define $E^{\mathrm{in}} = E^{\mathrm{r}}$ and $E^{\mathrm{in}} = E_{e,k}^{\mathrm{r}}$ for the AR and k'th UR, respectively. The minimum required harvested power by AR and $k$'th UR are denoted by $E_{\min}$ and $E_{\min_{e,k}}$, respectively. The energy efficiency at the AR and $k$'th UR are denoted by $\eta$ and $\eta_{e,k}$, respectively. Problem (9) is non-convex and hard to solve due to $\boldsymbol{\theta}$ and coupled variables in the constraint (9b) and (9d). Thus we split the problem into two sub-problem and solve problem (9) by alternating optimization. In the first subproblem we optimize $\mathbf{w}$, $\mathbf{Z}$, $\rho$ for given $\boldsymbol{\theta}$. For convenience, we reformulate (5) as

$$E^{\mathrm{in}}(E^{\mathrm{H}}) = b - \frac{1}{a} \ln \left( \frac{e^{ab}(\Lambda - E^{\mathrm{H}})}{\Lambda + e^{ab} E^{\mathrm{H}}} \right). \tag{10}$$

After rewriting problem (9) into SDP form, by defining $\mathbf{W} = \mathbf{w}\mathbf{w}^H$, we have

$$\min_{\mathbf{W},\mathbf{Z},\rho} \quad \mathrm{Tr}(\mathbf{W}) + \mathrm{Tr}(\mathbf{Z}) \tag{11a}$$

$$\text{s.t.} \quad \mathbf{h}^H \left( \frac{\mathbf{W}}{\gamma} - \mathbf{Z} \right) \mathbf{h} \geq \sigma_n^2 + \frac{\sigma_s^2}{\rho}, \tag{11b}$$

$$\mathbf{g}_k^H \left( \frac{\mathbf{W}}{\gamma_{e,k}} - \mathbf{Z} \right) \mathbf{g}_k \leq \sigma_n^2 + \sigma_s^2, \quad \forall k, \tag{11c}$$

$$\mathbf{h}^H (\mathbf{W} + \mathbf{Z}) \mathbf{h} + \sigma_n^2 \geq \frac{E^{\mathrm{r}}(E_{\min})}{\eta(1-\rho)}, \tag{11d}$$

$$\mathbf{g}_k^H (\mathbf{W} + \mathbf{Z}) \mathbf{g}_k + \sigma_n^2 \geq \frac{E_{e,k}^{\mathrm{r}}(E_{\mathrm{min}e,k})}{\eta_{e,k}}, \quad \forall k, \tag{11e}$$

$$0 < \rho < 1, \tag{11f}$$

$$\mathrm{rank}(\mathbf{W}) = 1, \tag{11g}$$

$$\mathbf{Z} \succeq 0, \ \mathbf{W} \succeq 0. \tag{11h}$$

Problem (11) is still non-convex due to constraint (11g). To overcome this, we drop constraint (11g) and then we have

$$\min_{\mathbf{W},\mathbf{Z},\rho} \quad \mathrm{Tr}(\mathbf{W}) + \mathrm{Tr}(\mathbf{Z}) \tag{12a}$$

$$\text{s.t.} \quad \text{(11b)–(11f), (11h).} \tag{12b}$$

Since problem (12) is convex and satisfies Slater's condition, the duality gap is zero. Thus by investigating KKT conditions for the dual problem of (12), we can verify the rank-one condition. Fortunately, in this problem, the optimal solution satisfies Rank($\mathbf{W}$) $=$ 1 and similarly it can be verified that Rank($\mathbf{Z}$) $= 1$ is already hold (see Appendix). After solving problem (12), we can find $\mathbf{w}^*$ and $\mathbf{z}^*$ by performing EVD. In the second sub-problem, we optimize $\boldsymbol{\theta}$ for given $\mathbf{w}$, $\rho$ and $\mathbf{z}$ in the second sub-problem. Thus, by defining $v = [e^{j\theta_1}, \ldots, e^{j\theta_M}]^H$ and $\mathbf{r}^H = [v^T, 1]$, the problem for the second part is expressed as

$$\text{find} \quad \mathbf{r} \tag{13a}$$

$$\text{s.t.} \quad \frac{|\mathbf{r}^H \mathbf{G}_s \mathbf{w}|^2}{|\mathbf{r}^H \mathbf{G}_s \mathbf{z}|^2 + \sigma_n^2 + \frac{\sigma_s^2}{\rho}} \geq \gamma, \tag{13b}$$

$$\frac{|\mathbf{r}^H \mathbf{G}_{q,k} \mathbf{w}|^2}{|\mathbf{r}^H \mathbf{G}_{q,k} \mathbf{z}|^2 + \sigma_n^2 + \sigma_s^2} \leq \gamma_{e,k}, \quad \forall k, \tag{13c}$$

$$|\mathbf{r}^H \mathbf{G}_s \mathbf{w}|^2 + |\mathbf{r}^H \mathbf{G}_s \mathbf{z}|^2 + \sigma_n^2 \geq \frac{E^{\mathrm{r}}(E_{\min})}{\eta(1-\rho)}, \tag{13d}$$

$$|\mathbf{r}^{\mathbf{H}} \mathbf{G}_{q,k} \mathbf{w}|^2 + |\mathbf{r}^{\mathbf{H}} \mathbf{G}_{q,k} \mathbf{z}|^2 + \sigma_n^2 \geq \frac{E_{e,k}^{\mathrm{r}}(E_{\min_{e,k}})}{\eta_{e,k}}, \quad \forall k, \tag{13e}$$

$$|\mathbf{r}_m| = 1, \quad m = 1, \ \ldots \ , M+1, \tag{13f}$$

where $\mathbf{G}_s = [\mathrm{diag}(\mathbf{h}_r^H)\mathbf{G}; \ \mathbf{h}_d^H]$, $\mathbf{G}_{q,k} = [\mathrm{diag}(\mathbf{g}_{r,k}^H)\mathbf{G}; \ \mathbf{g}_{d,k}^H]$. We define $\mathbf{R} = \mathbf{r}\mathbf{r}^H$, $\mathbf{W} = \mathbf{w}\mathbf{w}^H$ and $\mathbf{Z} = \mathbf{z}\mathbf{z}^H$, and then we have the following transformations

$$\begin{aligned} |\mathbf{r}^H \mathbf{G}_s \mathbf{w}|^2 &= \mathrm{Tr}\left(\mathbf{G}_s \mathbf{W} \mathbf{G}_s^H \mathbf{R}\right), \\ |\mathbf{r}^H \mathbf{G}_s \mathbf{z}|^2 &= \mathrm{Tr}\left(\mathbf{G}_s \mathbf{Z} \mathbf{G}_s^H \mathbf{R}\right), \\ |\mathbf{r}^H \mathbf{G}_{q,k} \mathbf{w}|^2 &= \mathrm{Tr}\left(\mathbf{G}_{q,k} \mathbf{W} \mathbf{G}_{q,k}^H \mathbf{R}\right), \\ |\mathbf{r}^H \mathbf{G}_{q,k} \mathbf{z}|^2 &= \mathrm{Tr}\left(\mathbf{G}_{q,k} \mathbf{Z} \mathbf{G}_{q,k}^H \mathbf{R}\right). \end{aligned} \tag{14}$$

Problem (13) can be rewritten as following

$$\text{find} \quad \mathbf{R} \tag{15a}$$

$$\text{s.t.} \quad \mathrm{Tr}\left( \mathbf{G}_s (\frac{\mathbf{W}}{\gamma} - \mathbf{Z}) \mathbf{G}_s^H \mathbf{R} \right) \geq \frac{\sigma_s^2}{\rho} + \sigma_n^2, \tag{15b}$$

$$\mathrm{Tr}\left(\mathbf{G}_{q,k}(\frac{\mathbf{W}}{\gamma_{e,k}}-\mathbf{Z})\mathbf{G}_{q,k}^{H}\mathbf{R}\right)\leq\sigma_s^2+\sigma_n^2,\quad\forall k, \tag{15c}$$

$$\mathrm{Tr}\left(\mathbf{G}_s(\mathbf{W}+\mathbf{Z})\mathbf{G}_s^{H}\mathbf{R}\right)+\sigma_n^2\geq\frac{E^{\mathrm{r}}(E_{\min})}{\eta(1-\rho)}, \tag{15d}$$

$$\mathrm{Tr}\left(\mathbf{G}_{q,k}(\mathbf{W}+\mathbf{Z})\mathbf{G}_{q,k}^{H}\mathbf{R}\right)+\sigma_n^2\geq\frac{E_{e,k}^{\mathrm{r}}(E_{\min_{e,k}})}{\eta_{e,k}},\quad\forall k, \tag{15e}$$

$$\mathrm{Tr}(\mathbf{T}_m\mathbf{R})=1,\quad m=1,\ \ldots\ ,M+1, \tag{15f}$$

$$\mathrm{Rank}(\mathbf{R})=1, \tag{15g}$$

$$\mathbf{R}\succeq 0, \tag{15h}$$

where

$$[\mathbf{T}_m]_{i,j}=\begin{cases}1 & i=j=m\\ 0 & o.w.\end{cases} \tag{16}$$

**Algorithm 1** Proposed SROCR algorithm for phase shifts

**Initialization:** set iteration number p=0, $L^{(0)}$, step size $\varphi^{(0)}$.
Solve problem (17) with initial parameters and obtain $\mathbf{R}^{(0)}$.
**repeat**
  Given $\{L^{(\mathrm{p})},\mathbf{R}^{(\mathrm{p})}\}$, Solve problem (17).
  **if** problem (17) is feasible, denote the optimal solution as $\mathbf{R}^{(\mathrm{p}+1)}$ **then**
    $\varphi^{(\mathrm{p}+1)}=\varphi^{(\mathrm{p})}$.
  **else**
    $\varphi^{(\mathrm{p}+1)}=\frac{\varphi^{(\mathrm{p})}}{2}$.
  **end if**
  update $L^{(\mathrm{p}+1)}=\min\left(1,\frac{\lambda_{max}(\mathbf{R}^{(\mathrm{p}+1)})}{\mathrm{Tr}(\mathbf{R}^{(\mathrm{p}+1)})}+\varphi^{(\mathrm{p}+1)}\right)$.
  set $\mathrm{p}=\mathrm{p}+1$.
**until** $L^{(\mathrm{p}-1)}\geq\epsilon_1$ and with obtained $\mathbf{R}^{(\mathrm{p})}$, the objective value attain a predefined convergence threshold $\epsilon_2\geq 0$.

The rank-one constraint in (15g) is non-convex and the relaxed problem may not lead to a rank-one solution. In most cases for obtaining rank-one solution, we can drop the rank-one constraint and after finding a solution with higher rank, we can employ different approaches to find a rank-one solution. For instance, Gaussian randomization method is used extensively to obtain a rank-one solution. However, there is no guarantee that the obtained solution being feasible to the original problem and the resulting solution is usually suboptimal solution. Another method that can be used to construct rank-one solution is penalty-based method. In this method, approximating the non-convex objective function plays an important role in the quality of the obtained rank-one solution. One drawback of this method is that the approximation in the objective function normally results a suboptimal solution that may be considerably different from the optimal solution. To overcome this issue, we apply SROCR technique [31] to our problem. In this method, we drop rank-one constraint and replace it with a new constraint and then the problem can achieve a rank-one solution gradually with the help of a relaxation parameter. The resulting problem can be reformulated as following

$$\text{find}\quad\mathbf{R} \tag{17a}$$

$$\text{s.t.}\quad(15b)-(15f),\ (15h), \tag{17b}$$

$$\mathrm{u}_{max}^{H}(\mathbf{R}^{(\mathrm{p})})\mathbf{R}\mathrm{u}_{max}(\mathbf{R}^{(\mathrm{p})})\geq L^{(\mathrm{p})}\mathrm{Tr}(\mathbf{R}), \tag{17c}$$

where $\mathrm{u}_{max}^{H}(\mathbf{R}^{(\mathrm{p})})$ denotes the largest eigenvector of $\mathbf{R}$, and $L^{(\mathrm{p})}$ denotes the relaxation parameter in the p'th iteration. The constraint (17c) is the relaxed form of the rank-one constraint in problem (15). Problem (17) can be solved by CVX [32]. The proposed iterative algorithm for finding a rank-one solution in problem (17), is summarized in Algorithm 1. The overall AO algorithm for convergence of the two sub-problems is summarized in Algorithm 2. The objective value of the first sub-problem is a monotonically decreasing function, hence, the convergence of algorithm 2 is guaranteed.

**Algorithm 2** Proposed AO algorithm

1: **Initialization:** set iteration number $\mathrm{p}'=0$, $\boldsymbol{\theta}^{(0)}$.
2: **repeat**
3:   set $\mathrm{p}'=\mathrm{p}'+1$.
4:   Given $\boldsymbol{\theta}^{(\mathrm{p}'-1)}$, obtain the optimal solution by solving problem (12) and performing EVD and then, denote it as $\{\mathbf{w}^{(\mathrm{p}')},\mathbf{z}^{(\mathrm{p}')},\rho^{(\mathrm{p}')}\}$.
5:   Given $\{\mathbf{w}^{(\mathrm{p}')},\mathbf{z}^{(\mathrm{p}')},\rho^{(\mathrm{p}')}\}$, solve problem (17) by using **Algorithm 1** to obtain $\boldsymbol{\theta}^{(\mathrm{p}')}$.
6: **until** The fractional decrease of total transmit power is below $\epsilon$.

## 4. Computational complexity

In this section, the complexity of our proposed scheme is provided. The complexity of the overall algorithm of the proposed scheme can be determined by step 3 and 4 in Algorithm 2. The complexity of step 3 and 4 are $\mathcal{O}_1=\mathcal{O}(2(k+N_T+2)^{0.5}[(2N_T^3+2K+4)n_1+(2N_T^2+2K+4)n_1^2+n_1^3])$, and $\mathcal{O}_2=\mathcal{O}((2(K+2)+M)^{0.5}[((M+1)^3+2k+3)n_2+((M+1)^2+2k+3)n_2^2+n_2^3])$, respectively, where $n_1=\mathcal{O}(2N_T^2+1)$, and $n_2=\mathcal{O}((M+1)^2)$. Finally, the computational complexity of the proposed algorithm is $\mathcal{O}(\ell_2(\mathcal{O}_1+\ell_1\mathcal{O}_2))$, where $\ell_1$ and $\ell_2$ denote the iteration number of Algorithm 1 and that of Algorithm 2, respectively.

## 5. Extension to multi-antenna URs

We consider that the URs have $N>1$ antennas. To this end, some modifications should be considered. For the multi-antenna scenario, the received SINR at the k'th UR is as follows

$$\frac{\sum_{n=1}^{N}|(\mathbf{g}_{r,k,n}^{H}\Theta\mathbf{G}+\mathbf{g}_{d,k,n}^{H})\mathbf{w}|^2}{\mathrm{Tr}((\mathbf{G}_{r,k}^{H}\Theta\mathbf{G}+\mathbf{G}_{d,k}^{H})\mathbf{Z}(\mathbf{G}^{H}\Theta^{H}\mathbf{G}_{r,k}+\mathbf{G}_{d,k}))+\sigma_n^2+\sigma_s^2}, \tag{18}$$

where $\mathbf{G}_{r,k}^{H}=[\mathbf{g}_{r,k,1},\ldots,\ \mathbf{g}_{r,k,N}]^{H}$ and $\mathbf{G}_{d,k}^{H}=[\mathbf{g}_{d,k,1},\ldots,\ \mathbf{g}_{d,k,N}]^{H}$ denote the channel matrix from the IRS to the $k$'th UR and from the BS to the $k$'th UR, respectively. For the k'th UR, $\mathbf{g}_{r,k,n}\in\mathbb{C}^{N_T\times 1}$ denotes the channel response between the n'th receive antenna and the transmit array. We introduce $\mathbf{G}_{\mathrm{U},k}^{H}=\mathbf{G}_{r,k}^{H}\Theta\mathbf{G}+\mathbf{G}_{d,k}^{H}$ and rewrite (18), then the relaxed SDP form of the first subproblem is formulated as

$$\min_{\mathbf{W},\mathbf{Z},\rho}\quad\mathrm{Tr}(\mathbf{W})+\mathrm{Tr}(\mathbf{Z}) \tag{19a}$$

$$\text{s.t.}\quad(11b),(11d),(11f),(11h), \tag{19b}$$

$$\mathrm{Tr}\left(\mathbf{G}_{\mathrm{U},k}^{H}\left(\frac{\mathbf{W}}{\gamma_{e,k}}-\mathbf{Z}\right)\mathbf{G}_{\mathrm{U},k}\right)\leq\sigma_n^2+\sigma_s^2,\quad\forall k, \tag{19c}$$

$$\mathrm{Tr}\left(\mathbf{G}_{\mathrm{U},k}^{H}\left(\mathbf{W}+\mathbf{Z}\right)\mathbf{G}_{\mathrm{U},k}+\sigma_n^2\right)\geq\frac{E_{e,k}^{\mathrm{r}}(E_{\min_{e,k}})}{\eta_{e,k}},\quad\forall k. \tag{19d}$$

For the second subproblem, we have $\mathbf{G}_{o,k}=\sum_{n=1}^{N}[\mathrm{diag}(\mathbf{g}_{r,k,n}^{H})\mathbf{G}\mathbf{w};\ \ \mathbf{g}_{d,k,n}^{H}\mathbf{w}]$, $\mathbf{G}_{I,k}=\sum_{n=1}^{N}[\mathrm{diag}(\mathbf{g}_{r,k,n}^{H})\mathbf{G}\mathbf{z};\ \ \mathbf{g}_{d,k,n}^{H}\mathbf{z}]$. Subsequently, the second subproblem for this scenario can be written as

$$\text{find}\quad\mathbf{R} \tag{20a}$$

$$\text{s.t.}\quad(15b),(15d),(15f),(15h),(17c), \tag{20b}$$

$$\mathrm{Tr}(\mathbf{G}_{o,k}\mathbf{G}_{o,k}^{H}\mathbf{R}) - \mathrm{Tr}(\mathbf{G}_{l,k}\mathbf{G}_{l,k}^{H}\mathbf{R}) \leq \sigma_n^2 + \sigma_s^2, \quad \forall k, \tag{20c}$$

$$\mathrm{Tr}(\mathbf{G}_{o,k}\mathbf{G}_{o,k}^{H}\mathbf{R}) + \mathrm{Tr}(\mathbf{G}_{l,k}\mathbf{G}_{l,k}^{H}\mathbf{R}) \geq \frac{E_{e,k}^{\mathrm{r}}(E_{\mathrm{min}_{e,k}})}{\eta_{e,k}}, \quad \forall k. \tag{20d}$$

## 6. Simulation results

In this section, we present numerical results to evaluate the proposed scheme. We assume a two-dimensional plane, where the BS and IRS are located at (0, 2) and (5, 2), respectively. The URs are closer to the BS than the AR. Accordingly, the URs and the AR are randomly distributed in the range of (4, 0) to (8, 0) m and (16, 0) to (20, 0) m, respectively. For all channel links, the small-scale fading are assumed to be Rician fading with the Rician factor of $\kappa_{\mathrm{r}} = 5$. The large-scale pathloss is modeled as $\mathrm{PL} = \mathrm{PL}_0 - 10\alpha \log_{10}(d)$, where $\mathrm{PL}_0 = -30$ dB is the pathloss at the distance of 1 m. The link distance is denoted as $d$, and $\alpha$ is the pathloss exponent and is set to 2.2 for all channel links. The Rician fading channel from BS to IRS is as follows

$$\mathbf{G} = \sqrt{\frac{\kappa_{\mathrm{r}}}{1+\kappa_{\mathrm{r}}}}\mathbf{G}^{\mathrm{LOS}} + \sqrt{\frac{1}{1+\kappa_{\mathrm{r}}}}\mathbf{G}^{\mathrm{NLOS}}, \tag{21}$$

where $\mathbf{G}^{\mathrm{LOS}}$ and $\mathbf{G}^{\mathrm{NLOS}}$ denote the Line-of-sight (LOS) component and Rayleigh fading component, respectively. The Rician fading channels from BS to AR and from IRS to AR can be expressed as

$$\mathbf{h}_i = \sqrt{\frac{\kappa_{\mathrm{r}}}{1+\kappa_{\mathrm{r}}}}\mathbf{h}_i^{\mathrm{LOS}} + \sqrt{\frac{1}{1+\kappa_{\mathrm{r}}}}\mathbf{h}_i^{\mathrm{NLOS}}, \quad i \in \{d,\ r\} \tag{22}$$

and similarly, the Rician fading channels from BS to k'th UR and from IRS to k'th UR can be expressed as

$$\mathbf{g}_{i,k} = \sqrt{\frac{\kappa_{\mathrm{r}}}{1+\kappa_{\mathrm{r}}}}\mathbf{g}_{i,k}^{\mathrm{LOS}} + \sqrt{\frac{1}{1+\kappa_{\mathrm{r}}}}\mathbf{g}_{i,k}^{\mathrm{NLOS}}, \quad i \in \{d,\ r\} \tag{23}$$

where $\mathbf{h}_i^{\mathrm{LOS}}$ and $\mathbf{g}_{i,k}^{\mathrm{LOS}}$ denote the LOS component, $\mathbf{h}_i^{\mathrm{NLOS}}$ and $\mathbf{g}_{i,k}^{\mathrm{NLOS}}$ indicate Rayleigh fading component. We consider a uniform linear array (ULA) at the BS with $\mathbf{h}_i^{\mathrm{LOS}} = [1 \ \ e^{j\varpi} \ \ e^{j2\varpi} \ \ \cdots \ \ e^{j(N_T-1)\varpi}]^T$ with $\varpi = -\frac{2\,\pi\,D\,\sin(\psi)}{\lambda_r}$, and $\mathbf{g}_{i,k}^{\mathrm{LOS}} = [1 \ \ e^{j\omega_k} \ \ e^{j2\omega_k} \ \ \cdots \ \ e^{j(N_T-1)\omega_k}]^T$ with $\omega_k = -\frac{2\,\pi\,D\,\sin(\varphi_k)}{\lambda_r}$, where $D$ is the antenna spacing and is set to be $D = \frac{\lambda_r}{2}$, $\lambda_r$ is the carrier wavelength, $\psi$ and $\varphi_k$ are the directions of the transmitter to the AR and k'th URs, respectively. The number of URs is assumed to be $k = 3$ and the number of BS antennas and the number of reflecting elements are set as $N_T = 4$ and $M = 50$, respectively. For the multi-antenna scenario, the number of URs antennas is considered to be $N = 3$. According to [33], we have $\Lambda = 24$ mW, $a = 150$, and $b = 0.014$, for the NLEH model. Other parameters are set as: $\sigma_n = -120$ dBm, $\sigma_s = -60$ dBm, $E_{\mathrm{min}} = E_{\mathrm{min}_{e,k}} = 0$ dBm, $\eta = \eta_{e,k} = 0.5$, $\epsilon = 10^{-3}$ and the maximum allowed SINR for the EHRs, $\gamma_{e,k} = -10$ dB. We compare our proposed scheme with the following benchmark schemes:

- **Isotropic-AN:** When CSI of the URs are not available, isotropic-AN design [34] can be used, where AN is uniformly sent in the null space of the AR. Particularly, the AN covariance matrix is as follows

$$\mathbf{Z}_{iso-AN} = \mu \Pi^{\perp}, \tag{24}$$

where $\Pi^{\perp} = \mathbf{I}_{N_T} - \frac{\mathbf{h}\mathbf{h}^H}{\|\mathbf{h}\|^2}$ denotes the orthogonal complement projector of $\mathbf{h}$, and $\mu \geq 0$ is the power factor that should be optimized in the first subproblem.

- **Random phase shifts:** In this scheme, the phase shifts of IRS elements are randomly chosen and not optimized.

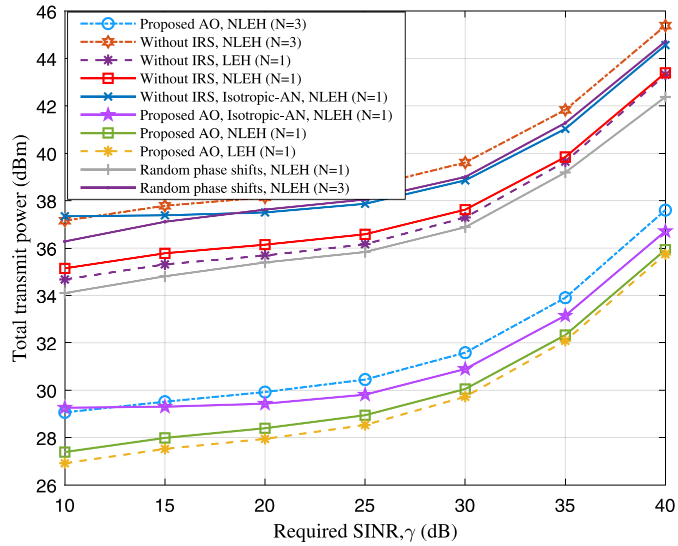


**Fig. 2.** Total transmit power versus the minimum required SINR, $\gamma$.

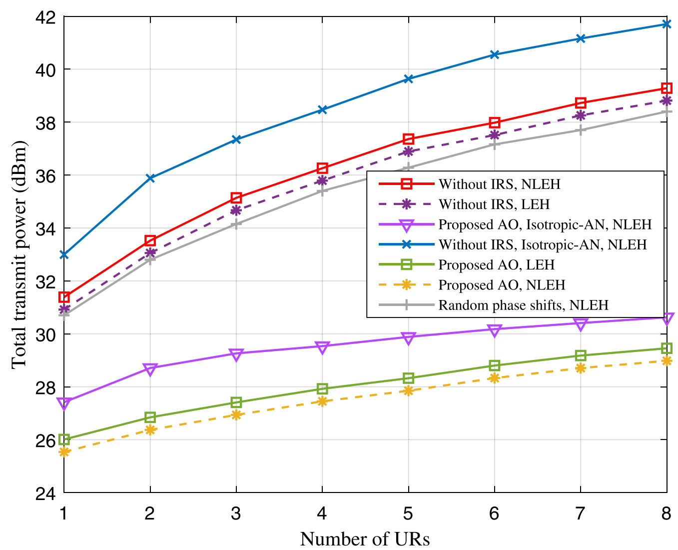


**Fig. 3.** Total transmit power versus the number of URs.

- **Without IRS:** In this scheme, there is no IRS in the system and only the BS-user links are considered.

Fig. 2 shows the total transmit power versus the minimum required SINR of the AR. The performance for both linear and non-linear EHRs are evaluated. It is observed that the total transmit power increases with the increase of the minimum required SINR of the AR. It can be seen, that the total transmit power in proposed-AO method decreases considerably and it demonstrates the advantage of using IRS. As we see, isotropic-AN design needs more power to be allocated than other methods, both for IRS case and no-IRS case, because it does not utilize CSI of URs. Furthermore, Our proposed-AO method is considerably outperforms the random phase shifts method. In multi-antenna URs scenario, it is more challenging to provide secure communication, because, the received SINR of URs is higher compared to the single antenna scenario. Therefore, the transmitter sends AN with higher power to keep SINR of URs lower than $\gamma_{e,k}$.

Fig. 3 shows the total transmit power versus the number of URs with fix $\gamma = 10$ dB for the single antenna URs. As the

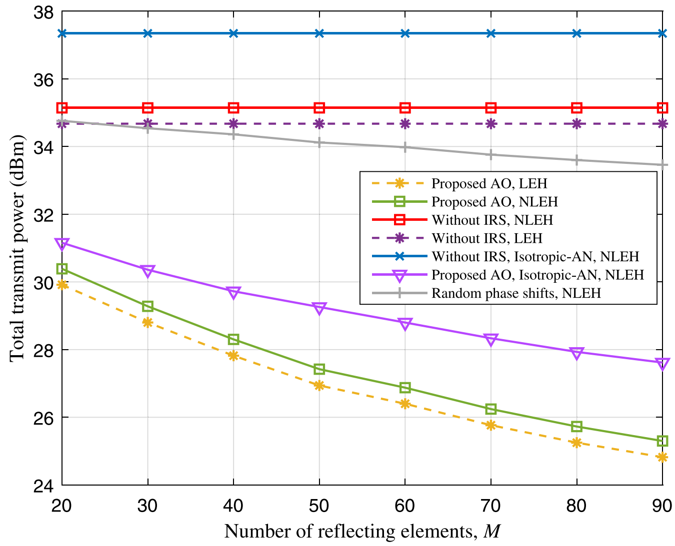


**Fig. 4.** Total transmit power versus the number of reflecting elements of IRS, $M$.

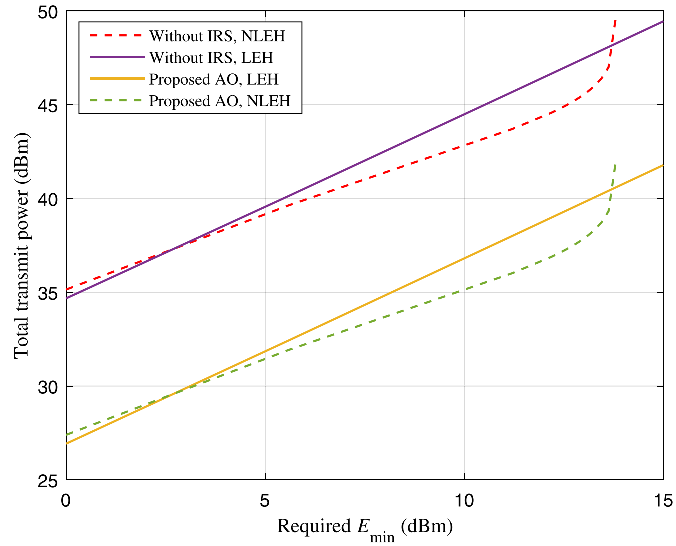


**Fig. 5.** Total transmit power versus $E_{\min}$.

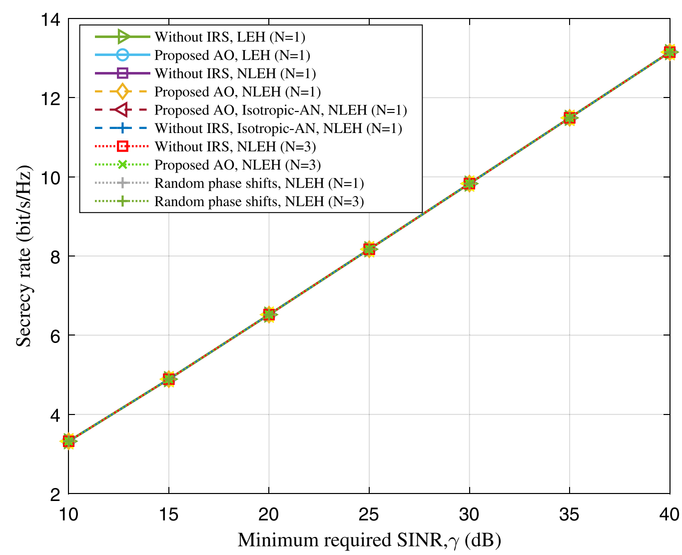


**Fig. 6.** Secrecy rate versus the minimum required SINR for AR.

number of URs increases, the more power should be allocated for AN generation to guarantee secure system and for satisfying constraints of the problem. It is worth noting that in IRS-aided case, the slope of the curves is considerably slower than no-IRS case. In other words, as the number of URs increases, the gap between these two methods increases. This indicates that employing IRS in scenarios with large number of receivers is more beneficial than scenarios with lower number of receivers. Similar to the previous figure, the Isotropic-AN design consumes more transmit power.

Fig. 4 shows the total transmit power versus the number of reflecting elements, $M$, with fix $\gamma = 10$ dB. It is obvious that in curves without IRS, the total transmit power does not change with increasing the number of reflecting elements. In contrast, it is observed that the total transmit power decreases with increasing the number of reflecting elements in IRS-aided case. In IRS-aided case, since the proposed-AO method utilizes CSI of URs' channels completely, it needs less transmit power than isotropic-AN design. In addition, the decrease of the transmit power in proposed-AO method is more than the isotropic-AN design and the gap between them tends to increase as the number of reflecting elements increases.

Fig. 5 illustrates the total transmit power versus $E_{\min}$. We set $\gamma = 10$ dB and as we expected, proposed-AO method requires less transmit power than no-IRS case. As we see, total transmit power for the non-linear model increases non-linearly with $E_{\min}$ and finally reaches saturation due to hardware limitation, while the linear model increases linearly and keeps increasing at higher $E_{\min}$ levels even after the saturation point of the non-linear model. The reason behind the dramatic increase at the end of non-linear model lies in the logarithmic function in (10). With regard to the property of logarithmic functions, in the NLEH scheme, as $E_{\min}$ gets close to the saturation point (24 mW), the parenthesis in front of the ln function in (10) gets close to zero, hence the power increase dramatically.

Finally, Fig. 6 shows the system secrecy rate in terms of the different minimum required SINR, $\gamma$, for the AR and fixed $\gamma_{e,k} = -10$ dB for the URs. As we can see, the secrecy rate increases with $\gamma$. The average secrecy rate for all schemes are the same, because in optimization problem, the equality holds for both SINR constraints for different schemes. On the other hand, the SINR threshold for URs, $\gamma_{e,k}$, is fixed for different schemes and is equal to $-10$ dB, and the SINR threshold for AR, $\gamma$, is the same in the $x$-axis of Fig. 6 for different schemes. Thus, different schemes result the same secrecy rate, but with different total transmit power. This figure confirms that our proposed AO with IRS guarantees secure system with the same secrecy rate as no-IRS case, but with consuming much lower power. This figure emphasize that the simulation results for the total transmit power of different schemes, have the same secrecy rate, and the decrease in total transmit power, does not deteriorate average secrecy rate.

## 7. Conclusion

This paper studied a robust and secure beamforming design in an IRS-assisted SWIPT system to minimize total transmit power. AN is transmitted with information signal to guarantee secure communication and provide energy for receivers. We split the optimization problem into two sub-problems and an iterative SROCR algorithm was proposed to build a rank-one solution. At the end, an AO algorithm was employed to optimize the original problem. Simulation results showed that using IRS is beneficial for energy consumption and we can achieve a desired secrecy rate by lower transmit power than no-IRS case.

## CRediT authorship contribution statement

**Mohammad Amin Parhizgar:** Conceptualization, Software, Formal analysis, Original draft, Methodology. **S. Mohammad Razavizadeh:** Review & editing, Methodology, Validation, Supervision, Project administration.

## Declaration of competing interest

The authors declare that they have no known competing financial interests or personal relationships that could have appeared to influence the work reported in this paper.

## Appendix. Proof

We define $E^{\mathrm{r}}(E_{\min}) = E_1$, and $E^{\mathrm{r}}_{e,k}(E_{\min_{e,k}}) = E_2$, and then the Lagrangian of problem (12) have

$$\begin{aligned}&\mathcal{L}(\mathbf{W},\mathbf{Z},\rho,\mathbf{Y},\mathbf{X},\xi,\alpha,\lambda,\phi)\\&=\mathrm{Tr}(\mathbf{W})+\mathrm{Tr}(\mathbf{Z})+\xi\left[\mathrm{Tr}(\mathbf{h}\mathbf{h}^H\mathbf{Z})-\frac{\mathrm{Tr}(\mathbf{h}\mathbf{h}^H\mathbf{W})}{\gamma}+\sigma_n^2+\frac{\sigma_s^2}{\rho}\right]\\&\quad+\sum_{k=1}^{K}\alpha_k\left[\mathrm{Tr}(\mathbf{g}_k\mathbf{g}_k^H\mathbf{W})-\gamma_{e,k}\left(\mathrm{Tr}(\mathbf{g}_k\mathbf{g}_k^H\mathbf{Z})+\sigma_n^2+\sigma_s^2\right)\right]\\&\quad+\lambda\left[\frac{E_1}{\eta(1-\rho)}-\mathrm{Tr}(\mathbf{h}\mathbf{h}^H\mathbf{W})-\mathrm{Tr}(\mathbf{h}\mathbf{h}^H\mathbf{Z})-\sigma_n^2\right]\\&\quad+\sum_{k=1}^{K}\phi_k\left[\frac{E_2}{\eta_{e,k}}-\mathrm{Tr}(\mathbf{g}_k\mathbf{g}_k^H\mathbf{W})-\mathrm{Tr}(\mathbf{g}_k\mathbf{g}_k^H\mathbf{Z})-\sigma_n^2\right]\\&\quad-\mathrm{Tr}(\mathbf{Y}\mathbf{W})-\mathrm{Tr}(\mathbf{X}\mathbf{Z})\end{aligned}\tag{25}$$

The dual problem of (12) is as follows

$$\max_{\substack{\xi,\alpha,\lambda,\phi\geq 0\\ \mathbf{X},\mathbf{Y}\succeq 0}}\ \min_{\mathbf{W},\mathbf{Z},\rho}\mathcal{L}(\mathbf{W},\mathbf{Z},\rho,\mathbf{Y},\mathbf{X},\xi,\alpha,\lambda,\phi)\tag{26}$$

Then the KKT conditions for $\mathbf{W}$ can be expressed as

$$\begin{aligned}\frac{\partial\mathcal{L}}{\partial\mathbf{W}}&=\mathbf{I}-\mathbf{Y}^*+\sum_{k=1}^{K}\alpha_k^*\mathbf{g}_k\mathbf{g}_k^H-\sum_{k=1}^{K}\phi_k^*\mathbf{g}_k\mathbf{g}_k^H\\&\quad-\frac{\xi^*}{\gamma}\mathbf{h}\mathbf{h}^{\mathbf{H}}-\lambda^*\mathbf{h}\mathbf{h}^{\mathbf{H}}=0,\quad\forall k\end{aligned}\tag{27a}$$

$$\mathbf{Y}^*\mathbf{W}^*=0\tag{27b}$$

$$\mathbf{Y}^*\succeq 0,\quad \xi^*,\alpha_k^*,\lambda^*,\phi_k^*\geq 0,\quad\forall k.\tag{27c}$$

We can rewrite (27a) as follows

$$\mathbf{I}+\sum_{k=1}^{K}(\alpha_k^*-\phi_k^*)\mathbf{g}_k\mathbf{g}_k^H=(\frac{\xi^*}{\gamma}+\lambda^*)\mathbf{h}\mathbf{h}^H+\mathbf{Y}^*,\quad\forall k\tag{28}$$

On the other hand, for finding the optimal PS ratio, we can solve the following problem

$$\min_{\rho}\ \frac{\lambda^*E_1}{\eta(1-\rho)}+\frac{\xi^*\sigma_s^2}{\rho},\tag{29}$$

and the optimal solution is as following

$$\rho^*=\frac{\sqrt{\eta\xi^*\sigma_s^2}}{\sqrt{\lambda^*E_1}+\sqrt{\eta\xi^*\sigma_s^2}}.\tag{30}$$

From (30), $\lambda^*>0$ and $\xi^*>0$ must hold. Because, if we assume $\lambda^*=0$ and $\xi^*=0$, the constraints in (11b) and (11d) lead to contradiction, respectively. Subsequently, similar to [15], by post-multiplying $\mathbf{W}^*$ by both sides of (27a) and considering (27b), we have

$$\left(\mathbf{I}+\sum_{k=1}^{K}(\alpha_k^*-\phi_k^*)\mathbf{g}_k\mathbf{g}_k^H\right)\mathbf{W}^*=\left((\frac{\xi^*}{\gamma}+\lambda^*)\mathbf{h}\mathbf{h}^{\mathbf{H}}\right)\mathbf{W}^*\tag{31}$$

Then we can rewrite (31) as follows

$$\mathbf{W}^*=\left(\mathbf{I}+\sum_{k=1}^{K}(\alpha_k^*-\phi_k^*)\mathbf{g}_k\mathbf{g}_k^H\right)^{-1}\left((\frac{\xi^*}{\gamma}+\lambda^*)\mathbf{h}\mathbf{h}^{\mathbf{H}}\right)\mathbf{W}^*\tag{32}$$

Then, we have

$$\begin{aligned}\mathrm{rank}(\mathbf{W}^*)&=\mathrm{rank}\left[\left(\mathbf{I}+\sum_{k=1}^{K}(\alpha_k^*-\phi_k^*)\mathbf{g}_k\mathbf{g}_k^H\right)^{-1}\right.\\&\quad\left.\times\left((\frac{\xi^*}{\gamma}+\lambda^*)\mathbf{h}\mathbf{h}^{\mathbf{H}}\right)\mathbf{W}^*\right]\leq\mathrm{rank}(\mathbf{h}\mathbf{h}^{\mathbf{H}})\leq 1\end{aligned}\tag{33}$$

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