Secrecy Energy Efficiency for Distributed-IRSs Assisted Uplink Networks

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***Abstract*—In this correspondence, we aim at achieving the energy-efficient secrecy transmission for an uplink network assisted by distributed intelligent reflecting surfaces (D-IRSs). The secrecy energy efficiency (SEE) maximization problem is investigated through jointly optimizing the discrete reflecting phase shifts, the transmit power, and the IRS switching status. To handle this non-convex problem, we decompose it into three subproblems, and then the majorization-minimization based algorithm, the Dinkelbach method and the Lagrangian dual method are adopted to obtain the solutions of reflecting phase shifts, transmit power and IRS switching status, respectively. The subproblems are alternatively optimized based on the iterative algorithm. Simulation results show that the proposed scheme can significantly improve the SEE compared to the centralized IRS, especially when the eavesdropping channel is stronger.**

***Index Terms*—Intelligent reflecting surface, secrecy energy efficiency, physical layer security, Lagrangian dual method.**

I. Introduction

Physical layer security (PLS) is a promising candidate to realize the secure information transmission and has received extensive research interests. However, some technologies based on PLS, such as artificial noise, cooperative jamming and anti-jamming [1], [2], will introduce additional power consumption, which motivates researchers to study the security problems from an energy-efficient perspective. Accordingly, secrecy energy efficiency (SEE), in terms of the amount of information that can be securely transmitted per joule of energy consumed, is emerging as another important performance metric [3].

Recently, the intelligent reflecting surface (IRS) has been regarded as a promising technique to effectively enhance the security performance [4]–[6] and energy efficiency [7], [8] by controlling reflecting phase shifts. However, most of the existing work has focused on the centralized IRS (C-IRS) assisted systems. Relative limited work has discussed the scenarios with distributed IRSs (D-IRSs). In fact, deploying D-IRSs is a more attractive solution since D-IRSs can provide multiple paths for the received signals to improve the channel strength and the robustness of transmission [9]–[13]. The power and reflecting coefficient were jointly optimized to maximize the achievable sum rate in [10], which illustrated the admirable performance of D-IRSs compared to traditional C-IRS. In [11], Zhang *et al*. investigated a D-IRSs assisted non-orthogonal multiple access network with an eavesdropper, and discovered that the D-IRSs scheme can achieve better secrecy rate (SR) due to the higher channel diversity. In [12], Dong *et al*. investigated a double IRS-aided network with inter-surface signal reflection, where the active beamforming of transmitter and the phase-shift coefficients at the double IRSs are jointly optimized to maximize the network secrecy rate. In [13], Qiao *et al*. studied the blocking-based path loss of two types of eavesdropping attacks in a multi-IRS-assisted terahertz (THz) system. Simulation results showed that this scheme can effectively enhance the secrecy performance compared with the single IRS.

On the other hand, when all the IRSs are operating, more energy will be consumed. Some prior works have tried to improve energy efficiency by dynamically switching IRSs [14], [15]. In [14], Xiu *et al*. considered a secure beamforming design to maximize the SR by controlling the on-off status of each IRS. In [15], the resource allocation problem was investigated by Yang *et al*. for D-IRSs assisted multiple-input single-output networks, to maximize the energy efficiency by turning on or off the IRS dynamically.

Motivated by the above works, we can deduce that when D-IRSs are adopted to help fight against the eavesdropping, activating all the IRSs may lead to low SEE due to improperly deployed IRSs. Thus, in order to enhance the SEE, we need to turn on the IRSs that are more suitable to assist the secure transmission. For clarity, we summarize the main contributions of this work as follows.

(1) In this correspondence, we propose a secrecy energy efficiency maximization scheme for a D-IRSs-assisted uplink network by jointly optimizing the reflecting phase shifts, transmit power, and IRS switching status, respectively. To the best of our knowledge, this is the first work to investigate the SEE maximization for an uplink system assisted by D-IRSs.

(2) In order to solve the non-convex problem, we propose a low-complexity algorithm, which decomposed the original problem into three subproblems. Specifically, the majorization-minimization (MM) based algorithm and the Dinkelbach method is developed to optimize the reflecting phase shifts

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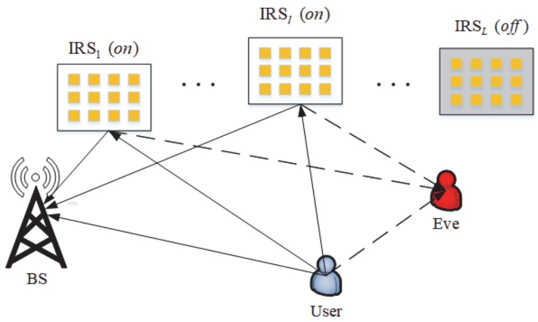

Fig. 1. Illustration of secure uplink system with multiple IRSs.

and transmit power, respectively, and we use the Lagrangian dual method to obtain the optimal solution of IRS switching status.

(3) Simulation results verify the importance of optimizing the D-IRSs switching status to improve SEE, and show that the proposed scheme outperforms the C-IRS scheme in terms of SEE performance, especially when the eavesdropping channel is stronger.

II. System Model and Problem Formulation

We present the uplink system model assisted by D-IRSs, which includes one base station (BS), L IRSs, one legitimate user (User), and one eavesdropper (Eve), as shown in Fig. 1. Define $\mathcal{L}$ and $\mathcal{N}_l$ as the set of indices of IRSs and reflecting elements of the l-th IRS (IRS_l), respectively, i.e., $l \in \mathcal{L} \triangleq \{1,2,...,L\}$, $n \in \mathcal{N}_l \triangleq \{1,2,...,N_l\}$, and $N_l = N, l = 1,...,L$. All the nodes except IRSs are equipped with a single antenna.

When User transmits the confidential message, the BS (Eve) receives the signals including both the direct link from User and the reflecting links through IRSs. The received signal at the BS (Eve) is $y_{uk} = \sqrt{P}(\sum_{l=1}^{L} x_l \mathbf{h}_{rlk}^H \mathbf{\Theta}_l \mathbf{h}_{url} + h_{uk})s + n_k, k \in \{b,e\}$, where P denotes User's transmit power. $x_l \in \{0,1\}$, and $x_l = 1$ means IRS_l is on, otherwise IRS_l is off. $n_k \sim \mathcal{CN}(0,\sigma^2)$ represents the additive white Gaussian noise (AWGN) at the BS (Eve), and s denotes the information symbol. $\mathbf{h}_{url} \in \mathbb{C}^{N\times 1}$ is the User-to-IRS_l channel, $\mathbf{h}_{rlk} \in \mathbb{C}^{N\times 1}$ is the IRS_l-to-BS (Eve) channel, and h_{uk} is the direct channel from User to the BS (Eve). $\mathbf{\Theta}_l = \text{diag}\{e^{j\phi_{l1}}, e^{j\phi_{l2}}, ..., e^{j\phi_{lN}}\} \in \mathbb{C}^{N\times N}$ is a diagonal matrix with ϕ_{ln} as the phase shift incurred by the n-th reflecting element of IRS_l.

Assume that the direct channel from User to BS (Eve) is

$$h_{uk} = \sqrt{L_0 d_{uk}^{-\alpha_1}} g_{uk}, k \in \{b,e\}, \tag{1}$$

where L_0 represents the path-loss constant in unit meter, d_{uk} is the distance from User to the BS (Eve), α_1 indicates the path attenuation exponent of the channel from User to the BS (Eve), and g_{uk} is the Rayleigh fading component. The IRS related channels can be written as

$$\mathbf{h}_j = \sqrt{L_0 d_j^{-\alpha_2}} \left(\sqrt{\frac{\beta_j}{1+\beta_j}} \mathbf{g}_j^{\text{LoS}} + \sqrt{\frac{1}{1+\beta_j}} \mathbf{g}_j^{\text{NLoS}} \right), j \in \{url, rlb, rle\}, \tag{2}$$

where β_j is the Rician factor of the corresponding channel, $\mathbf{g}_j^{\text{LoS}}$ and $\mathbf{g}_j^{\text{NLoS}}$ represent the deterministic line-of-sight (LoS) component and the Rayleigh fading component, respectively. Due to the obstructions blocking the signal transmission, there is almost no LoS component for the direct links. In this case, the direct path is usually modeled as the Rayleigh fading. As for the IRS related reflecting links, the IRS is usually installed on the building surfaces or high walls, resulting in little shading. Therefore, we consider the Rician fading channel model for the IRS related reflecting links, where both LoS and NLoS components are involved in the reflecting channels with high probability [4], [16].

The phase shift is discretized for each element of IRSs. The number of phase-shift levels is 2^b, and b denotes the number of bits. Thus, the set of discrete phase-shift values at each element is $\mathcal{F} = \{0, \Delta\phi, ..., (2^b - 1)\Delta\phi\}$, and $\Delta\phi = 2\pi/2^b$.

Based on the aforementioned expressions, the achievable rate at the BS (Eve) can be expressed as

$$R_k = \log_2\left(1 + \rho \left| \sum_{l=1}^{L} x_l \mathbf{h}_{rlk}^H \mathbf{\Theta}_l \mathbf{h}_{url} + h_{uk} \right|^2\right), k \in \{b,e\}, \tag{3}$$

where $\rho = P/\sigma^2$. According to (3), the secrecy rate can be given by $R_s = [R_b - R_e]^+$, and $[x]^+ \triangleq \max(x,0)$.

The system energy consumption can be divided into three parts: the transmit power of User P, the circuit power consumption of BS (P_B) and User (P_U) P_g, and the total power consumption of all elements of D-IRSs. Thus, the total power consumption of the legitimate system is $P_t = P + P_g + \sum_{l=1}^{L} x_l N P_n(b)$, where $P_n(b)$ is the energy consumed by the n-th element of IRS_l, which is related to the number of bits b for phase shifts [7].

To make the secure transmission more energy-efficient, we maximize the SEE by jointly optimizing the reflecting phase shifts, the transmit power, and the IRS switching status. The optimization problem can be formulated as

$$\max_{\mathbf{\Theta},\mathbf{P},\mathbf{x}} \frac{\log_2\left(1+\rho\left|\sum_{l=1}^{L} x_l h_{Bl} + h_{ub}\right|^2\right) - \log_2\left(1+\rho\left|\sum_{l=1}^{L} x_l h_{El} + h_{ue}\right|^2\right)}{P + P_g + \sum_{l=1}^{L} x_l N P_n(b)} \tag{4a}$$

$$s.t. \quad P \le P_{\max}, \tag{4b}$$

$$\phi_{ln} \in [0, 2\pi), \quad \forall l \in \mathcal{L}, \forall n \in \mathcal{N}_l, \tag{4c}$$

$$x_l \in \{0,1\}, \quad \forall l \in \mathcal{L}, \tag{4d}$$

where $\mathbf{\Theta} = \{\mathbf{\Theta}_1; ...; \mathbf{\Theta}_l; ...; \mathbf{\Theta}_L\}$, $\mathbf{x} = \{x_1, ..., x_l, ..., x_L\} \in \mathbb{C}^{1\times L}$, $P_{\max}$ is the maximum transmit power of User, $h_{Bl} = \mathbf{h}_{rlb}^H \mathbf{\Theta}_l \mathbf{h}_{url}$, and $h_{El} = \mathbf{h}_{rle}^H \mathbf{\Theta}_l \mathbf{h}_{url}$. (4b) and (4c) limit the maximum value of User's transmit power and the phase of each reflecting element, respectively. The problem (4) is a mixed-integer nonlinear program which is difficult to solve.

III. SEE Optimization With D-IRSs

In this section, an iterative algorithm is proposed to solve the problem (4). Specifically, the original problem is decomposed into three subproblems, which are solved iteratively until convergence.

A. Optimization of Reflecting Phase Shifts

With the given transmit power and IRS switching status, we first optimize the reflecting phase shifts. For simplicity, rewrite $\sum_{l=1}^{L} x_l h_{kl} = \mathbf{v}^H \mathbf{g}_k, k \in \{B, E\}$, where

$$\mathbf{v}^H = [\boldsymbol{\vartheta}_1; ...; \boldsymbol{\vartheta}_L]^T \in \mathbb{C}^{1\times Q}, \mathbf{g}_k = [x_1 \mathbf{g}_{r1k}; ...; x_L \mathbf{g}_{rLk}] \in \mathbb{C}^{Q\times 1},$$
$$\boldsymbol{\vartheta}_l = \left[e^{j\phi_{l1}}, ..., e^{j\phi_{lN}}\right]^T \in \mathbb{C}^{N\times 1}, \mathbf{g}_{rlk} = \mathrm{diag}\left(\mathbf{h}_{rlk}^H\right)\mathbf{h}_{url} \in \mathbb{C}^{N\times 1}.$$

Accordingly, the problem (4) can be changed into

$$\min_{\mathbf{v}} \ \frac{\rho\left|\mathbf{v}^H \mathbf{g}_E + h_{ue}\right|^2 + 1}{\rho\left|\mathbf{v}^H \mathbf{g}_B + h_{ub}\right|^2 + 1} \tag{5a}$$
$$s.t. \ \ |v_i| = 1, i = 1, 2, ..., Q, \tag{5b}$$

where $v_i = e^{j\phi_{ln}}$, and Q is the sum of all the elements of IRSs, i.e., $L \times N = Q$.

Due to the non-convexity, we transform (5) into a subtraction form by the Dinkelbach method. By introducing a parameter $\mu > 0$, (5) can be rewritten as

$$\min_{\mathbf{v}} \ \rho\left|\mathbf{v}^H \mathbf{g}_E + h_{ue}\right|^2 + 1 - \mu\left(\rho\left|\mathbf{v}^H \mathbf{g}_B + h_{ub}\right|^2 + 1\right) \tag{6a}$$
$$s.t. \ \ |v_i| = 1, i = 1, 2, ..., Q. \tag{6b}$$

The problem is still non-convex, and the MM based algorithm is applied to transform (6). According to [17], the upper bound of the objective function (6a) can be denoted by

$$H(\mu) = \rho\mathbf{v}^H\left(\mathbf{g}_E\mathbf{g}_E^H - \mu\mathbf{g}_B\mathbf{g}_B^H\right)\mathbf{v} - 2\rho\Re\left\{\mathbf{v}^H\left(\mu\mathbf{g}_B h_{ub}^* - \mathbf{g}_E h_{ue}^*\right)\right\}$$
$$+ \rho|h_{ue}|^2 + 1 - \mu\rho|h_{ub}|^2 - \mu \leq -2\rho\Re\left\{\mathbf{v}^H\boldsymbol{\beta}\right\} + c, \tag{7}$$

where $c = \rho\left[\left(\mathbf{v}^H\mathbf{X}\mathbf{v} + (\mathbf{v}^{(k)})^H(\mathbf{X} - \mathbf{H}_D)\mathbf{v}^{(k)} + |h_{ue}|^2 - \mu|h_{ub}|^2\right)\right] + 1 - \mu$, $\boldsymbol{\beta} = (\mathbf{X} - \mathbf{H}_D)\mathbf{v}^{(k)} - \mathbf{g}_E h_{ue}^* + \mu\mathbf{g}_B h_{ub}^*$, $\mathbf{X} = \lambda_{max}(\mathbf{H}_D)\mathbf{I}_Q$ and $\mathbf{H}_D = \mathbf{g}_E\mathbf{g}_E^H - \mu\mathbf{g}_B\mathbf{g}_B^H$. $\mathbf{v}^{(k)}$ is the k-th iteration of $\mathbf{v}$. The corresponding optimization problem can be rewritten as

$$\min_{\mathbf{v}} \ -2\rho\Re\left\{\mathbf{v}^H\boldsymbol{\beta}\right\} + c \tag{8a}$$
$$s.t. \ \ |v_i| = 1, i = 1, 2..., Q. \tag{8b}$$

We can figure out that $-\Re\{\mathbf{v}^H\boldsymbol{\beta}\}$ takes the minimum value only when the phases of v_i and β_i are equal. Therefore, we can obtain the optimal solution to (6) as

$$\mathbf{v}^*(\mu) = \left(\exp\left(j\arg\left(\boldsymbol{\beta}\right)\right)\right)^T, \tag{9}$$

and continue to discretize $\mathbf{v}$ by replacing it with the phase shifts in $\mathcal{F}$.

By alternatively updating $\mathbf{v}$ and $\mu = \frac{\rho\left|\mathbf{v}^H\mathbf{g}_E + h_{ue}\right|^2 + 1}{\rho\left|\mathbf{v}^H\mathbf{g}_B + h_{ub}\right|^2 + 1}$ until $H(\mu)$ converges, the solution to (5) can be obtained.

B. Transmit Power Optimization

With fixed phase shifts and IRS switching status, we optimize the transmit power of User. Defining $t_1 = \left|\mathbf{v}^H\mathbf{g}_B + h_{ub}\right|^2$ and $t_2 = \left|\mathbf{v}^H\mathbf{g}_E + h_{ue}\right|^2$, the problem (4) can be reduced as

$$\max_{P} \ \frac{\log_2(\rho t_1 + 1) - \log_2(\rho t_2 + 1)}{P + P_g + \sum_{l=1}^{L} x_l N P_n(b)} \tag{10a}$$
$$s.t. \ \ P \leq P_{max}. \tag{10b}$$

By introducing a parameter $\lambda > 0$, (10a) can be rewritten as

$$\log_2(\rho t_1 + 1) - \log_2(\rho t_2 + 1) - \lambda\left(P + P_g + \sum_{l=1}^{L} x_l N P_n(b)\right). \tag{11}$$

Denote (11) as $H(\lambda)$. Proposition 1 is presented to calculate the solution to (10) as follows.

Proposition 1: We consider the following two cases to get the optimal solution of P.

Case 1: If $t_1 - t_2 > 0$, the optimal solution to (10) with given λ can be expressed as

$$P^*(\lambda) = \begin{cases} P(t_1, t_2), & \text{if } 0 < P(t_1, t_2) < P_{max}, \\ P_{max}, & \text{if } P(t_1, t_2) \geq P_{max}, \end{cases} \tag{12}$$

where $P(t_1, t_2) = \frac{-\sigma^2(t_1 + t_2) + \sqrt{\sigma^4(t_1 - t_2)^2 + 4t_1t_2\frac{\sigma^2(t_1 - t_2)}{\lambda \ln 2}}}{2t_1t_2}$.

Case 2: If $t_1 - t_2 \leq 0$, $R_s \leq 0$, and we have SEE=0.

Proof: To identify the concavity of $H(\lambda)$, we first calculate its second-order derivative with respect to P, the sign of which is equivalent to $\frac{\sigma^2(t_2 - t_1)}{(Pt_2 + \sigma^2)(Pt_1 + \sigma^2)}$. It can be deduced that $H(\lambda)$ is strictly concave when $t_1 - t_2 > 0$. Otherwise, $R_s \leq 0$, and SEE$= 0$. Thus, the root of $\mathrm{d}H(\lambda)/\mathrm{d}P = 0$ is the optimal solution to (10) as

$$\frac{\mathrm{d}H(\lambda)}{\mathrm{d}P} = \frac{t_1}{\ln 2(Pt_1 + \sigma^2)} - \frac{t_2}{\ln 2(Pt_2 + \sigma^2)} - \lambda = 0. \tag{13}$$

Since (13) is a quadratic equation with respect to P, its solution can be derived as $P(t_1, t_2)$.

If $P(t_1, t_2) < 0$, it means that the legitimate channel condition is poor or the gap between the legitimate and eavesdropping channels is not large, and we set $P = 0$. ■

By alternatively updating P and $\lambda = \frac{(s-q)\log_2(e)}{P + P_g + \sum_{l=1}^{L} x_l N P_n(b)}$ until $H(\lambda)$ converges, the solution to (10) can be obtained.

C. Optimization of IRS Switching Status

After obtaining the solutions of reflecting phase shifts and transmit power, the problem related to the IRS switching status can be formulated as

$$\max_{\mathbf{x}} \ \frac{\log_2\left(1 + \rho\left|\sum_{l=1}^{L} x_l h_{Bl} + h_{ub}\right|^2\right) - \log_2\left(1 + \rho\left|\sum_{l=1}^{L} x_l h_{El} + h_{ue}\right|^2\right)}{P + P_g + \sum_{l=1}^{L} x_l N P_n(b)} \tag{14a}$$
$$s.t. \ \ x_l \in \{0, 1\}, \ \ \forall l \in \mathcal{L}. \tag{14b}$$

To handle the non-convex problem (14), two auxiliary variables s and q, and a non-negative parameter λ are introduced. Thus, the problem (14) can be converted as

$$\max_{\mathbf{x}, s, q} \ (s - q)\log_2(e) - \lambda\left(P + P_g + \sum_{l=1}^{L} x_l N P_n(b)\right) \tag{15a}$$
$$s.t. \ \ \rho\left|\sum_{l=1}^{L} x_l h_{Bl} + h_{ub}\right|^2 + 1 \geq e^s, \tag{15b}$$
$$\rho\left|\sum_{l=1}^{L} x_l h_{El} + h_{ue}\right|^2 + 1 \leq e^q, \tag{15c}$$
$$x_l \in \{0, 1\}, \ \ \forall l \in \mathcal{L}. \tag{15d}$$

Even though (15a) is convex, the constraints (15b) and (15c) are still non-convex. We first transform them as

$$\left|\sum_{l=1}^{L} x_l h_{Bl} + h_{ub}\right|^2 = B_0 + \sum_{l=1}^{L} B_l x_l + \sum_{l=2}^{L}\sum_{m=1}^{L-1} B_{lm} x_l x_m, \quad (16a)$$

$$\left|\sum_{l=1}^{L} x_l h_{El} + h_{ue}\right|^2 = E_0 + \sum_{l=1}^{L} E_l x_l + \sum_{l=2}^{L}\sum_{m=1}^{L-1} E_{lm} x_l x_m, \quad (16b)$$

where $B_0 = h_{ub}h_{ub}^H, E_0 = h_{ue}h_{ue}^H$,

$B_l = \mathbf{h}_{rlb}^H \boldsymbol{\Theta}_l \mathbf{h}_{url}\mathbf{h}_{url}^H \boldsymbol{\Theta}_l^H \mathbf{h}_{rlb} + h_{ub}\mathbf{h}_{url}^H \boldsymbol{\Theta}_l^H \mathbf{h}_{rlb} + \mathbf{h}_{rlb}^H \boldsymbol{\Theta}_l \mathbf{h}_{url} h_{ub}^H$,

$E_l = \mathbf{h}_{rle}^H \boldsymbol{\Theta}_l \mathbf{h}_{url}\mathbf{h}_{url}^H \boldsymbol{\Theta}_l^H \mathbf{h}_{rle} + h_{ue}\mathbf{h}_{url}^H \boldsymbol{\Theta}_l^H \mathbf{h}_{rle} + \mathbf{h}_{rle}^H \boldsymbol{\Theta}_l \mathbf{h}_{url} h_{ue}^H$,

$B_{lm} = \mathbf{h}_{rlb}^H \boldsymbol{\Theta}_l \mathbf{h}_{url}\mathbf{h}_{urm}^H \boldsymbol{\Theta}_m^H \mathbf{h}_{rmb} + \mathbf{h}_{rmb}^H \boldsymbol{\Theta}_m \mathbf{h}_{urm}\mathbf{h}_{url}^H \boldsymbol{\Theta}_l^H \mathbf{h}_{rlb}$,

$E_{lm} = \mathbf{h}_{rle}^H \boldsymbol{\Theta}_l \mathbf{h}_{url}\mathbf{h}_{urm}^H \boldsymbol{\Theta}_m^H \mathbf{h}_{rme} + \mathbf{h}_{rme}^H \boldsymbol{\Theta}_m \mathbf{h}_{urm}\mathbf{h}_{url}^H \boldsymbol{\Theta}_l^H \mathbf{h}_{rle}$.

To handle the coupled item $x_l x_m$, we introduce a variable z_{lm} satisfying the following constraints due to $x_l \in \{0,1\}$.

$$z_{lm} \geq x_l + x_m - 1, 0 \leq z_{lm} \leq 1, z_{lm} \leq x_l, z_{lm} \leq x_m, \quad (17)$$

where $z_{lm} = \left[z_{21}, z_{31}, z_{32}, ..., z_{L(L-1)}\right]_{l=2,...,L,m=1,...,L-1}$.

Substituting (16) and $z_{lm} = x_l x_m$ into (15), (15b) and (15c) can be rewritten as

$$\rho\underbrace{\left(B_0 + \sum_{l=1}^{L} B_l x_l + \sum_{l=2}^{L}\sum_{m=1}^{L-1} B_{lm} z_{lm}\right)}_{B(\mathbf{x},\mathbf{z})} + 1 \geq e^s, \quad (18a)$$

$$\rho\underbrace{\left(E_0 + \sum_{l=1}^{L} E_l x_l + \sum_{l=2}^{L}\sum_{m=1}^{L-1} E_{lm} z_{lm}\right)}_{E(\mathbf{x},\mathbf{z})} + 1 \leq e^q. \quad (18b)$$

It is easy to find that (18a) is convex, but (18b) is still non-convex. Thus, we expand e^q in the first-order Taylor at $\overline{q}$ as

$$e^q \geq e^{\overline{q}} + e^{\overline{q}}\left(q - \overline{q}\right) = \mathcal{L}(q, \overline{q}). \quad (19)$$

According to the above mathematical transformations, we can rewrite the problem (15) as

$$\max_{\mathbf{x},\mathbf{z},s,q} \ (s-q)\log_2(e) - \lambda\left(P + P_g + \sum_{l=1}^{L} x_l N P_n(b)\right) \quad (20a)$$

$$s.t. \ \rho B(\mathbf{x},\mathbf{z}) + 1 \geq e^s, \quad (20b)$$

$$\rho E(\mathbf{x},\mathbf{z}) + 1 \leq \mathcal{L}(q, \bar{q}), \quad (20c)$$

$$z_{lm} \geq x_l + x_m - 1, z_{lm} \leq x_l, z_{lm} \leq x_m, \quad (20d)$$

$$0 \leq z_{lm} \leq 1, \ \forall l = 2, ..., L, m = 1, ..., L-1, \quad (20e)$$

$$x_l \in \{0,1\}, \ \forall l \in \mathcal{L}. \quad (20f)$$

By relaxing (20f) as $0 \leq x_l \leq 1$, (20) becomes a standard convex optimization problem. The Lagrangian expression corresponding to (20) can be denoted as

$$\begin{aligned}
&\mathrm{L}\left(\mathbf{x}, \mathbf{z}, \boldsymbol{\gamma}, \alpha, \beta, s, q\right) \\
&= (s-q)\log_2(e) - \lambda\left(P + P_g + \sum_{l=1}^{L} x_l N P_n(b)\right) \\
&+ \alpha\left[\rho B(\mathbf{x},\mathbf{z}) + 1 - e^s\right] - \beta\left[\rho E(\mathbf{x},\mathbf{z}) + 1 - \mathcal{L}(q, \bar{q})\right] \\
&+ \sum_{l=2}^{L}\sum_{m=1}^{L-1}\left[\gamma_{1lm}(z_{lm} - x_l - x_m + 1)\right. \\
&\left. + \gamma_{2lm}(x_l - z_{lm}) + \gamma_{3lm}(x_m - z_{lm})\right], \quad (21)
\end{aligned}$$

where $\boldsymbol{\gamma} = \{\gamma_{1lm}, \gamma_{2lm}, \gamma_{3lm}\}_{l=2,...,L,m=1,...,L-1}$, and α and β are non-negative Lagrangian multipliers. Accordingly, the Lagrangian dual problem related to (20) can be formulated as

$$\min_{\alpha,\beta,\boldsymbol{\gamma}} \ \Gamma\left(\alpha, \beta, \boldsymbol{\gamma}\right) \quad (22a)$$

$$s.t. \ 0 \leq z_{lm} \leq 1, \forall l = 2, ..., L, m = 1, ..., L-1, \quad (22b)$$

$$0 \leq x_l \leq 1, \ \forall l \in \mathcal{L}, \quad (22c)$$

where $\Gamma\left(\alpha, \beta, \boldsymbol{\gamma}\right) = \max_{\mathbf{x},\mathbf{z},s,q} \ \mathrm{L}\left(\mathbf{x}, \mathbf{z}, \boldsymbol{\gamma}, \alpha, \beta, s, q\right)$.

In order to obtain the optimal solution to (22), we first introduce the following proposition.

Proposition 2: For the dual problem (22), the optimal solutions of x_l and z_{lm} are given by

$$x_l = \begin{cases} 1, \ \text{if} \ \ C_l > 0, \\ 0, \ \text{otherwise}, \end{cases} \quad (23)$$

$$z_{lm} = \begin{cases} 1, \ \underbrace{\alpha\rho B_{lm} - \beta\rho E_{lm} + \gamma_{1lm} - \gamma_{2lm} - \gamma_{3lm}}_{C_{lm}} > 0, \\ 0, \ \text{otherwise}, \end{cases} \quad (24)$$

where

$$C_l = \begin{cases} -\lambda N P_n(b) + \alpha\rho B_1 - \beta\rho E_1 + \sum_{m=2}^{L}(\gamma_{3ml} - \gamma_{1ml}), \ \text{if} \ l=1, \\ -\lambda N P_n(b) + \alpha\rho B_l - \beta\rho E_l + \sum_{m=1}^{L-1}(\gamma_{2lm} - \gamma_{1lm}) \\ + \sum_{m=l+1}^{L}(\gamma_{3ml} - \gamma_{1ml}), \ \text{if} \ 2 \leq l \leq L-1, \\ -\lambda N P_n(b) + \alpha\rho B_L - \beta\rho E_L + \sum_{m=1}^{L-1}(\gamma_{2lm} - \gamma_{1lm}), \ \text{if} \ l=L. \end{cases}$$

Proof: By taking the first-order derivative of (21), we find that it is a linear function of x_l and z_{lm}. In order to maximize (21), x_l and z_{lm} should be taken as 1 when their corresponding coefficient is positive, i.e., $C_l > 0$ and $C_{lm} > 0$. ■

To elaborate the update of $\boldsymbol{\gamma}$, we take $\gamma_{1lm}^{(k)}$ for example. If $z_{lm}^{(k)} \geq x_l + x_m - 1$, update $\gamma_{1lm}^{(k)}$=0; otherwise, according to (24), find $\gamma_{1lm}^{(k)}$ to satisfy $z_{lm}^{(k)} \geq x_l + x_m - 1$ by the bisection method. $\gamma_{2lm}^{(k)}$ and $\gamma_{3lm}^{(k)}$ can be updated similarly based on the constraints $x_l \geq z_{lm}^{(k)}$ and $x_m \geq z_{lm}^{(k)}$.

Based on the KKT conditions, the optimal solution of s can be derived from

$$\frac{\partial \mathrm{L}\left(\mathbf{x}, \mathbf{z}, \boldsymbol{\gamma}, \alpha, \beta, s, q\right)}{\partial s} = \log_2(e) - \alpha e^s = 0. \quad (25)$$

It can be observed that the update of s is related to α, and we can update $s^{(k)}$ and $\alpha^{(k)}$ by analyzing (20b). If $s^{(k-1)}$ satisfies (20b), we update

$$s^{(k)} = \ln\left(\log_2(e) / \alpha^{(k-1)}\right), \alpha^{(k)} = \alpha^{(k-1)}, \quad (26)$$

Otherwise, we force $s^{(k)}$ to satisfy (20b) and update

$$s^{(k)} = \ln\left(\rho B(\mathbf{x},\mathbf{z}) + 1\right), \alpha^{(k)} = \frac{\log_2(e)}{e^{s^{(k)}}}. \quad (27)$$

In addition, (21) is a linear function of q. Thus, q can be updated by using the sub-gradient method as

$$q^{(k)} = \left[q^{(k-1)} - \phi_1\left(\beta e^{q^{(k-1)}} - \log_2(e)\right)\right]^+. \quad (28)$$

We also use the sub-gradient method to update β as

$$\beta^{(k)}=\left[\beta^{(k-1)}-\phi_2\left(\mathcal{L}\left(q^{(k)},\bar{q}^{(k)}\right)-\rho E\left(\mathbf{x},\mathbf{z}\right)-1\right)\right]^{+}, \quad (29)$$

where $\bar{q}^{(k)}=q^{(k-1)}$ and $\phi_i>0$ is a step-size sequence, which adjusts dynamically to satisfy constraint (20c).

Details of solving $\mathbf{x}$ are summarized in Algorithm 1.

Algorithm 1 Lagrangian dual method for (20)

Initialize λ and $(\alpha,\beta,\bar{q})$, and set the accuracy $\varepsilon=10^{-6}$.
repeat
  Update $(\mathbf{x},\mathbf{z})$ according to Proposition 2.
  Update $\boldsymbol{\gamma}$, $(s,q,\alpha,\beta,\bar{q})$ according to (26)-(29).
  Update $\lambda=\frac{(s-q)\log_2(e)}{(P+P_g+\sum_{l=1}^{L}x_l N P_n(b))}$.
until : $H(\lambda)<\varepsilon$.

D. Algorithm Summary

Based on the above derivations, we propose an iterative algorithm to solve (4). Via alternately optimizing $\mathbf{v}$, P and $\mathbf{x}$, the suboptimal solution to (4) can be obtained. Details of the algorithm are summarized in Algorithm 2.

Algorithm 2 Iterative Algorithm for Problem (4)

Initialize $\left(\mathbf{v}^{(0)},P^{(0)},\mathbf{x}^{(0)}\right)$, and set the iteration index t=1.
repeat
  Update $\mathbf{v}^{(t)}$ with given $\left(P^{(t-1)},\mathbf{x}^{(t-1)}\right)$ by MM-based algorithm.
  Update $\left(P^{(t)},\mathbf{x}^{(t)}\right)$ with given $\mathbf{v}^{(t)}$ by Proposition 1 and Algorithm 1, respectively.
  Set $t=t+1$.
until : The objective value of (4a) converges.

Convergence analysis: The SEE in Algorithm 2 is non-decreasing with iterations. In addition, due to the transmit power constraint, the objective value of SEE has a finite upper bound. Thus, the convergence of the proposed algorithm can be guaranteed.

Computational complexity analysis: In Algorithm 2, the main complexity lies in the reflecting phase shifts and the IRS switching status, which involve the complexity of $\mathcal{O}(T_1 Q)$ and $\mathcal{O}(T_2 L^2)$, respectively. Hence, the total complexity of solving the problem (4) is $\mathcal{O}(T_0(T_1 Q+T_2 L^2))$, where T_0 is the total number of iterations for Algorithm 2, T_1 is the total number of iterations for solving the problem (5) of reflecting phase shifts and T_2 is the total number of iterations for updating λ in Algorithm 1.

IV. Simulation Results

In this section, we provide numerical results to evaluate the SEE of uplink network assisted by 3 D-IRSs. We first give the main parameters: $b=3$, $P_n(b)=1.5$ mW, $\forall n\in\mathcal{N}$, $P_B=200$ mW, $P_U=10$ mW, $P_g=210$ mW, $\sigma^2=-110$ dBm, $L_0=-30$ dBm, $\alpha_1=3.6$, $\alpha_2=2.2$ and $\beta_j=3$ dB.

Fig. 2 shows the performance of SEE and SR with the increase of P_{max}. We consider three schemes including the optimized IRS switching status (OISS), the non-optimized IRS switching status (NISS) where all IRSs are turned on, and the random IRS switching status (RISS). The coordinates of User, the BS and Eve are set as [8,0,0], [8,100,0] and [5,80,0] in meters respectively. 3 IRSs are placed at [5,0,5], [5,100,5] and [2,150,5], respectively. N is set to 16. From Fig. 2, it can be observed that the SEE performance of OISS is much better than that of the other two benchmarks. As for the SR, we can find that the SR of OISS (turn on the IRSs at [5,0,5] and [5,100,5]) is slightly lower than that of NISS, and they almost overlap with higher P_{max}. Thus, optimizing the IRS switching status is crucial to enhance the SEE for D-IRSs assisted networks.

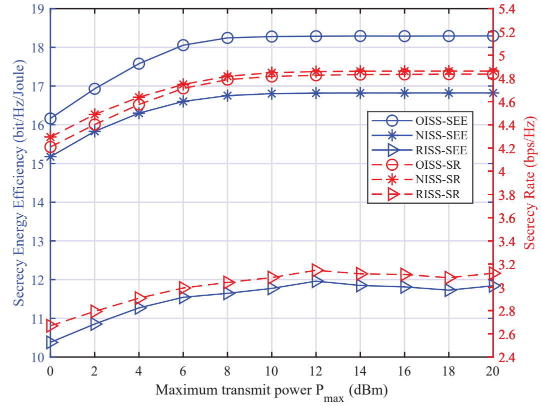

Fig. 2. SEE and SR versus the maximum transmit power P_{max} ($N=16$).

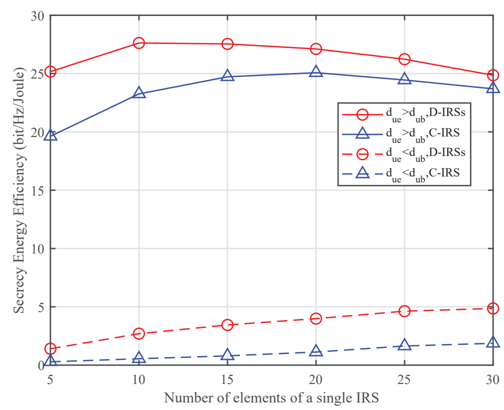

Fig. 3. SEE versus the number of elements of a single IRS.

Then, we compare the SEE performance of C-IRS and D-IRSs with the increase of N, as shown in Fig. 3. The number of elements of C-IRS is the same as the total number of elements in D-IRSs. We consider two scenarios where $d_{ub}<d_{ue}$ and $d_{ub}>d_{ue}$. In the first scenario, the coordinates of User, the BS and Eve are set as [8,40,0], [8,100,0] and [5,140,0], respectively, while in the second scenario, the coordinate of Eve is changed as [5,60,0]. In both scenarios, 3 IRSs are placed near User, the BS and Eve, respectively. Set the location of C-IRS as [5,90,5]. From Fig. 3, the SEE of D-IRSs is always superior to that of C-IRS in both scenarios. Specifically, when $d_{ub}<d_{ue}$, the SEE of D-IRSs and C-IRS both increases first and then decreases with the number of elements of a single IRS. This indicates that too many IRS elements may be harmful for the SEE. To gain more insights, we records

TABLE I
RATIO OF SEE BETWEEN D-IRSS AND C-IRS

	$N=5$	$N=10$	$N=15$	$N=20$	$N=25$	$N=30$
$d_{ub}<d_{ue}$	1.2948	1.1968	1.1090	1.0824	1.0623	1.0404
$d_{ub}>d_{ue}$	5.0843	3.9636	3.8485	3.4178	2.8565	2.5927

the ratio of SEE between D-IRSs and C-IRS in Table I. It can be observed that the performance gap between D-IRSs and C-IRS becomes smaller with N due to the stronger reflection channel of C-IRS. More interestingly, the ratio in the case of $d_{ub} > d_{ue}$ is much larger than that in the case of $d_{ub} < d_{ue}$. This reveals that the SEE of D-IRSs has more advantage over that of C-IRS where the eavesdropping channel is stronger.

V. CONCLUSIONS

In this correspondence, we have proposed an SEE maximization scheme for the D-IRSs assisted uplink network by jointly optimizing the phase shifts, transmit power and IRS switching status. Due to the non-convexity of the formulated problem, the iterative algorithm is adopted. We use the MM-based algorithm and the Lagrangian dual method to obtain the optimal solutions of the phase shifts and IRS switching status, respectively. Simulation results demonstrate the superiority of SEE performance of D-IRSs compared with C-IRS in challenging scenarios.

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