Computer Science and Information Systems 20(2):595–0000 https://doi.org/10.2298/CSIS220712011L

Security Performance Analysis of Active Intelligent Reflective Surface Assisted Wireless Communication

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Abstract. As a new communication technology, Intelligent Reflecting Surface (IRS) can intelligently reconfigure the wireless propagation environment by integrating many passive/active reflective elements on the plane. According to the characteristics that IRS can adjust the propagation channel intelligently, this paper applies IRS to wireless security communication, and studies how to make the security rate reach the optimal security capacity from the perspective of optimization technology. In this paper, two schemes of passive/active IRS are considered, and the corresponding safety rate maximization algorithm is proposed. In view of the non-convexity of the objective function, on the one hand, in the passive IRS scheme, the Dinkelbach method is used to transform the objective function into a form that is easy to handle, and the original problem is transformed into a convex problem through the continuous convex approximation method; On the other hand, under the active IRS scheme, aiming at the complexity of the original problem, the first order Taylor expansion is used to obtain the lower bound of the optimization problem, and a minimax optimization algorithm is proposed. Finally, the performance of the proposed algorithm is verified by simulation. The simulation results show that the algorithm designed with active IRS has better security rate than the algorithm designed with passive IRS under the same parameter settings.

Keywords: IRS, Physical layer Secrecy, Convex optimization.

1. Introduction

Intelligent Reflecting Surfaces (IRSs) are man-made surfaces of electromagnetic(EM) material controlled by integrated electronic devices, which have attracted widespread attention due to their unique wireless communication capabilities[1]. The development of IRS has evolved from related researches on metamaterials and metasurfaces, and current implementations include traditional reflectarray, liquid crystal surfaces, and software-defined primitive surfaces [2,3,4]. With the emergence of IRS objects, IRS technology provides a new idea for improving the channel capacity of wireless communication. Under the premise that IRS can be realized, more and more researchers analyze the positive effects of IRS on wireless communication from different research fields, such as performance optimization, channel estimation, security, etc [5].

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Specifically, the passive IRS can adjust the phase of the incident electromagnetic wave by controlling the components deployed on the IRS, thereby controlling the direction of the electromagnetic wave at the receiver. Compared with the traditional decoding and forwarding relaying technology, passive IRS can shape the wireless signals through software programming without additional energy sources, increasing the channel capacity without consuming spectrum resources. Most of the current research shows that passive IRS can improve the channel capacity of target users [6,7,8,9], but these studies ignore the "double fading" effect brought by passive IRS, and only consider that the direct link is very poor atypical scenario. Therefore, in order to overcome the basic limitation of the "double fading" effect, some researchers proposed the application of active IRS in wireless communication systems [12,13,14,15,16,17,18]. Unlike passive IRS, which passively reflects signals without amplification, active IRS can not only adjust the phase of incident electromagnetic waves, but also amplify reflected signals. However, the noise in the signal transmission is also amplified accordingly, and additional power is required. However, In comparison with the amplify-and-forward relay technology, active IRS does not require encoding and decoding operations. Reference 11 specifically expounds the concept of active IRS, its hardware structure and signal model, introduces how active IRS amplifies the incident signal, and compares and analyzes the channel capacity gain brought by active IRS and passive IRS through experimental measurements. Reference 12 introduces an active IRS into a single-input multiple-output system, and maximizes the received signal power of the system and minimizes the IRS-related noise at the receiver by optimizing the reflection coefficient matrix of the IRS and the receiving beamforming at the receiver. The results prove that the active IRS- assisted system has better performance than the traditional passive IRS- assisted system under the same power budget. However, if the two systems have the same total power budget and the number of reflection units of the IRS is sufficient, the channel capacity brought by the passive IRS is better than that of the active IRS[13]. References 14,15,16 also demonstrate the superiority of active IRS over passive IRS. It is worth noting that, unlike the existing passive IRS, which passively reflects the signal without amplitude amplification, the active IRS requires extra power to support the active amplification of the reflected signal, so with the same number of reflective elements, Active IRS requires a larger power budget than passive IRS [17]. In order to reduce the power loss of active IRS, the literature 18 proposes a sub-connection architecture of active IRS. At the cost of reduced degrees of freedom required for beamforming design, the components on the IRS independently control their phase shifts, but share the same power amplifier, the simulation results demonstrate that the sub-connection structure is an energy-efficient implementation of the active IRS. More importantly, the team of Mr. Dai from Tsinghua University has a mature IRS hardware design method[19], which has promoted more and more researchers to study the beneficial effects of IRS on wireless communication from different fields.

With the rapid development of the Internet of Things and the continuous upgrading of various eavesdropping methods, the disadvantages brought by traditional security technologies are also increasing. The emergence of IRS technology provides a new direction for PLS research. Although PLS for different wireless systems has been extensively studied in many literatures [20,21,22,23], the application of IRS to improve PLS is still not well studied. Through literature review, it is found that most authors consider passive IRS to improve the PLS of the system. As in the literature

24,25,26,27,28,29,32,33,34,35,36,37,38,39, the security performance of a passive IRS-assisted system is considered, and the security capacity is maximized by optimizing the reflection phase shift and/or transmission probability of the passive IRS. Few literatures consider the application of active IRS. Furthermore, existing research on active IRS aspects is limited to non-secure aspects, that is, eavesdropping users are not considered in the system model. Although active IRS greatly improves the communication quality of users, it will also leak more information to eavesdropping users due to the broadcast nature of wireless channels. Therefore, it is also necessary to consider the safety performance gain brought by active IRS.

2. System Model

In order to reduce the influence of ”double fading”, this section further considers the application of active IRS in PLS, and the considered system model is shown in Fig. 1. In the active IRS-assisted communication system model, h_D, h_E are the channels from the base station to the legitimate user and the base station to the eavesdropping user, respectively. $\mathbf{h}_{br} \in \mathbb{C}^{N\times 1}$, $\mathbf{h}_{rd} \in \mathbb{C}^{N\times 1}$ are the channels between the base station to the IRS and the IRS to the legitimate user, respectively, $\mathbf{h}_{re} \in \mathbb{C}^{N\times 1}$ is a channel from the IRS to the eavesdropping user. In addition, considering that the base station knows the CSI of all users, it is assumed that the eavesdropping user passively eavesdrops on the information transmitted by the base station to the legitimate users, without causing any interference to the legitimate users.

The reflection coefficient diagonal matrix of IRS is defined as $\Theta = diag(a_1 e^{j\theta_1}, \ldots, a_N e^{j\theta_N})$, a_n and $\theta_n \in [0, 2\pi]$ are the amplification factor matrix and phase shift matrix of IRS, respectively. In particular, the active IRS can not only change the phase shift of the channel, but also amplify the signal and then reflect it to the user[12], So the magnification factor $a_n \geq 1, \forall n$. However, the active IRS requires additional power to support the amplification of the reflected signal. In addition, active IRS amplifies the useful signal while the amplified noise signal can not be ignored.. Therefore, the received signals at the legitimate user and the eavesdropping user can be respectively expressed as[31]

$$y_s = \sqrt{p_t}\left(h_D + \mathbf{h}_{rd}^H \Theta \mathbf{h}_{br}\right) x_s + \underbrace{\mathbf{h}_{rd}^H \Theta n_r}_{\substack{\text{noise by}\\ \text{active RIS}}} + n_s \tag{1}$$

$$y_s = \sqrt{p_t}\left(h_D + \mathbf{h}_{re}^H \Theta \mathbf{h}_{br}\right) x_s + \underbrace{\mathbf{h}_{re}^H \Theta n_r}_{\substack{\text{noise by}\\ \text{active RIS}}} + n_e \tag{2}$$

where n_r is the noise introduced for active IRS amplification, and $n_r \sim \mathcal{CN}(0, \sigma_r^2)$. Then, the channel capacities of the legitimate end and the eavesdropping end are respectively defined as

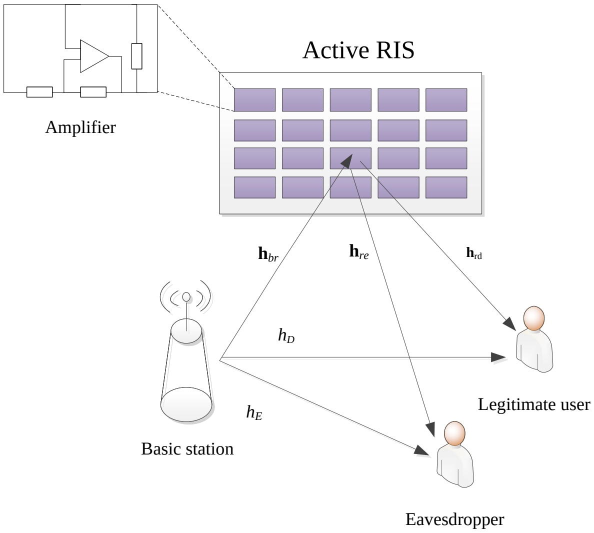

Fig. 1. Active IRS-assisted Communication system model

$$C_S = \log\left(1 + \frac{p_t\left|h_D + \mathbf{h}_{rd}^H \Theta \mathbf{h}_{br}\right|^2}{\left\|\mathbf{h}_{rd}^H \Theta\right\|^2 \sigma_r^2 + \sigma_s^2}\right) \tag{3}$$

$$C_E = \log\left(1 + \frac{p_t\left|h_E + \mathbf{h}_{re}^H \Theta \mathbf{h}_{br}\right|^2}{\left\|\mathbf{h}_{re}^H \Theta\right\|^2 \sigma_r^2 + \sigma_e^2}\right) \tag{4}$$

As mentioned above, it should be noted that, compared to passive IRS, active IRS needs to consider the amplified noise. To ensure that legitimate users can correctly decode the signal sent by the base station, the noise introduced by the active IRS should satisfy the following constraints

$$\left\|\Theta(\phi)\mathbf{h}_{br}\right\|^2 + \left\|\Theta(\phi)\mathbf{I}_N\right\|^2 \bar{\gamma}_r \leq \bar{P}_A \tag{5}$$

where $\bar{P}_A = P_A/p_t$, $\bar{\gamma}_r = \sigma_r/p_t$, P_A is the maximum noise amplification power budget for active IRS, $\mathbf{I}_N$ is the identity matrix of dimension N. In addition, the amplification gain of the active IRS also needs to satisfy the upper bound constraint. Let $a_{\max}$ be the maximum gain of amplification, assuming that the maximum gain of each reflector of each IRS is the same. Then, the reflection coefficient constraint of the IRS can be expressed as

$$|\phi_i| \leq a_{i,\max}, i = 1, \cdots, N \tag{6}$$

where $\phi = \mathrm{diag}(\Theta) = \left[a_1 e^{j\theta_1}, \cdots, a_N e^{j\theta_N}\right]^H$. According to the above constraints, it is necessary to optimize the phase shift and amplification gain to maximize the safe rate of the system. The optimization problem can be expressed as

$$\begin{aligned} \mathrm{P2}: \max_{\phi} \log & \left(1 + \frac{\left|h_D + \phi^H \mathbf{h}_d\right|^2}{\left|\phi^H \operatorname{diag}\left(\mathbf{h}_{rd}^H\right)\right|^2 \bar{\gamma}_r + \bar{\gamma}_s}\right) - \\ & \log\left(1 + \frac{\left|h_E + \phi^H \mathbf{h}_e\right|^2}{\left|\phi^H \operatorname{diag}\left(\mathbf{h}_{re}^H\right)\right|^2 \bar{\gamma}_r + \bar{\gamma}_e}\right) \end{aligned} \tag{7}$$

$$\text{s.t. } |\phi_i| \le a_{\max}, i = 1, \cdots, N \tag{8}$$

$$\left\|\phi^H \operatorname{diag}\left(\mathbf{h}_{br}\right)\right\|^2 + |\phi|^2 \bar{\gamma}_s \le \bar{P}_A \tag{9}$$

where $\bar{\gamma}_i = \sigma_i^2 / p_t$, $i \in \{r, s, e\}$, $\mathbf{h}_i = \operatorname{diag}\left(\mathbf{h}_{ri}^H\right) \mathbf{h}_{br}$, $i \in \{d, e\}$. The objective function and constraints in problem P2 are both non-convex, so they cannot be solved directly. In order to solve this problem, a joint optimization algorithm based on SDR method and min-max algorithm is proposed. The algorithm to solve this problem is introduced below.

3. Optimal Solution

Since the numerator and denominator of the objective function in problem P2 and all the constraints are related to the optimization variables ϕ, the optimization solution cannot be directly solved. First, let $\mathrm{V} = [\phi; 1]\left[\phi^H, 1\right]$, rewrite the objective function in P2. The channel capacity of the legal end can be reformulated as

$$\begin{aligned} & \log\left(1 + \tfrac{\left|h_D + \phi^H \mathbf{h}_d\right|^2}{\left|\phi^H \operatorname{diag}\left(\mathbf{h}_{rd}^H\right)\right|^2 \bar{\gamma}_r + \bar{\gamma}_d}\right) \\ & = \log\left(\left|\phi^H \operatorname{diag}\left(\mathbf{h}_{rd}\right)\right|^2 \bar{\gamma}_r + \bar{\gamma}_d + \left|h_D + \phi^H \mathbf{h}_d\right|^2\right) \\ & - \log\left(\left|\phi^H \operatorname{diag}\left(\mathbf{h}_{rd}^H\right)\right|^2 \bar{\gamma}_r + \bar{\gamma}_d\right) \\ & = \log\left(\operatorname{tr}\left(\mathrm{H}_D \, \mathrm{V}\right)\right) - \log\left(\operatorname{tr}\left(\tilde{\mathrm{H}}_D \, \mathrm{V}\right)\right) \end{aligned} \tag{10}$$

where $\mathrm{H}_D = \begin{bmatrix} \left(\operatorname{diag}\left(\mathbf{h}_{rd}\right) \operatorname{diag}\left(\mathbf{h}_{rd}\right)^H \bar{\gamma}_r + \mathbf{h}_d \mathbf{h}_d^H\right), \mathbf{h}_d h_D^H \\ h_D \mathbf{h}_d^H, \bar{\gamma}_d + h_D h_D^H \end{bmatrix}$, $\tilde{\mathrm{H}}_D = \begin{bmatrix} \operatorname{diag}\left(\mathbf{h}_{rd}\right) \operatorname{diag}\left(\mathbf{h}_{rd}\right)^H \bar{\gamma}_r, 0_{N \times 1} \\ 0_{1 \times N}, \bar{\gamma}_d \end{bmatrix}$. Similarly, the channel capacity of the eavesdropping side can be expressed as

$$\log\left(1 + \frac{\left|h_E + \phi^H \mathbf{h}_e\right|^2}{\left|\phi^H \operatorname{diag}\left(\mathbf{h}_{re}^H\right)\right|^2 \bar{\gamma}_r + \bar{\gamma}_e}\right) = \log\left(\operatorname{tr}\left(\mathrm{H}_E \, \mathrm{V}\right)\right) - \log\left(\operatorname{tr}\left(\tilde{\mathrm{H}}_E \, \mathrm{V}\right)\right) \tag{11}$$

where, $\mathrm{H}_E = \begin{bmatrix} \left(\operatorname{diag}\left(\mathbf{h}_{re}\right) \operatorname{diag}\left(\mathbf{h}_{re}\right)^H \bar{\gamma}_r + \mathbf{h}_e \mathbf{h}_e^H\right), \mathbf{h}_e h_E^H \\ h_E \mathbf{h}_e^H, \bar{\gamma}_e + h_E h_E^H \end{bmatrix}$, $\tilde{\mathrm{H}}_E = \begin{bmatrix} \operatorname{diag}\left(\mathbf{h}_{re}\right) \operatorname{diag}\left(\mathbf{h}_{re}\right)^H \bar{\gamma}_r, 0_{N \times 1} \\ 0_{1 \times N}, \bar{\gamma}_e \end{bmatrix}$. Then, let $\mathit{\Pi} =$

$\begin{bmatrix} p_t \operatorname{diag}\left(\mathbf{h}_{br}\right) \operatorname{diag}\left(\mathbf{h}_{br}\right)^H + \sigma_r^2 \mathbf{I}_N, 0_{N\times 1} \\ 0_{1\times N}, 0 \end{bmatrix}$, Constraints (9) can be expressed as

$$\mathrm{tr}(\bar{\mathit{\Pi}}\mathrm{V}) \le \bar{P}_A \tag{12}$$

Therefore, the optimization problem P2 can be reformulated as

$$\begin{aligned} \mathrm{P}2a : \max\nolimits_{\mathrm{V}} C(\mathrm{V}) = \log\left(\mathrm{tr}\left(\mathrm{H}_D\,\mathrm{V}\right)\right) - \log\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_D\,\mathrm{V}\right)\right) - \\ \log\left(\mathrm{tr}\left(\mathrm{H}_E\,\mathrm{V}\right)\right) + \log\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_E\,\mathrm{V}\right)\right) \end{aligned} \tag{13}$$

$$\text{s.t. } \mathrm{V}(i,i) \le a_{\max}, i = 1, \cdots, N \tag{14}$$

$$\mathrm{V}(N+1, N+1) = 1,\ \mathrm{V} \ge 0 \tag{15}$$

$$\mathrm{tr}(\bar{\mathit{\Pi}}\mathrm{V}) \le \bar{P}_A \tag{16}$$

After processing, it is found that the objective function of problem P2a is still nonconvex. We consider approximating the objective function, and then optimize the approximated objective function through a minimum-maximization algorithm. According to the first-order Taylor expansion, any concave function $f(\mathrm{x})$, for points within the domain $\mathrm{x}, \tilde{\mathrm{x}}$, can be written as $f(\mathrm{x}) \le f(\tilde{\mathrm{x}}) + (\nabla f(\tilde{\mathrm{x}}))^T(\mathrm{x} - \tilde{\mathrm{x}})$. Any concave function $g(X)$, for points within the domain $\mathrm{X}, \tilde{\mathrm{X}}$, can be written as $g(\mathrm{X}) \le g(\tilde{\mathrm{X}}) + \mathrm{tr}(\nabla g(\tilde{\mathrm{X}})(\mathrm{X} - \tilde{\mathrm{X}}))$.Then Since $\mathrm{tr}\left(\tilde{\mathrm{H}}_D\,\mathrm{V}\right)$ is a linear function of V, $\log_2\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_D\,\mathrm{V}\right)\right)$ is a convex function of V. Therefore, for any feasible point $\tilde{\mathrm{V}}$, it is satisfied $\log_2\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_D\,\mathrm{V}\right)\right) \le \log_2\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_D\tilde{\mathrm{V}}\right)\right) + \mathrm{tr}\left(\tilde{\mathrm{H}}_D / \mathrm{tr}\left(\tilde{\mathrm{H}}_D\tilde{\mathrm{V}}\right)(\mathrm{V} - \tilde{\mathrm{V}})\right)$. Similarly, for any feasible point , it is also satisfied $\log_2\left(\mathrm{tr}\left(\mathrm{H}_E\,\mathrm{V}\right)\right) \le \log_2\left(\mathrm{tr}\left(\mathrm{H}_E\tilde{\mathrm{V}}\right)\right) + \mathrm{tr}\left(\mathrm{H}_E / \mathrm{tr}\left(\mathrm{H}_E\tilde{\mathrm{V}}\right)(\mathrm{V} - \tilde{\mathrm{V}})\right)$. Therefore, the lower bound of the objective function of problem P2a can be expressed as

$$\begin{aligned} C(\mathrm{V}) \ge & \log\left(\mathrm{tr}\left(\mathrm{H}_D\,\mathrm{V}\right)\right) + \log\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_E\,\mathrm{V}\right)\right) - \\ & \log_2\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_D\tilde{\mathrm{V}}\right)\right) - \mathrm{tr}\left(\tilde{\mathrm{H}}_D / \mathrm{tr}\left(\tilde{\mathrm{H}}_D\tilde{\mathrm{V}}\right)(\mathrm{V} - \tilde{\mathrm{V}})\right) - \\ & \log_2\left(\mathrm{tr}\left(\mathrm{H}_E\tilde{\mathrm{V}}\right)\right) - \mathrm{tr}\left(\mathrm{H}_E / \mathrm{tr}\left(\mathrm{H}_E\tilde{\mathrm{V}}\right)(\mathrm{V} - \tilde{\mathrm{V}})\right) \end{aligned} \tag{17}$$

Then, by replacing the objective function in P2a with equation (17), and removing the constant term, the optimization problem can be approximately expressed as

$$\begin{aligned} \mathrm{P}2b : \max_{\mathrm{V}} \log\left(\mathrm{tr}\left(\mathrm{H}_D\,\mathrm{V}\right)\right) + \log\left(\mathrm{tr}\left(\tilde{\mathrm{H}}_E\,\mathrm{V}\right)\right) - \\ \mathrm{tr}\left(\tilde{\mathrm{H}}_D / \mathrm{tr}\left(\tilde{\mathrm{H}}_D\tilde{\mathrm{V}}\right)\mathrm{V}\right) - \mathrm{tr}\left(\mathrm{H}_E / \mathrm{tr}\left(\mathrm{H}_E\tilde{\mathrm{V}}\right)\mathrm{V}\right) \end{aligned} \tag{18}$$

$$\text{s.t. } \mathrm{V}(i,i) \le a_{\max}, i = 1, \cdots, N \tag{19}$$

$$\mathrm{V}(N+1, N+1) = 1,\ \mathrm{V} \ge 0 \tag{20}$$

$$\mathrm{tr}(\bar{\mathit{\Pi}}\mathrm{V}) \le \bar{P}_A \tag{21}$$

The problem P2b is already a convex problem and can be solved with the CVX toolbox. After the matrix V is obtained after optimization, if the rank of the matrix V is not

1, it can be restored to the rank 1 matrix by Gaussian randomization. Then, according to $\mathit{\Theta} = \mathrm{diag}\left(u_{\max}(\mathrm{V})\sqrt{\lambda_{\max}(\mathrm{V})}\right)[1 : N]$, the original phase shift matrix $\mathit{\Theta}$ can be obtained, where $\mu_{\max}(\bullet)$ is the eigenvector corresponding to the largest eigenvalue of the matrix, and $\lambda_{\max}(\bullet)$ is the largest eigenvalue of the matrix. The entire optimization algorithm flow is shown in the algorithm Table 1.

Table 1. Algorithms for problem P2

```
Algorithm 1 Algorithm for Problem P2
Initialize feasible point $\tilde{\mathrm{V}}$, convergent value $\varepsilon$
Calculate the value of $C(\tilde{\mathrm{V}})$,set $C(\mathrm{V}) = C(\tilde{\mathrm{V}}) + 2\varepsilon\mathbf{I}$
While $\left|C(V) - C(\tilde{V})\right| \geq \varepsilon$ :
    Solving the optimization problem P2b gets ;
    if not $rank(V) = 1$ then
        use the Gaussian randomization method;
    end if
    Calculate the value of $C(\mathrm{V})$, update $\tilde{\mathrm{V}} = \mathrm{V}$;
end while
according to $\mathit{\Theta} = \mathrm{diag}\left(u_{\max}(\mathrm{V})\sqrt{\lambda_{\max}(\mathrm{V})}\right)[1 : N]$, obtain the IRS coefficient matrix $\mathit{\Theta}$.
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4. Simulation And Result Analysis

In order to illustrate the channel gain brought by the active IRS, compared with the passive IRS, the fading model of each channel adopted in this section is the same as that in Fig. 2. Using the parameters in Fig. 2, we simulate and analyze the safety performance of the active IRS-assisted communication system. Where ”active-IRS” represents the simulation curve in the presence of active IRS, ”without-IRS” represents the simulation curve without IRS, and ”passive-IRS” represents the simulation curve in the presence of passive IRS. In Fig. 2, the transmit power of the base station is set as $p_t = 40dBm$, and we compared the curve of the relationship between the number of reflection units N and the safety rate of the IRS under different schemes. The interference power threshold of the active IRS is $\overline{P}_A = 30dBm$, and the maximum value of the amplitude amplification is set as $a_{\max}^2 = 35dB$. As the number of reflector elements increases, the security rate of both passive IRS and active IRS-aided communication systems increases. Under the same parameter settings, the performance gain brought by the active IRS is significantly better than that of the passive IRS and the communication system without IRS. As can be seen from the figure, due to the influence of ”double fading”, the number of reflection units required by the active IRS is smaller than the number of reflection units required by the passive IRS when the same safe rate is achieved. This shows that under the condition of achieving the same performance gain, the active IRS can save the number of reflection units and reduce the complexity of the IRS. When the legitimate users are far away from the IRS, that is $x_{IRS} = 40m$, the number of IRS reflection units has little effect on the system performance. The signal received by the user is mainly the signal reflected by the

IRS, and the average safe rate increases significantly with the increase of the number of reflecting units.

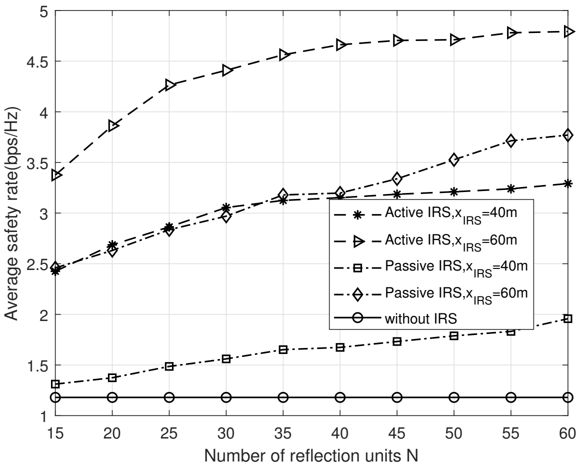

Fig. 2. Average safety rate of different reflection units

In Fig. 3, the graph is the relationship between the transmission power of different base stations and the system security rate. We set the number of reflection units $N = 32$ of IRS, and compare the improvement of system safety performance between active IRS and passive IRS under the same parameter settings. The interference power threshold of IRS is fixed as $\overline{P}_A = 20dBm$, and the maximum value of the amplitude amplification is set as $a_{\max}^2 = 35dB$. It can be seen from the figure that the larger the value of p_t, the higher the security rate of both passive IRS and active IRS-assisted communication systems. But when the power increases to a certain value, the upward trend of the safe rate curve using the active IRS scheme is gradually smoothed. This is because it is subject to the maximum interference power constraint of the IRS. Therefore, the achievable performance gain offered by active IRS is limited. Compared with communication systems without IRS, IRS can significantly improve the security performance of the system. This is because through the proposed optimization algorithm, the performance loss of the signal reflected by the IRS received by the legitimate user is smaller than that of the eavesdropping user, thereby improving the security rate of the system.

In Fig. 4, the comparative active IRS and passive IRS system safety rate curves under different parameter settings. The horizontal coordinate of the IRS is set as $x_{IRS} = 60m$, the transmit power of the base station is $p_t = 40dBm$, and the interference power threshold of the IRS is $\overline{P}_A = 20dBm$. As can be seen from the figure, the achievable performance gain provided by the active IRS is significantly better than that of the passive IRS. With the increase of $a_{\max}^2$, the security rate of the active IRS-assisted communication system is continuously improved. This shows that the active IRS can enhance the incident signal by increasing the amplification factor and the number of reflective elements.

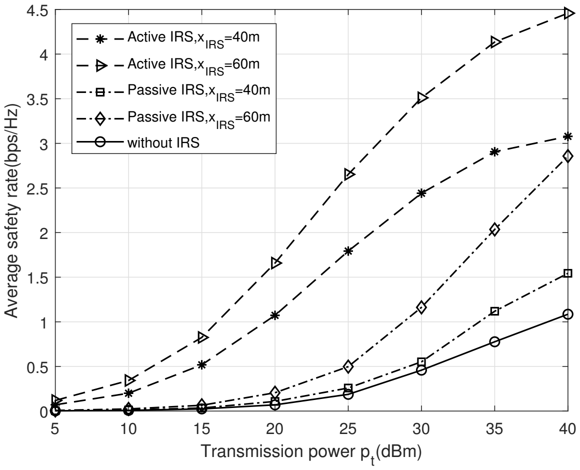

Fig. 3. Average safe rate for different transmit powers p_t

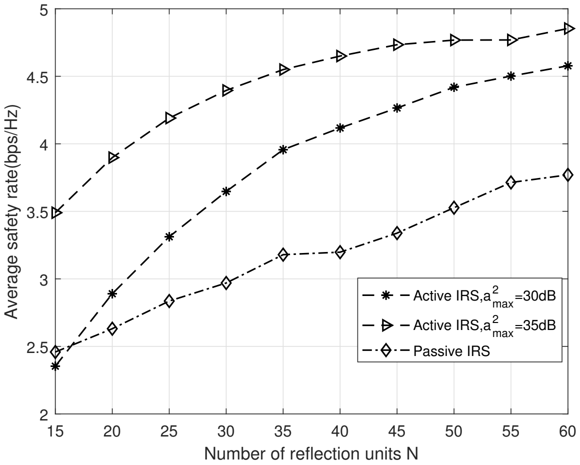

Fig. 4. Average safe rate for different amplification coefficients $a^2_{\max}$

5. Conclusion

In this chapter, we studied the IRS-assisted single-input single-output secure communication system, designed the optimal phase shift based on two types of IRS, passive and active, and proposed a secure rate maximization algorithm for IRS-assisted wireless communication networks. For the passive IRS-assisted communication system, the influence of the number of IRS reflective elements on the security performance of the system is studied. The simulation results show that the more reflective units in the IRS, the higher the performance gain brought by the IRS. However, due to the influence of ”double fading”, passive IRS improves the performance of the system to a limited extent. When the number of reflection elements increases to a certain value, the slope of the average safe rate curve also decreases continuously. For the communication system assisted by the active intelligent reflective surface, the active IRS can not only adjust the phase shift of the signal, but also amplify the amplitude of the signal. In order to solve the non-convex secrecy rate optimization problem based on this design, a min-max optimization algorithm is proposed. The simulation results show that the influence of ”double fading” effect can be effectively alleviated with the assistance of active IRS. Under the actual power consumption model, the performance gain brought by the proposed active IRS-assisted system is better than that of the traditional passive IRS system.

Acknowledgments. This work was supported by the Open Fund of Advanced Cryptography and System Security Key Laboratory of Sichuan Province under Grant SKLACSS-202115, the Projects of Hunan Provincial Department of Education Department under Grant 21A0408 and 20C0879, and the Hunan Province for Postgraduate No.QL20220248.

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Received: January 31, 2020; Accepted: July 15, 2020.