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Latency Consideration in Programmable Wireless Environment for 6G Systems

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Abstract

Intelligent reflecting surface (IRS) is a kind of cost-effective material which can optimize the beamforming of the transmission signal by adjusting the magnitude and phase shift of the signal and send to users. It is very useful in solving the latency problem facing by the 6th generation (6G) wireless communication. Latency is annoying in wireless communication because a lot of areas need low latency in communication. In this paper two systems are being built, one is system model of IRS the other is system model of IRS adding latency constraint. For the first system semidefinite relaxation (SDR) and alternating optimization are being used to maximize the power at access point (AP) and jointly design the beamforming vector of IRS, they are all of optimal solution and are all better than that without IRS. In the second system, penalty successive convex approximation (PSCA) algorithm and low-complexity algorithm are being used to maximize signal to noise ratio (SNR) at Bob, which can tell the influence of latency. Although the low-complexity algorithm do not perform well, we can also say that IRS can solve the latency problem in wireless communication better.

Acknowledgements

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Contents

1 Introduction

This chapter introduces the research background, the front edge research and unsolved problems about IRS aided wireless communication system. The aims and objectives of this paper will be illustrated in this chapter. The structure of this thesis will also be given in the following.

1.1 Overview

Wireless communication is widely used in our daily life. It is known as some equipment using together, to change the propagation direction of electromagnetic waves, in order to achieve the goals of connections between devices. The signals are send out and received through antennas and it will reflect and refract because of the media like walls. These barriers may cause attenuation depending on the characteristics of the constituent materials, and that is why sometimes, our phone will have no signals in a cave or elevator. In recent years, 5th Generation (5G) has been widely used in our daily life and has the capacity to support massive data traffic almost everywhere [2]. The capacity will reach to its potential limit in 10 years and new techniques need to be developed to support new kinds of applications such as remote surgery, virtual reality, augmented reality and so on. A lot of techniques which 5G cannot achieve will be realized based on the 6th Generation (6G) [1]. Artificial intelligence (AI) is an essential part of 6G wireless communication because of its capacity and able to carry big data and 6G will achieve the aim of collective AI. Also some areas which is not able by using 5G will emerge in 6G like wireless power transfer, radio frequency (RF) energy harvesting for massively distributed battery-less nanosensors, optical wireless communication and Li-Fi. In 6G communication, the latency will be limited to a lower value of less than 1 ms and it could also integrate with satellites which can enable global mobile coverage. In 6G the physical barriers caused by distance will be overcome, immersive extended reality, internet of intelligent things integrated ultra smart city-life and industry X.0 will be realized. In order to achieve the goal, the following key techniques will be used, pervasive AI, radar-enabled contextual communications, cell-free networks, smart metamaterials based programmable radio environment, visible light communication, etc 5G usually uses together with internet of things, massive machine type communications etc. But these techniques require a massive data rate, a smaller latency and a higher security, which 5G cannot reach the requirement right now. 6G will developing fast in recent years and in this area IRS is a very useful facility in 6G.

With the fast development of technology in material intelligent reflecting surface (IRS) comes up in this area, it is a revolutionary new technology, which can reconstruct the propagation environment intelligently and significantly improve the latency appears in the process of wireless communication network, which is strengthen the desired signal and attenuate the interference signal [3]. The cost of IRS is very low but the quality of communication can be improved a lot by using it. As is mentioned in [4], IRS can greatly improve the security during the communication process. Beamforming optimization is always a main work when designing an IRS system, the magnitude and phase shift must be jointly designed to achieve the goal [4].

According to [5] in recent years semidefinite relaxation (SDR) is widely used in area of signal processing and communication. It is a very efficient technology suitable for many optimization problems. SDR is always applied to non-convex quadratically constrained quadratic programs (QCQPs) to make it easier to solve the problem. According to [6], alternating optimization is an iterative procedure to maximize or minimize the objective function by meeting the requirements in turns until it is convergence. For penalty successive convex approximation (PSCA) algorithm, it is a kind of method used to solve complex convex optimization function, because the process is very complex, in the following, a low complexity one is being provided.

1.2 Aims and Objectives

Based on the overview about wireless communication and IRS, the aim of this project is to find the optimal beamforming of IRS-aided 6G wireless communication with less transmit power consumpted and improve the performance of the IRS-aided wireless communication system after adding latency constraint. To solve the first problem, two methods are being used, one is SDR, the other is alternating optimization. To solve the second problem two methods are being used, one is penalty successive convex approximation, the other is a lower complexity one. The specific objectives of this paper are as follows:

- Two kinds of methods, SDR and alternating optimization, are studied two solve the problem of the IRS aided wireless communication. By using these two methods, the problem is formulated.

- The result with different horizontal distances and number of reflecting surface is provided, with comparison we can get the performance of these two kinds of methods and how the performance varies with different horizontal distances and number of reflecting surface.

- Two kinds of methods, PSCA and low-complexity algorithm, are studied two solve the problem of the system adding latency constraint. By using these two methods, the problem is formulated.

- The result with different IRS horizontal location and number of reflecting surface is provided, with comparison we can get the performance of these two kinds of methods and how the performance varies with different IRS horizontal location and number of reflecting surface.

1.3 Thesis Structure

The thesis structure is as follows:

- In Chapter 2, we will introduce the research background which is mainly about latency and IRS;

- In Chapter 3, a wireless communication system model with IRS is introduced and the corresponding methods to formulate the problem. In this chapter, two kinds of methods are used to solve the problem.

- In Chapter 4, latency constraints will be added to the IRS-aided wireless communication model and its corresponding problem formulation will be provided. In this chapter, two kinds of methods are used to solve the problem.

- In Chapter 5, the simulation results are illustrated, both system model will be simulated in this chapter.

- In Chapter 6, the results of this paper will be summarized conclusions will be given. The direction of future will also be discussed.

1.4 Notations

In this paper, italic letters x donate scalars, bold lower-case $\mathbf{x}$ and upper-case letters $\mathbf{X}$ donate vectors and matrices respectively, $\mathbb{C}^{x \times y}$ represents a $x \times y$ complex-valued matrix. For a complex-valued vector $\mathbf{x}$, $||\mathbf{x}||$ means its Euclidean norm, arg$(\mathbf{x})$ represents a vector with each elements means the angle of the vector $\mathbf{x}$, diag$(\mathbf{x})$ represents a diagonal matrix of vector $\mathbf{x}$

2 Background

In this chapter, the concept of latency in wireless communication and some corresponding technologies to improve the latency performance are briefly introduced; among them, a new technology called IRS will be introduced.

2.1 Latency in Wireless Communication

In recent years, 5G will reach to its limitation and 6G is in a fast development speed. The aim of 6G is high security and low latency. In the following I will introduce the latency in wireless communication and the latest work about latency.

Latency is caused by the time difference of reflection, for example, when some one send signal out, there are two ways for the signal reach to the user, one is directly from transmitter to user, the other is after one or some times of reflection and then transmit to user. The overlap of these two kinds of signals will cause noise to the receiver, for the signal at the user at a wrong time is useless and will disturb the communication.

Delay is very annoying in wireless communication, wireless communication in a lot of conditions needs low latency, like remote surgery and networked video game. For the remote surgery, it requires in time response to the patient's condition, in 2019 a doctor in Fujian finished a remote surgery on animal by using 5G, but nowadays we cannot perform this on human and 6G will solve this problem. For the video game played on the internet, in order to have a good time, low latency for internet must be required, this will do good to both the player and its partner.

Recent studies on delay in wireless communication appears in a lot of areas. In [7] it mentions that for Mobile Edge Computing (MEC) system non-orthogonal multiple-access (NOMA) and time-division multiple-access (TDMA) are used to solve the problem of delay and the latter one has a better performance. In [8], it mentions that by using multidomain orchestration 5G network can configure for tactile internet services, but the latency still need to be improved for a better user experience. In [9] for vehicle-to-vehicle (V2V) communications, it performs well when balancing the current single-hop latency and the residual multi-hop latency. In [11] a delay-intolerant communication system is built it mentions that when the allowable delay is in a short span or the covert requirement is very strict, randomly changing the transmit power can enhance the communication.

2.2 Introduction to IRS

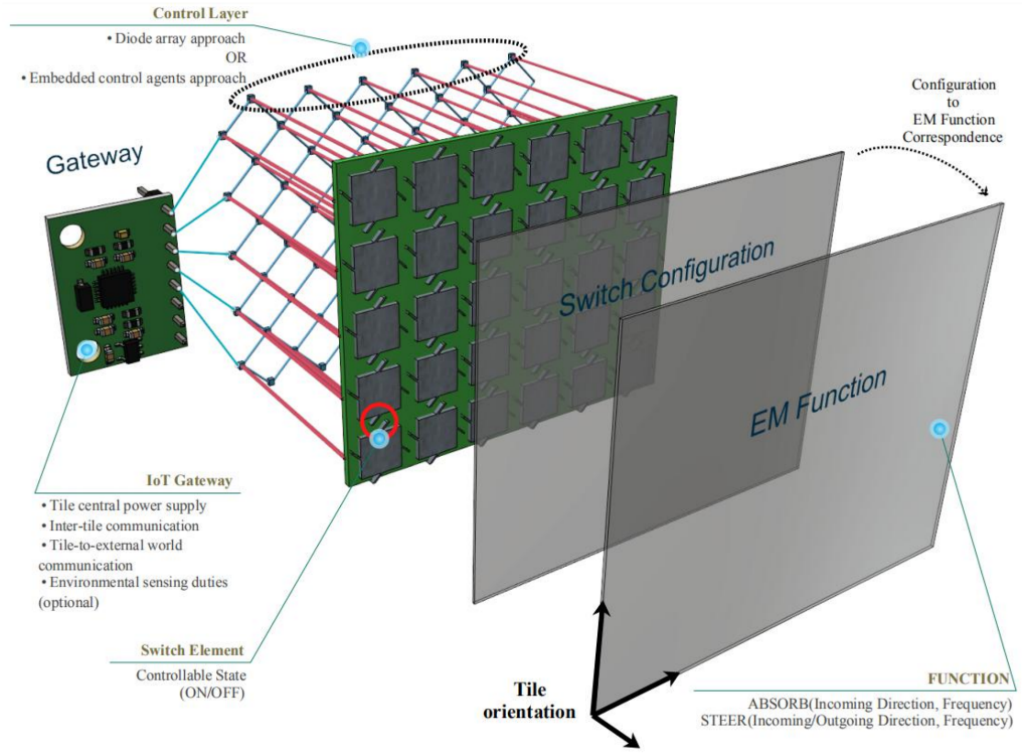


Figure 1: Working procedure of IRS [10]

IRS is a revolutionary new technology, which can intelligently reconstruct the infinite propagation environment and significantly improve the security performance of wireless communication network. The structure of IRS is in figure 1 is a planar array composed of a large number of auxiliary components, such as low-cost printed dipole. Each component can independently induce a certain phase shift on the incident electromagnetic wave through an intelligent controller.The main material of it is metasurface and hypersurface.

Metasurface is a main technology in programmable wireless communication, it has a silicon substrate and over it there are some atoms. The structure of it is planar and artificial. The meta-atom is a kind of conductor with one-tenth of the incident wave whose form can determine the EM behavior. The metasurface is able to control EM waves in a corresponding frequency. It has two kinds, static which comprise split ring resonators and dynamic which can change the meta-atom structure. It has functions include redirection, beam splitting, wave absorbing, wave polarizing, wave front focus and phase control. [10]

Hypersurfaces which can be controlled by software interface are a new metasurfaces, in the control the structure and electromagnetic behavior can be changed. The hardware includes a dynamic metasurface, a set of networked, miniaturized controllers which can monitor and modify the state of at least one metasurface switch element, switching elements that can monitor and modify the metasurface, and a gateway that provides connections between internal chips and external environment, it can also exchange data with other system. For the tile inter-networking, the internet of things communication protocols is being used which can accumulate data and diffuses EM actuation commands. For the environment control software, it is an application programming interface. [10]

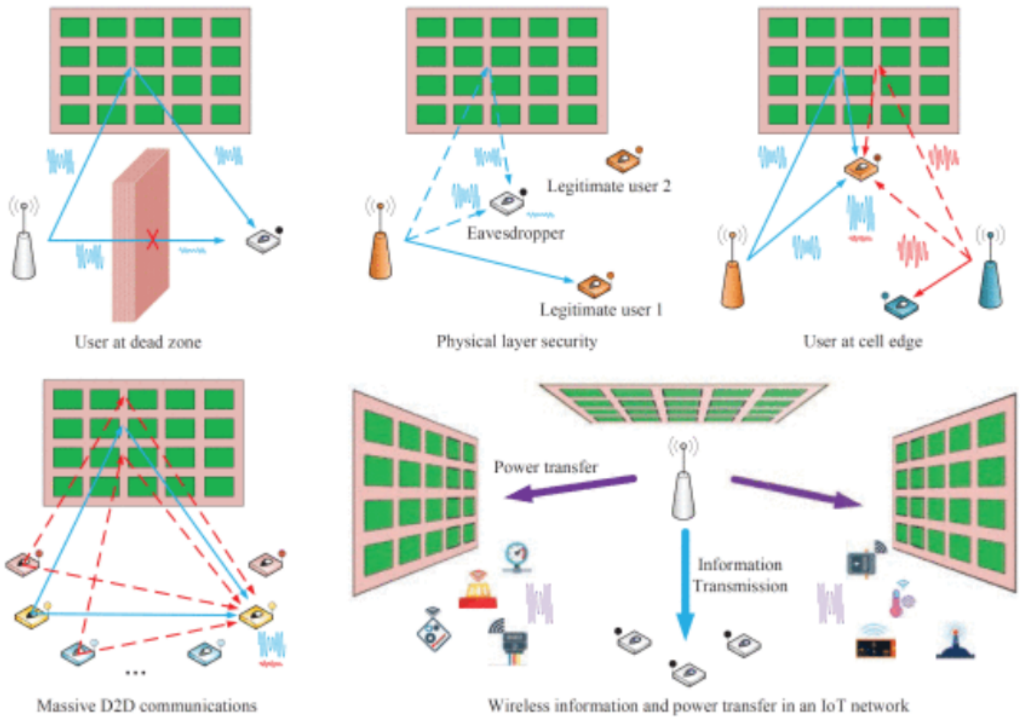


Figure 2: Using Condition of IRS [24]

Figure 2 shows the conditions where IRS can be used: [24]

- At a corner where the user is located, IRS can help creating a reflection path properly, so that the user can still receive the signal.
- Assuming that the eavesdropper is in the same line with the base station and the user, no matter how to design the beamforming of the base station, the eavesdropper will always get a information from transmitter. Therefore, appropriate design of the reflection factor of the IRS can largely weaken the information get by eavesdropper.The signal can be enhanced by superposition of the reflecting signals.
- In the case when the user is at cell edge, signal of the base station in the cell will seriously attenuate and meanwhile disturb by the adjacent ones. By using IRS the signal of this cell will be enhanced and the disturbance to others will be reduced.

- In the scenario of device to device (D2D), IRS can be used to reduce interference between them and enhance the needed signal.
- For transmitting signal and power through long distance, the efficient of transmission will be better.

Comparing with other methods, IRS has a low cost, it is green with no pollution to the environment. IRS is also to be operated. When detecting where user is, the IRS can jointly design the magnitude and the phase shift of the signal to let it work. It can also increase the security and decrease the latency influence during the communication.

3 System Model of IRS and Problem Formulation

This chapter introduces the IRS-aided wireless communication system model and the corresponding problem formulation. The system problem can be solved by two methods, one is SDR, the other is alternating optimization.

3.1 System Model

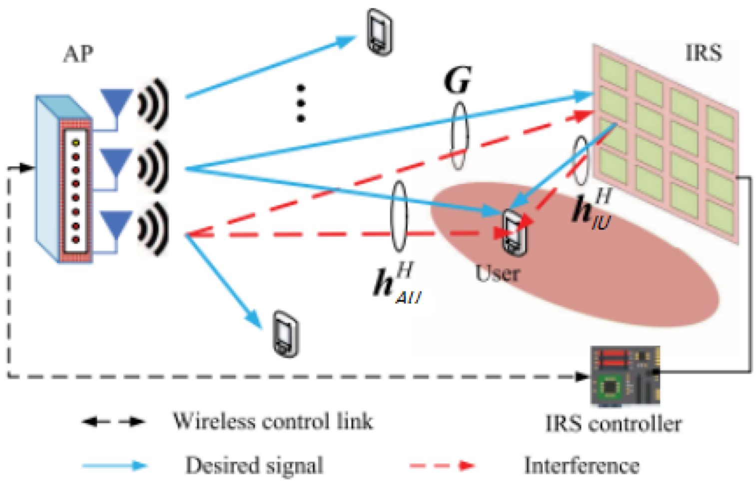


Figure 3: IRS system model [12]

Figure 3 shows the structure of the IRS-aided wireless communication. In order to simplify this problem, only a single user will be considered in this part.

As it can bee seen from the figure that the structure consist of three parts, the access point (AP), a user and an IRS which is controlled by the controller. There are two ways for a signal to transmit from the AP to the users, one is direct transform without any reflection, the other is aided with IRS which can control the magnitude and phase shift of the signal to be the optimal choice. In this article, the one aided with the IRS will be considered and in order to simplify this problem, only a single user will be considered in this part. Which means there is only one antenna at the user.

The base band equivalent channels from the AP to IRS, from the IRS to user and from the AP to user are defined as $\mathbf{G} \in \mathbb{C}^{N\times M}$, $\mathbf{h}^H_{IU} \in \mathbb{C}^{1\times N}$, $\mathbf{h}^H_{AU} \in \mathbb{C}^{1\times M}$, respectively. During the communication, each element of the IRS will receive the superposed multi-path signal from the AP and adjust the phase of the received signal and transmit to user. Suppose the reflection angle of the IRS is defined as $\theta = [\theta_1, \theta_2, ..., \theta_N]$ where $\theta_n \in [0, 2\pi]$ donate the angle of the n^{th} element. Then define a diagonal matrix $\Theta = \text{diag}(\beta_1 e^{j\theta_1}, \cdots, \beta_N e^{j\theta_N})$ where $\beta_N \in [0, 1]$

which means the n^{th} amplitude of the reflection coefficient. In the following article, in order to simplify calculation we use $\beta_n = 1$ and the matrix Θ is simplified to $\Theta = \text{diag}(e^{j\theta_1}, \cdots, e^{j\theta_N})$

As the user will both receive signal directly from the AP and from the IRS, the signal received at user is [12]

$$y = (\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H)\mathbf{w}s + n \tag{1}$$

where s means the transmitted data for the user and $w \in \mathbb{C}^{M\times 1}$ means the corresponding beamforming vector, $N \sim CN(0, \sigma)$ denotes the additive white Gaussian noise (AWGN) at the user. Then we can defined the corresponding signal-to-interference-plus-noise ratio (SINR) which is [12]

$$SINR = \frac{|(\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H)\mathbf{w}|^2}{\sigma^2}, \forall k \tag{2}$$

3.2 Problem Formulation

The aim of this project is to optimize the transmit beamforming at the AP and reflection beamforming at the IRS, subject to SINR constraints at user. The problem can be formulated as [12]

$$\text{(P1):} \quad \min_{\mathbf{w},\theta} \ ||\mathbf{w}||^2 \tag{3a}$$

$$\text{s.t.} \ |(\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H)\mathbf{w}|^2 \geq \gamma\sigma^2 \tag{3b}$$

$$0 \leq \theta_n \leq 2\pi, n = 1, ..., N \tag{3c}$$

where γ is the minimum SINR requirement of the user.

This is a non-convex optimization problem because of the left-hand-side of (3c) is not jointly concave with respect to **w** and θ. We will use two methods to solve the optimization problem.

3.2.1 SDR

Applying SDR to problem (P1). The lower bound of the problem is given by evaluating the performance gaps from other suboptimal solutions. For random phase shift θ, the optimal transmit beamforming can be expressed as $\mathbf{w}^* = \sqrt{P}\frac{(\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H)^H}{|\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H|}$, which is called the maximum-ratio transmission (MRT) [12], where P is the transmit power at AP. By using $\mathbf{w}^*$, the form of problem (P1) can be change into [12]

$$\min_{P,\theta} \ P \tag{4a}$$

$$\text{s.t.} \ \ P||\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H||^2 \geq \gamma\sigma^2 \tag{4b}$$

$$0 \leq \theta_n \leq 2\pi, \forall n \tag{4c}$$

From this the optimal solution for this problem can be given by $P^* = \frac{\gamma\sigma^2}{||\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H||^2}$. As the minimum value of the power is the subject, the maximum value of the denominator need to be find. Then the problem change into [12]

$$\max_{\theta} \ ||\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H||^2 \tag{5a}$$

$$\text{s.t.} \ \ 0 \leq \theta_n \leq 2\pi, \forall n \tag{5b}$$

Suppose $\mathbf{v} = [v_1, \cdots, v_N]^H$, according to (5b) we can find that $|v_n|^2 = 1, \forall n$. Suppose $\mathbf{h}_{IU}^H \Theta G = \mathbf{v}^H \Phi$, where $\Phi = \text{diag}(\mathbf{h}_{IU}^H)\mathbf{G}$, then problem (5a) can be changed into [12]

$$\max_{\mathbf{v}} \ \mathbf{v}^H \Phi \Phi^H \mathbf{v} + \mathbf{v}^H \Phi \mathbf{h}_{AU} + \mathbf{h}_{AU}^H \Phi^H \mathbf{v} + ||\mathbf{h}_{AU}^H||^2 \tag{6a}$$

$$\text{s.t.} \ \ |v_n|^2 = 1, \forall n \tag{6b}$$

As is mention in [13] that (6a) is a non-convex quadratically constrained quadratic program (QCQP) which can be reformulated as a homogeneous QCQP by introducing an auxiliary variable t, (6b) is changed into [12]

$$\max_{\mathbf{v}} \ \bar{\mathbf{v}}^H \mathbf{R} \bar{\mathbf{v}} + ||\mathbf{h}_{AU}^H||^2 \tag{7a}$$

$$\text{s.t.} \ \ |v_n|^2 = 1, n = 1, ..., N+1 \tag{7b}$$

where $\mathbf{R} = \begin{bmatrix} \Phi\Phi^H & \Phi\mathbf{h}_{AU} \\ \mathbf{h}_{AU}^H \Phi^H & 0 \end{bmatrix}, \bar{\mathbf{v}} = \begin{bmatrix} \mathbf{v} \\ t \end{bmatrix}.$

If we want to use CVX in MATLAB to solve this problem, the object function must be convex. But as is mentioned in [13] that (7a) is not a convex function so we use SDR to relax the constraint. (7a) is reduced into [12]

$$\max_{\mathbf{V}} \ \mathrm{tr}(\mathbf{RV}) + ||\mathbf{h}_{AU}^H||^2 \tag{8a}$$

$$\text{s.t.} \ \mathbf{V}_{n,n} = 1, n = 1, ..., N+1 \tag{8b}$$

$$\mathbf{V} \geq 0 \tag{8c}$$

As we can see (8a) is a convex function and by using CVX [14] in MATLAB, we can solve the SDR problem.

3.2.2 Alternating Optimization

Besides SDR, we also have a method to solve the optimal problem. Which is alternating optimization.

Let $\mathbf{w} = \sqrt{P}\bar{\mathbf{w}}$ where $\bar{\mathbf{w}}$ donates the direction of the beamforming vector.

To start with we use fixed $\bar{\mathbf{w}}$ to optimize the reflection angle of the IRS θ.

By using $\mathbf{w} = \sqrt{P}\bar{\mathbf{w}}$ problem (P1) can be written into [12]

$$\max_{\theta} \ |(\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H)\bar{\mathbf{w}}|^2 \tag{9a}$$

$$\text{s.t.} \ 0 \leq \theta_n \leq 2\pi, n = 1, ..., N \tag{9b}$$

$$\tag{9c}$$

An inequality according to triangle inequality can be applied to solve (9a) and reduce (9a) into [12]

$$\begin{aligned} |(\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H)\bar{\mathbf{w}}| &= |\mathbf{h}_{IU}^H \Theta \mathbf{G}\bar{\mathbf{w}} + \mathbf{h}_{AU}^H \bar{\mathbf{w}}| \\ &= |\mathbf{h}_{IU}^H \Theta \mathbf{G}\bar{\mathbf{w}}| + |\mathbf{h}_{AU}^H \bar{\mathbf{w}}| \end{aligned} \tag{10}$$

This holds when $\arg(\mathbf{h}_{IU}^H \Theta \mathbf{G}\bar{\mathbf{w}}) = \arg(\mathbf{h}_{AU}^H \bar{\mathbf{w}}) \equiv \phi_0$.

Then let $\mathbf{h}_{IU}^H \Theta \mathbf{G}\mathbf{w} = \mathbf{v}^H a$ where $\mathbf{v} = [e^{j\theta_1}, ..., e^{j\theta_N}]^H$ and $\mathbf{a} = \mathrm{diag}(\mathbf{h}_{IU}^H \mathbf{G}\bar{\mathbf{w}})$.

Then (9a) is equivalent to [12]

$$
\begin{aligned}
\max_{\mathbf{v}} \quad & |\mathbf{v}^H \mathbf{a}|^2 && \text{(11a)} \\
\text{s.t.} \quad & |v_n| = 1.n = 1, ..., N && \text{(11b)} \\
& \arg(\mathbf{v}^H \mathbf{a}) = \phi_0 && \text{(11c)}
\end{aligned}
$$

From this the optimal solution of θ_n can be solved by given [12] $\mathbf{v}^* = e^{j(\phi_0 - \arg(\mathrm{diag}(\mathbf{h}_r^H)\mathbf{G}\bar{\mathbf{w}}))}$ and

$$
\theta_n^* = \phi_0 - \arg(h_{n,IU}^H) - \arg(\mathbf{g}_n^H \bar{\mathbf{w}}) \tag{12}
$$

where $h_{n,IU}^H$ is the n^{th} element of $\mathbf{h}_r^H$ and $\mathbf{g}_n^H$ is the nth row of $\mathbf{G}$.

Then we use fixed reflection angle of the IRS θ to optimize $\bar{\mathbf{w}}$. We can find that the solution is [12]

$$
\bar{\mathbf{w}}^* = \frac{(\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H)^H}{||\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H||^2} \tag{13}
$$

and the optimal solution to the power is [12]

$$
P^* = \frac{\gamma \sigma^2}{||\mathbf{h}_{IU}^H \Theta \mathbf{G} + \mathbf{h}_{AU}^H||^2} \tag{14}
$$

4 System Model Adding Latency Constraints and Problem Formulation

This chapter introduces the IRS-aided wireless communication system adding latency constraints and its problem formulation. This system is a little bit different from the former one by adding a warden Willie. The system can be solved by two kinds of methods, one is PSCA algorithm, the other is low-complexity algorithm.

4.1 System Model Adding Latency Constraints

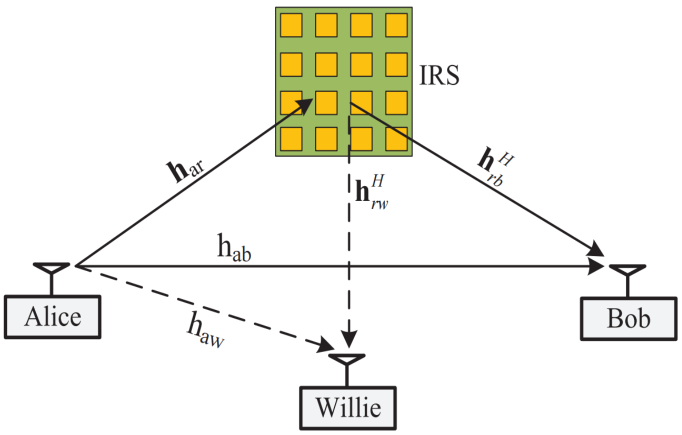


Figure 4: System Model Adding Latency Constraints [16]

Figure 4 shows the structure of the IRS-aided wireless communication system adding latency constraints. In order to simplify this system every antenna is considered to be single. Because of large path loss the signal reflected by the IRS twice or more are being ignored [17].

As it can bee seen from the figure that the structure consist of four parts, Alice is a signal transmitter at a unit time it only send a unit signal, it transmit signal to the signal receiver Bob. There is also an IRS used to control the magnitude and phase shift of the signal from Alice and reflect to Bob. In this system, Willie is a warden which is used to detect the existence of this transmission. The IRS is equipped with N passive reflecting elements which can dynamically adjust the direction and the power of the signal. The channel from Alice to IRS, Bob and Willie can be expressed as $\mathbf{h}_{ar} = \sqrt{\chi_{ar}}\overline{\mathbf{h}}_{ar}$, $h_{ab} = \sqrt{\chi_{ab}}\overline{h}_{ab}$, $h_{aw} = \sqrt{\chi_{aw}}\overline{h}_{aw}$, respectively. The channel from IRS to Bob and Willie are $\mathbf{h}_{rb}^H = \sqrt{\chi_{rb}}\overline{\mathbf{h}}_{rb}^H$, $\mathbf{h}_{rw}^H = \sqrt{\chi_{rw}}\overline{\mathbf{h}}_{rw}^H$ respectively. In those formulas, χ_{ij} means the path loss from node i to node j, where $ij \in \{ar, ab, aw, rb, rw\}$ donates different channels, and $\overline{\mathbf{h}}_{ar}$, $\overline{h}_{ab}$, $\overline{h}_{aw}$, $\overline{\mathbf{h}}_{rb}^H$, $\overline{\mathbf{h}}_{rw}^H$ means the small-scale fading coefficients. The diagonal reflecting matrix of IRS is $\Theta = \text{diag}(\beta_1 e^{j\theta_1}, ..., \beta_N e^{j\theta_N})$ where $\beta_N \in [0,1]$ means the nth amplitude reflection coefficient and $\theta_n \in [0, 2\pi)$ means the phase shift of the nth element. Suppose the signal transmitted by Alice in the ith channel is $x[i] \sim CN(0,1), \forall i \in \{1, 2, ..., L\}$, where L is the number of channel uses.

4.1.1 Binary Hypothesis Testing at Willie

As is mentioned in the former section that Willie works as a warden, which can differentiate these two hypotheses [16]

$$y_w[i] = \begin{cases} n_w[i], & H_0, \\ \sqrt{P_a}(\mathbf{h}_{rw}^H \Theta \mathbf{h}_{ar} + h_{aw})x[i] + n_w[i], & H_1, \end{cases} \tag{15}$$

where $y_w[i]$ is the ith channel received signal at Willie, H_0 means the null hypothesis that Alice does not transmit, and H_1 is a alternative hypothesis which means Alice transmit information to Bob. P_a means the transmit power at Alice and $n_w[i]$ is the AWGN at Willie whose mean is zero and with σ_w^2 variance. At Willie the false alarm rate is $Pr\{D_1|H_0\}$ and the miss detection rate is $Pr\{D_1|H_0\}$, where D_0 and D_1 are used to express the binary meaning of whether transmission at Alice occurred or not. According to this, the total error rate at Willie can be expressed as [16]

$$\xi = Pr\{D_1|H_0\} + Pr\{D_0|H_1\} \tag{16}$$

During the transmission performance, the total error rate should be minimized in order to have a better performance. The likelihood ratio test optimal test that minimizes ξ is [16]

$$\frac{\mathbb{P}_1 \triangleq \prod_{i=1}^{L} f(y_w[i]|H_1)}{\mathbb{P}_0 \triangleq \prod_{i=1}^{L} f(y_w[i]|H_0)} \underset{D_0}{\overset{D_1}{\gtrless}} 1 \tag{17}$$

where $\mathbb{P}_0$ and $\mathbb{P}_1$ are the likelihood functions of Willie's observation vector over L independent channel uses under H_0 and H_1, respectively. The likelihood function of $y_w[i]$ under H_0 and H_1 can be express as $f(y_w[i]|H_0) = CN(0, \sigma_{w^2})$ and $f(y_w[i]|H_1) = CN(0, P_a|\mathbf{h}_{rw}^H \Theta \mathbf{h}_{ar} + h_{aw}|^2 + \sigma_w^2)$, respectively.

According to (17) the minimum detection error rate ξ^* at Willie can be obtained [11]. But the expression of ξ^* has gamma function which is incomplete. It is not easy to deal with the subsequent analysis and design. In order to solve this problem, we put forward a lower bound on ξ^*, which is [18]:

$$\xi^* \geq 1 - \sqrt{\frac{1}{2} D(\mathbb{P}_0|\mathbb{P}_1)} \tag{18}$$

where $D(\mathbb{P}_0|\mathbb{P}_1)$ is the KL divergence from $\mathbb{P}_0$ to $\mathbb{P}_1$ and according to [11]:

$$D(\mathbb{P}_0|\mathbb{P}_1) = L\left[\ln\left(1 + \frac{P_a|\mathbf{h}_{rw}^H \Theta \mathbf{h}_{ar} + h_{aw}|^2}{\sigma_w^2}\right) - \frac{P_a|\mathbf{h}_{rw}^H \Theta \mathbf{h}_{ar} + h_{aw}|^2}{P_a|\mathbf{h}_{rw}^H \Theta \mathbf{h}_{ar} + h_{aw}|^2 + \sigma_w^2}\right] \tag{19}$$

suppose $\mathbf{v} = [v_1, v_2, \cdots, v_N]^T$ where $v_n = \rho_n e^{-j\theta_n}$, $\forall n$ and $\mathbf{a} = \text{diag}(\mathbf{h}_{rw}^H)\mathbf{h}_{ar}$, it is known that $\mathbf{h}_{rw}^H \Theta \mathbf{h}_{ar} = \mathbf{v}^H \text{diag}(\mathbf{h}_{rw}^H)\mathbf{h}_{ar}$, then $|\mathbf{h}_{rw}^H \Theta \mathbf{h}_{ar} + h_{aw}|^2$ can be expressed as $|\mathbf{v}^H \mathbf{a} + h_{aw}|^2$.

In this communication system, $\xi^* \geq 1 - \epsilon$ is usually used as a covertness constraint, where ϵ, a small value, means the covertness level being required. According to per (18), $D(\mathbb{P}_0|\mathbb{P}_1) \leq 2\epsilon^2$ is more stringent than constraint $\xi^* \geq 1 - \epsilon$. Because of this, $D(\mathbb{P}_0|\mathbb{P}_1) \leq 2\epsilon^2$ can be used as the covertness constraint.

4.1.2 Transmission from Alice to Bob

The signal received by Bob at the ith channel transmitted by Alice is [16]

$$y_b[i] = \sqrt{P_a}(\mathbf{h}_{rb}^H \Theta \mathbf{h}_{ar} + h_{ab})x[i] + n_b[i] \tag{20}$$

where $n_b[i]$ is the AWGN at Bob with zero mean and σ_b^2 variance.Signal-to-noise-ratio (SNR) at Bob can be expressed as [16]

$$\gamma_b = \frac{P_a}{\sigma_b^2}|\mathbf{h}_{rb}^H \Theta \mathbf{h}_{ar} + h_{ab}|^2 = \frac{P_a}{\sigma_b^2}|\mathbf{v}^H \mathbf{b} + h_{ab}|^2 \tag{21}$$

where $\mathbf{b} = \text{diag}(\mathbf{h}_{rb}^H)\mathbf{h}_{ar}$

4.2 Problem Formulation

Maximize SNR at Bob is the goal of the work, to achieve this the transmit power at Alice and the phase shift and reflection amplitude of the reflecting beamforming vector **v** at IRS should be jointly designed. They are constrained by covertness level, maximum transmit power and IRS reflection coefficient at Alice. The formulated optimization problem can be expressed as [16]

$$
\begin{aligned}
\text{(P2):} \quad & \max_{P_a,\mathbf{v}} \; P_a|\mathbf{v}^H\mathbf{b}+h_{ab}|^2 && \text{(22a)}\\
& \text{s.t.} \; D(\mathbb{P}_0|\mathbb{P}_1) \le 2\epsilon^2 && \text{(22b)}\\
& P_a \le P_{max} && \text{(22c)}\\
& |v_n| \le 1, \forall n=1,2,...,N && \text{(22d)}
\end{aligned}
$$

where σ_b^2 in the objective function (22a) can be ellipsised, because it is constant. (22b) is covertness and (22d) can be get from $\rho_n \in [0,1]$ and $\theta \in [0,2\pi)$, for IRS cannot amplify signals, it can only adjust the amplitude and phase.

The design of reflection amplitude and phase shift of IRS is combined in (22). The phase shift of IRS is designed in particular. Based on (22) $\rho_n = 1$ is optimal or not can be revealed. Considering the limited block length, because Willie mixes signals from both direct and reflected path in the system, IRS aided wireless communication is very different from that without IRS. Willie can detect the beamforming from Alice to Bob when the design is not suitable, even though the quality of communication is improved. First, we solve the problem of non covert transmission with $D(\mathbb{P}_0|\mathbb{P}_1)=0$. We notice that perfect concealment is impossible in a covert communication system without IRS when the block length is limited [19].

***Theorem** 1*: Only with non zero transmit power perfect concealment can be achieved. The below problem P1' [16]

$$
\begin{aligned}
\text{(P2'):} \quad & \max_{P_a,\mathbf{v}} \; P_a|\mathbf{v}^H\mathbf{b}+h_{ab}|^2 && \text{(23a)}\\
& \text{s.t.} \; P_a|\mathbf{v}^H\mathbf{a}+h_{aw}|^2 = 0 && \text{(23b)}\\
& P_a \le P_{max} && \text{(23c)}\\
& |v_n| \le 1, \forall n=1,2,...,N && \text{(23d)}
\end{aligned}
$$

can work only if $\sum_{n=1}^{N} |a_n| \le |h_{aw}|$, where a_n is the nth element of $\mathbf{a} = \text{diag}(\mathbf{h}_{rw}^H)\mathbf{h}_{ar}$.

The equivalent channel coefficient of reflect-path from Alice to Willie through the IRS can be expressed as $\mathbf{v}^H\mathbf{a}$. As $\sum_{n=1}^{N} |a_n| = ||\mathbf{a}||_1$ can be written as $||\mathbf{a}||_1 = |\mathbf{v}^H\mathbf{a}|_1$ when $\rho_n = 1, \forall n$. Therefore, **Theorem 1** means that perfect concealment can be achieved when channel quality of the reflected path is higher than the direct one. What's more, Theorem 1 also reveals that when the reflection beamforming is designed properly, the additional reflection path will reduce Willie's detection performance. This essentially shows that IRS can effectively improve the communication concealment, which will be clearly tested by solving the optimization problem (P2).

As problem P2' is non-convex, it need to be converted into a convex function. In the following, a joint design based on penalty successive convex approximation (PSCA) algorithm will be used to solve the problem.

4.2.1 PSCA

The term $P_a|\mathbf{v}^H\mathbf{b}+h_{ab}|^2$ in (22a) and the term $P_a|\mathbf{v}^H\mathbf{a}+h_{aw}|^2$ in (22b)can be written as [16]

$$P_a(\mathbf{v}^H\mathbf{b}\mathbf{b}^H\mathbf{v}+2\mathrm{Re}(\mathbf{v}^H\mathbf{b}h_{ab}^*t)+|h_{ab}|^2|t|^2) \tag{24}$$

$$P_a(\mathbf{v}^H\mathbf{a}\mathbf{a}^H\mathbf{v}+2\mathrm{Re}(\mathbf{v}^H\mathbf{a}h_{aw}^*t)+|h_{aw}|^2|t|^2) \tag{25}$$

respectively, where t satisfies $|t|^2 = 1$. The above two equations (24) and (25) can be expressed as quadratic forms $P_a\mathbf{u}^H\mathbf{B}\mathbf{u}$ and $P_a\mathbf{u}^H\mathbf{A}\mathbf{u}$, respectively, where [16]

$$\mathbf{u}=\begin{bmatrix}\mathbf{v}\\ t\end{bmatrix},\mathbf{B}=\begin{bmatrix}\mathbf{b}\mathbf{b}^H & \mathbf{b}h_{ab}^*\\ h_{ab}\mathbf{b}^H & |h_{ab}|^2\end{bmatrix},\mathbf{A}=\begin{bmatrix}\mathbf{a}\mathbf{a}^H & \mathbf{a}h_{aw}^*\\ h_{aw}\mathbf{a}^H & |h_{aw}|^2\end{bmatrix} \tag{26}$$

According to this formula, problem (P2) can be written as [16]

$$\begin{aligned}
\text{(P2.1):}\quad & \max_{P_a,\mathbf{u}}\ P_a\mathbf{u}^H\mathbf{B}\mathbf{u} && \text{(27a)}\\
& \text{s.t.}\ P_a\le P_{max} && \text{(27b)}\\
& \ln\left(1+\tfrac{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}}{\sigma_w^2}\right)-\frac{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}}{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}+\sigma_w^2}\le\frac{2\epsilon^2}{L} && \text{(27c)}\\
& |v_n|\le 1,\forall n=1,2,...,N && \text{(27d)}\\
& |v_{N+1}|=1 && \text{(27e)}
\end{aligned}$$

As $\mathbf{u}$ is an optimal solution to problem (P2.1) and $\frac{\mathbf{v}}{t}$ is an optimal solution to problem (P2) [20], it is still difficult to solve the problem because there are two variables P_a and beamforming vector **u** both appear in (27a). As P_a is a scalar, problem (P2.1) can be simplify into [16]

$$
\begin{aligned}
\text{(P2.2):}\quad & \max_{P_a,\mathbf{w}} \ \mathbf{w}^H\mathbf{B}\mathbf{w} && \text{(28a)}\\
& \text{s.t.}\ \ln\left(\tfrac{\mathbf{w}^H\mathbf{A}\mathbf{w}+\sigma_w^2}{\sigma_w^2}\right) - \frac{\mathbf{w}^H\mathbf{A}\mathbf{w}}{\mathbf{w}^H\mathbf{A}\mathbf{w}+\sigma_w^2} \le \frac{2\epsilon^2}{L}, && \text{(28b)}\\
& P_a \le P_{max}, && \text{(28c)}\\
& |w_n| \le \sqrt{P_a}, \forall n = 1,2,...,N, && \text{(28d)}\\
& |w_{N+1}| = \sqrt{P_a}, && \text{(28e)}
\end{aligned}
$$

where $\mathbf{w} = \sqrt{P_a}\mathbf{u}$. As constraints (28b) and (28e) is non-convex constraints, problem (P2.2) is hard to solve. In order to solve this problem, we define $\mathbf{W} = \mathbf{w}\mathbf{w}^H$. The problem (P2.2) can be simplified as [16]

$$
\begin{aligned}
\text{(P2.3):}\quad & \max_{P_a,\mathbf{w}} \ \mathrm{Tr}(\mathbf{B}\mathbf{W}) && \text{(29a)}\\
& \text{s.t.}\ \ln\left(1+\tfrac{\mathrm{Tr}(\mathbf{A}\mathbf{W})}{\sigma_w^2}\right) - \frac{\mathrm{Tr}(\mathbf{A}\mathbf{W})}{\mathrm{Tr}(\mathbf{A}\mathbf{W})+\sigma_w^2} \le \frac{2\epsilon^2}{L}, && \text{(29b)}\\
& P_a \le P_{\max}, && \text{(29c)}\\
& \mathbf{W}_{n,n} \le P_a, \forall n = 1,2,...,N, && \text{(29d)}\\
& \mathbf{W}_{N+1,N+1} = P_a, && \text{(29e)}\\
& \mathbf{W} \ge 0, && \text{(29f)}\\
& \mathrm{rank}(\mathbf{W}) = 1, && \text{(29g)}
\end{aligned}
$$

It can be seen that function (29a) is linear and the constraints (29c), (29d), (29e) and (29f) are also linear. The convexity of covertness constraint (29b) can be determine as the following Lemma.

Lemma 1: By adding a slack variable α, the constraint (29b) can be equivalently expressed as

$$
\begin{aligned}
\ln\left(1+\tfrac{\mathrm{Tr}(\mathbf{A}\mathbf{W})}{\sigma_w^2}\right) &\le \alpha && \text{(30a)}\\
\alpha - \frac{\mathrm{Tr}(\mathbf{A}\mathbf{W})}{\mathrm{Tr}(\mathbf{A}\mathbf{W})+\sigma_w^2} &\le \frac{2\epsilon^2}{L} && \text{(30b)}
\end{aligned}
$$

For known α, (30) is a convex constraint with respect to $\mathbf{W}$. We can find the optimal value of α by a one-dimensional line search. But the constraints (29g) still not being solved. To recover the rank-one solution, a PSCA iterative algorithm is used to solve problem (P2.3). To start with, the rank-one constraints can be written as [16]

$$\text{Tr}(\mathbf{W}) - \lambda_{\max}(\mathbf{W}) \leq 0, \tag{31}$$

where $\lambda_{\max}(\mathbf{W})$ is the maximal eigenvalue of $\mathbf{W}$. When $\mathbf{W} \geq 0$ hold we can get $\text{Tr}(\mathbf{W}) - \lambda_{\max}(\mathbf{W}) \geq 0$. According to this the constraint (31) is equivalent to $\text{Tr}(\mathbf{W} = \lambda_{\max}(\mathbf{W}))$, which implies that $\mathbf{W}$ has only one non-zero eigenvalue. As $\lambda_{\max}(\mathbf{W})$ is a spectral function and is convex with respect to $\mathbf{W}$. For any feasible solution $\widetilde{\mathbf{W}}$ the constraint (31) can be transform into a more strict convex constraint [21].

By applying successive convex approximation (SCA) method it is difficult to find the initial $\widetilde{\mathbf{W}}$. So a slack variable $\eta \geq 0$ is introduced to expand the feasible solution set of constraint (31). Then problem (P2.3) can be written as [16]

$$
\begin{aligned}
\text{(P2.4):} \quad & \max_{P_a, \mathbf{W}, \eta} \ \text{Tr}(\mathbf{BW}) - \tau\eta && \text{(32a)}\\
& \text{s.t.} \ (30)(29c)(29e)(29d)(29f) && \text{(32b)}\\
& \qquad \text{Tr}(\mathbf{W}) - \lambda_{\max}(\mathbf{W}) \leq \eta, && \text{(32c)}\\
& \qquad \eta \geq 0, && \text{(32d)}
\end{aligned}
$$

where $\tau > 0$ is a penalty parameter. (P1.3) and (P1.4) are equivalent when $\tau > \tau_0$.

As for constraint (32c). The function $\lambda_{\max}(\mathbf{W})$ is non-smooth, its sub-gradient can be given by $\mathbf{w}_{\max}\mathbf{w}_{\max}^H$ [22], where $\mathbf{w}_{\max}$ is the eigenvector according to the maximum eigenvalue of $\lambda_{\max}(\mathbf{W})$. The first-order restrictive approximation of $\lambda_{\max}(\mathbf{W})$ is replaced by [16]

$$\lambda_{\max}(\mathbf{W}) \geq \lambda_{\max}(\tilde{\mathbf{W}}) + \text{Tr}\left(\tilde{\mathbf{w}}_{\max}\tilde{\mathbf{w}}_{\max}^H(\mathbf{W} - \tilde{\mathbf{W}})\right), \tag{33}$$

where $\tilde{\mathbf{W}}$ is a given feasible point and $\tilde{\mathbf{w}}$ is the eigenvector for the maximum eigenvalue $\lambda_{\max}(\tilde{\mathbf{W}})$ of matrix $\tilde{\mathbf{W}}$. So the constraint (32c) can be written as [16]

$$\text{Tr}(\mathbf{W}) - \lambda_{\max}(\tilde{\mathbf{W}}) - \text{Tr}\left(\tilde{\mathbf{w}}_{\max}\tilde{\mathbf{w}}_{\max}^H(\mathbf{W} - \tilde{\mathbf{W}})\right) \leq \eta. \tag{34}$$

The above formula can be simplified as [16]

$$\mathrm{Tr}(\mathbf{W}) - \mathrm{Tr}(\tilde{\mathbf{w}}_{\max}\tilde{\mathbf{w}}_{\max}^H\mathbf{W}) \leq \eta, \tag{35}$$

because $\lambda_{\max}(\tilde{\mathbf{W}}) = \tilde{\mathbf{w}}_{\max}^H\tilde{\mathbf{W}}\tilde{\mathbf{w}}_{\max}$. The problem (P2.4) can be written as [16]

$$\text{(P2.5)}: \quad \max_{P_a,\mathbf{W},\eta} \ \mathrm{Tr}(\mathbf{B}\mathbf{W}) - \tau\eta \tag{36a}$$

$$\text{s.t.} \ \ (29\text{c}), (29\text{d}), (29\text{e}), (29\text{f}), (30), (32\text{d}), (35) \tag{36b}$$

$$\tag{36c}$$

For a given penalty parameter τ and an initial feasible solution $\tilde{\mathbf{W}}$, it can be solved by CVX [15]. The above table is the step to solve the problem (P2.2).

Algorithm 1 PSCA algorithm for Solving (P2.2) [16]

1: Given an initial feasible solution $\tilde{\mathbf{W}}^0$ and an initial penalty parameter τ^0; Given $c > 1$ and $\tau_{\max}$; Set $r = 0$

2: **repeat**

3: Solve (P2.5) with given a feasible solution $\tilde{\mathbf{W}}^r$ and obtain the current optimal solution $\{\mathbf{W}^{r+1}, P_a^{r+1}, \eta^{r+1}\}$.

4: Update $\tau^{r+1} = \min\{c\tau^r, \tau_{\max}\}$ and set $\tilde{\mathbf{W}}^{r+1} = \mathbf{W}^{r+1}$; Set the iteration number$r = r + 1$.

5: **until** Convergence.

Algorithm 1 illustrates the over all PSCA algorithm for solving problem (P2.2). Convergence in this algorithm can be understand as, for every round after calculate **W** out, we can get a value of SNR at Bob. When $\frac{SNR^{r+1}-SNR^r}{SNR^r} < 0.0001$, we can say that the solution is convergence.

4.2.2 Low-Complexity Algorithm

Based on the former section, a two-stage algorithm is developed to do the formulation. The reflection beamforming of IRS is designed in the first stage, the transmit power of Alice is designed in the second stage.

IRS Beamforming Design: We notice that $D(\mathbb{P}_1|\mathbb{P}_0)$ is covertness, (27c) is a monotonically increasing function of $P_a\mathbf{u}^H\mathbf{A}\mathbf{u}$, which means the hidden level is controlled by energy received at Willie. According to this, the ratio between energy received at Bob and energy received at Willie can be expressed as $\frac{P_a\mathbf{u}^H\mathbf{B}\mathbf{u}}{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}}$. The optimization problem can be developed into [16]

$$(\text{P3}): \quad \max_{\mathbf{u}} \; \frac{\mathbf{u}^H\mathbf{B}\mathbf{u}}{\mathbf{u}^H\mathbf{A}\mathbf{u}} \tag{37a}$$

$$\text{s.t. } |u_n| \le 1, \forall n = 1, 2, ..., N, \tag{37b}$$

$$|u_{N+1}| = 1. \tag{37c}$$

As the objective function (37a) and the constraint (37c) are all non-convex and it is impossible to be solved. We should start with developing a lower bound for the objective function (37a) and then develop a SCA algorithm to solve problem (P3). According to [23], (37a) can be changed into [16]

$$\frac{\mathbf{u}^H\mathbf{B}\mathbf{u}}{\mathbf{u}^H\mathbf{A}\mathbf{u}} \ge \mathrm{Re}(\mathbf{f}^H\mathbf{u}) + \zeta_1, \tag{38}$$

where [16]

$$\mathbf{f} = \left(\frac{\mathbf{B}}{\tilde{\mathbf{u}}^H\mathbf{A}\tilde{\mathbf{u}}} - \frac{(\mathbf{A} - \lambda_{\max}(\mathbf{A})\mathbf{I}_{N+1})\tilde{\mathbf{u}}^H\mathbf{B}\tilde{\mathbf{u}}}{(\tilde{\mathbf{u}}^H\mathbf{A}\tilde{\mathbf{u}})^2} \right), \tag{39}$$

$$\zeta_1 = \frac{\tilde{\mathbf{u}}^H\mathbf{B}\tilde{\mathbf{u}}}{\tilde{\mathbf{u}}^H\mathbf{A}\tilde{\mathbf{u}}} - \frac{\lambda_{\max}(\mathbf{A})(||\mathbf{u}||^2 + ||\tilde{\mathbf{u}}||^2)\tilde{\mathbf{u}}^H\mathbf{B}\tilde{\mathbf{u}}}{(\tilde{\mathbf{u}}^H\mathbf{A}\tilde{\mathbf{u}})^2}, \tag{40}$$

$\tilde{\mathbf{u}}$ is a given value calculated in the loop. As is defined earlier that $\mathbf{A}$ is a rank-one matrix and $\lambda_{\max}(\mathbf{A}) = \bar{\mathbf{a}}^H\bar{\mathbf{a}}$, where $\bar{\mathbf{a}} = \begin{bmatrix}\mathbf{a}^H & h_{aw}^*\end{bmatrix}^H$. Going after the constraints (37b) and (37c), a lower bound on ζ_1 is put as [16]

$$\zeta_1 \ge \zeta_2 = \frac{\tilde{\mathbf{u}}^H\mathbf{B}\tilde{\mathbf{u}}}{\tilde{\mathbf{u}}^H\mathbf{A}\tilde{\mathbf{u}}} - \frac{2\bar{\mathbf{a}}^H\bar{\mathbf{a}}(N+1))\tilde{\mathbf{u}}^H\mathbf{B}\tilde{\mathbf{u}}}{(\tilde{\mathbf{u}}^H\mathbf{A}\tilde{\mathbf{u}})^2}, \tag{41}$$

According to this, problem (P3) can be changed into [16]

$$(\text{P3.1}): \quad \max_{\mathbf{u}} \; 2\mathrm{Re}(\mathbf{f}^H\mathbf{u}) + \zeta_2 \tag{42a}$$

$$\text{s.t. } |u_n| \le 1, \forall n = 1, 2, ..., N, \tag{42b}$$

$$|u_{N+1}| = 1. \tag{42c}$$

In order to solve the above problem (P3.1), the objective function (42a) can be written into

[16]

$$2\text{Re}\left(\sum_{n=1}^{N+1}(|f_n||u_n|e^{j(\arg(u_n)-\arg(f_n))})\right)+\zeta_2, \tag{43}$$

where f_n is the *n*th element of $\mathbf{f}$. If $\tilde{\mathbf{u}}$ is known, the optimal solution to problem (P2.1) is $\arg(u_n)=\arg(f_n)$ and $|u_n|=1, \forall n$.

2) Transmit Power Design: For a given $\mathbf{u}$, problem (P2.1) can be simplified into [16]

$$\text{(P4):} \quad \max_{P_a} \ P_a\mathbf{u}^H\mathbf{B}\mathbf{u} \tag{44a}$$

$$\text{s.t.} \ \ln\left(1+\frac{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}}{\sigma_w^2}\right)-\frac{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}}{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}+\sigma_w^2}\leq\frac{2\epsilon^2}{L} \tag{44b}$$

$$P_a\leq P_{\max}. \tag{44c}$$

We notice that $\frac{2\epsilon^2}{L}$ is usually a small value. A conservative approximation of constraint (44b) can be expressed as [16]

$$\frac{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}}{\sigma_w^2}-\frac{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}}{P_a\mathbf{u}^H\mathbf{A}\mathbf{u}+\sigma_w^2}\leq\frac{2\epsilon^2}{L}. \tag{45}$$

Then problem (P4) can be changed into [16]

$$\text{(P4.1):} \quad \max_{P_a} \ P_a\mathbf{u}^H\mathbf{B}\mathbf{u} \tag{46a}$$

$$\text{s.t.} \ (44\text{b}), (44\text{c}), (45). \tag{46b}$$

For the LHS of constraint (45) and the objective function of (P3.1) they are monotonically increasing functions of P_a. According to this, the constraint (45) can take the equal sign, which lead to $P_a=\frac{\sigma_w^2(\epsilon^2\pm\sqrt{\epsilon^4+2\epsilon^2L})}{L\mathbf{u}^H\mathbf{A}\mathbf{u}}$. As $\epsilon^2\leq\sqrt{\epsilon^4+2\epsilon^2L}$ and $0\leq P_a\leq P_{max}$, the optimal transmit power can be calculate as [16]

$$P_a^*=\min\left\{\frac{\sigma_w^2(\epsilon^2+\sqrt{\epsilon^4+2\epsilon^2L})}{L\mathbf{u}^H\mathbf{A}\mathbf{u}}, P_{\max}\right\} \tag{47}$$

The steps to solve the low-complexity algorithm are:

The above table shows the steps of the low-complexity method. As it can be seen from the

Algorithm 2 Proposed Low-Complexity Algorithm [16]

1:Given an initial feasible solution $\tilde{\mathbf{u}}^0 = ones(1, N+1)$ and set the iteration index $r = 0$.

2: **repeat**

3: Compute $\mathbf{u} = e^{j\arg(\mathbf{f})}$ to obtain the current optimal solution of problem (P3.1).

4: Update $\tilde{\mathbf{u}}^r = \mathbf{u}$ and set $r = r + 1$.

5: **until** Convergence.

6: Compute P_a according to (47).

table that we should start with setting an initial value to $\tilde{\mathbf{u}}^0$. Convergence in this algorithm can be understand as, for every round after calculate **u** out, we can get a value of P_a, when $\frac{P_a^{r+1} - P_a^r}{P_a^r} < 0.0001$, we can say that the answer is convergence and the calculation can be stopped.

5 Simulation Results

This chapter shows the simulation results of wireless communication aided with IRS and adding latency constraints.

5.1 Simulation of IRS-aided Wireless Communication

In the simulation part, there are two kind of simulation, one is with fixed IRS $N = 25$, change AP-user horizontal distance and see how the transmit power at the AP varies. The other is with fixed AP-user horizontal distance $d = 50m$, $d = 41m$ and $d = 15m$ change the IRS N and see how the transmit power at the AP changes. In the end 4 figures will be plotted.

To start with, the distance loss must be considered, the model of the distance dependent path loss model is:

$$L(d) = C_0 \left(\frac{d}{D_0}\right)^{-\alpha} \tag{48}$$

where C_0 is the path loss at the reference distance $D_0 = 1m$, d is the distance to the IRS surface and α is the path loss exponent.

Rician fading channel model can be used to discribe small-scale fading for all channels between AP-user, AP-IRS and IRS-user. The AP-IRS channel $\mathbf{G}$ is given by:

$$\mathbf{G} = \sqrt{\frac{\beta_{A1}}{1+\beta_{A1}}}\mathbf{G}^{LoS} + \sqrt{\frac{1}{1+\beta_{A1}}}\mathbf{G}^{NLoS} \tag{49}$$

where β_{A1} is the Rician factor and G^{LoS} and G^{NLoS} are the deterministic LoS (specular) and Rayleigh fading channel components respectively. G is of rank 1. The representation of each is the product of channel fading and the path loss model.

The Rician factor and the pass loss components of each channel is list below.

	Rician factor (β)	Path loss component (α)
AP-IRS (AI)	Inf	2
AP-user (AU)	0	3.5
IRS-user (IU)	0	2.8

The simulation should run 1000 times in order to make the result more accurate. Other system parameters are given as: $C_0 = -30dB$, $\sigma = -80dB$ and $d_0 = 51m$. The user SNR target $\gamma = 10dB$ and $M = 4$.

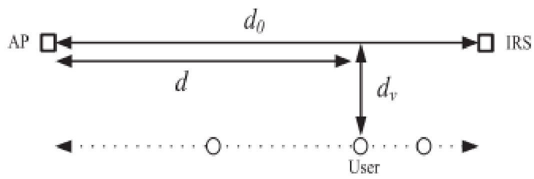


Figure 5: Distance model between AP, IRS and user.

The vertical distance between user and AP is $d_v = 2m$. As we can see from Figure 5 that the distance between AP-user is $d_{AU} = \sqrt{d^2 + d_v^2}$ and the distance between IRS-user is $d_{IU} = \sqrt{(d_0 - d)^2 + d_v^2}$. In the result, 7 schemes are being compared.

- **Lower bound**: the minimum transmit power based on the SDR problem (8);
- **SDR**: the solution to the optimal question by using SDR in part 3;
- **Alternating optimization**: the solution to part 4 by using alternating optimization, the initial $\bar{w} = \frac{h_d}{||h_d||}$, then calculate the optimized phase and based on the phase shift get the optimized direction then calculate the power;
- **AP-user MRT**: set $\bar{w} = \frac{h_d}{||h_d||}$ calculate the phase shift and calculate the optimal power;
- **AP-IRS MRT**: set $\bar{w} = \frac{g}{||g||}$ calculate the phase shift and calculate the optimal power;
- **Random phase shift**: set the phase to be random value between $[0, 2\pi]$, calculate $\bar{w}$ and calculate the optimal solution;
- **Without IRS**: which means the variable Θ is 0 and the optimal solution $P^* = \frac{\gamma\sigma^2}{||h_d^H||^2}$.

Below are the simulation results.

1) AP Transmit Power Versus AP-User Distance: In Figure 6 we compare the power at AP by the 7 schemes versus the horizontal distance between the AP and the user. First as it can be seen from the figure that comparing with the lower bound case, both of the two methods we use achieve near-optimal transmit power and they both better than the other schemes. Second, for the one without IRS, because of the attenuation of the signals, the users far from AP also needs signals with great power. But this problem is solved when deploying IRS, for we can see in the picture that with the increase of the distance, the lines with IRS all go down, which means that a greater AP-user distance do not result to a bigger transmit power and IRS can display the signal well by adjust the magnitude and phase. For the a big AP-user distance, the user-IRS distance is short, and the user can receive a strong signal from IRS. We can also find that when distance

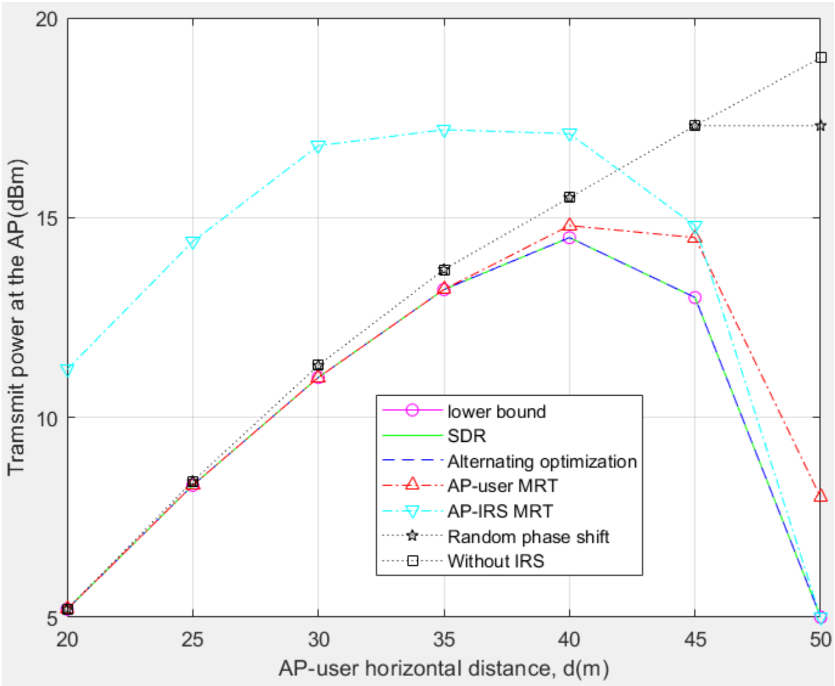
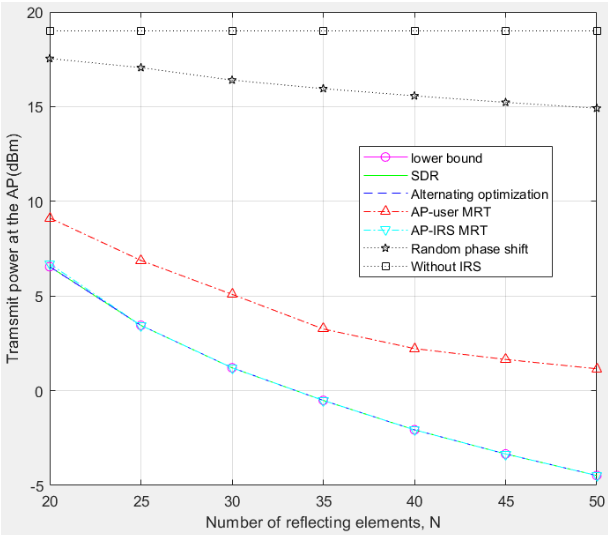


Figure 6: Transmit power at the AP vs AP-user horizontal distance.

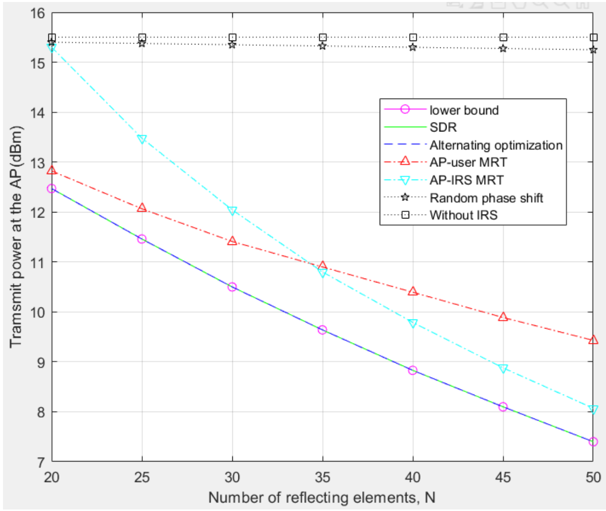
between AP-user or IRS-user is short, the requirement to transmit power is lower than that are both long.

We can also find that the AP-user MRT is closer to the optimal plan when the user is closer to AP, while the AP-IRS MRT is of bad condition and even worse than that without IRS, it is closer to the optimal plan when the user is closer to IRS. This is because when the user is closer to AP, the signal received by user is controlled by AP-user direct link. Under the AP-IRS MRT case when the user is near AP, we should let AP directly send signal to user instead of using IRS.

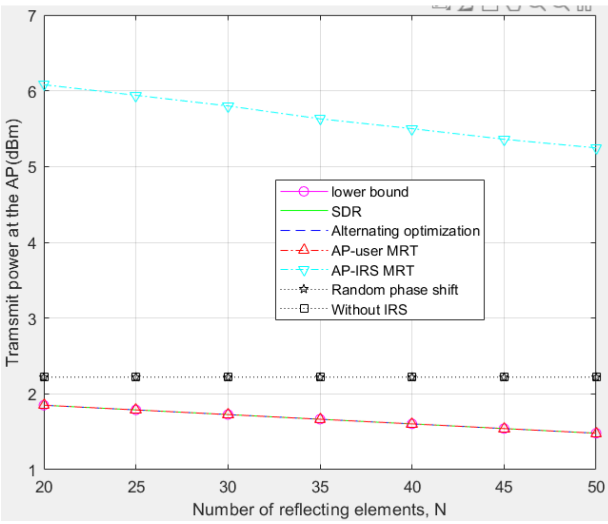
2) AP Transmit Power Versus Number of Reflecting Elements: In Figure 7, we compare the relationship between the transmitting power at AP versus the number of reflective elements at IRS when $d = 50$, 40 and $15m$, respectively. Looking through those 3 schemes, the common thing is that with the increase of number of reflecting elements, the transmit power at AP will decrease. In 7a when $d = 50m$ we can find that the AP-IRS MRT is comparable to the optimal solution. This is because more signal received from IRS rather than directly from AP. In 7c when $d = 15m$ we can find that the AP-user MRT is comparable to the optimal solution, while the AP-IRS is even worse than that without IRS. In 7b when $d = 40m$, with few reflecting elements, the AP-user MRT is better while with much reflecting elements, the AP-IRS MRT is better, the more reflecting elements being used, the nearer the transmit power is to the optimal



(a) d=50m



(b) d=40m



(c) d=15m

Figure 7: Transmit power at the AP vs number of reflecting elements.

one.

3) AP Transmit Power Versus AP-IRS Distance: In Figure 8, we compare the relationship between the transmitting power at AP verses distance between AP and IRS when $d = 50m$. As it can be seen from the figure that the two methods perform good results, which is better than that without IRS. Comparing with the lower bound, we can say that they perform near-optimal solution. From the result we can see that it is more useful to place IRS when the IRS is close to user or AP. For the AP-user MRT, it is near optimal while for the AP-IRS MRT case it is only useful when the IRS is close to the user or AP.

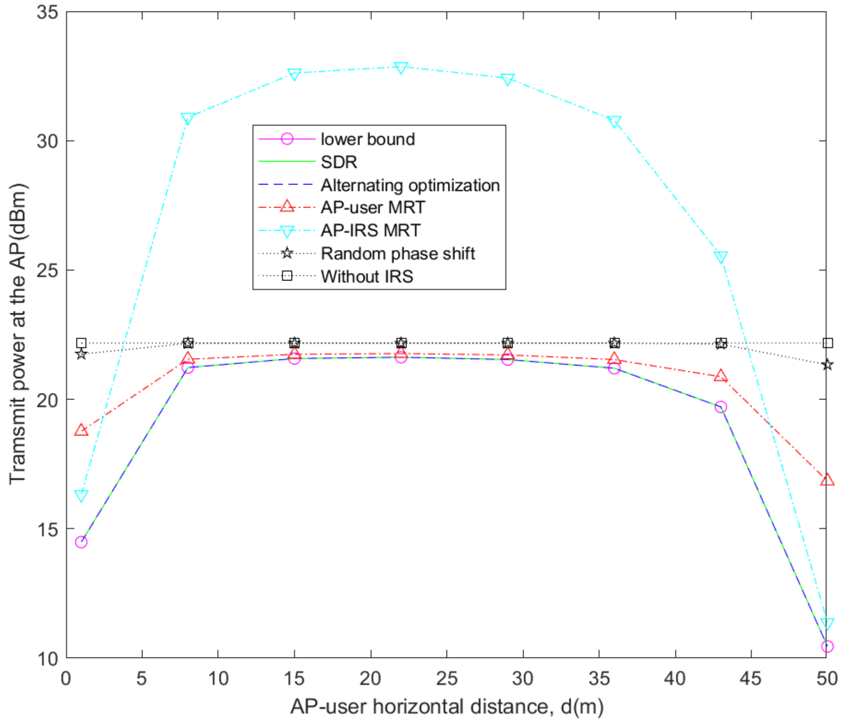


Figure 8: Transmit power at the AP vs AP-IRS distance.

5.2 Simulation of IRS adding Latency Constraints

In the simulation part, there are two kind of simulation, one is with fixed location of IRS, change the IRS N and see how the SNR at Bob change. The other is with fixed IRS $N = 50$ change the horizontal location of IRS and see how the SNR at Bob change. In the end 2 figures will be plotted.

The location of IRS, Alice, Bob and Willie is considered in a three-dimensional coordinate system and their location are $(0, 0, 5)m$, $(100, 0, 5)m$, $(70, 10, 0)m$ and $(100, 10, 0)m$, respectively. The noise variance at both Bob and Willie are set to $\sigma_b^2 = \sigma_w^2 = -80dBm$, The maximum transmit power at Alice is $P_{max} = 36dBm$, the number of channel uses $L = 100$. Without special requirement, the value of covertness level is $\epsilon = 0.1$. In order to simplify the calculation, IRS reflection amplitudes are fixed to 1.

As is mention in the previous subsection, there are Rician fading and Tayleigh fading considered in the channel. Because of the location of IRS is usually based on Alice's location, the channel from Alice to IRS is randomly selected from the Rician Fading, the Rician factor is set to be $5dB$, for other channels they can all be realized drawing from Rayleigh fading.

The large-scale path loss also needs to be considered. From node i to node j, the component can be expressed as:

$$\chi_{ij} = \beta_0 \left(\frac{d_{ij}}{d_0}\right)^{-\alpha_{ij}} \tag{50}$$

where β_0 is the channel power gain at $d_0 = 1m$, and it is set to be $\beta_0 = -30dB$, d_{ij} is the distance between node i and node j, α_{ij} is the path loss exponents they are $\alpha_{ar} = 2.4$, $\alpha_{ar} = 4.2$, $\alpha_{ar} = 4.2$, $\alpha_{ar} = 3$ and $\alpha_{ar} = 3$.

In order to reduce error and avoid chance, the simulation shou run 1000 times.

In the result, 4 schemes are being compared.

- **Upper bound**: without considering the rank-one constraint (29g) solve the problem (P2.3);
- **PSCA algorithm**: jointly design Alice's transmit power and phase shift of each IRS reflection element by using Algorithm 1;
- **Low-complexity algorithm**: using Algorithm 2 to solve the problem;
- **Without IRS**: no IRS is considered in the system and only Alice's transmit power P_a needs to be designed.

Below are the simulation results.

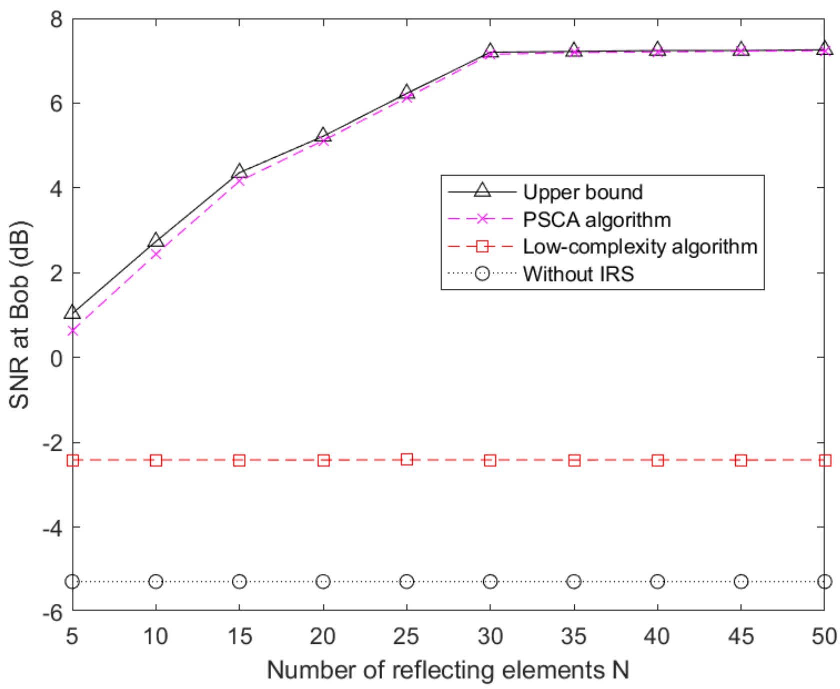


Figure 9: SNR at Bob Versus Number of Reflecting Elements.

1) SNR at Bob Versus Number of Reflecting Elements: In figure 9 we compare the relationship between SNR at Bob versus the number of reflecting elements at IRS. As it can been seen from the whole that the one with IRS receives a better result than that without IRS, this means IRS is benefit to the communication system for with big SNR at Bob, the system performs well, but with some problem, the low-complexity algorithm runs not very well so in the simulation next part, it is not going to be used. With the increasing of reflecting elements, SNR at Bob will increase sharply and then the increase become slow. So in order to simplify the system, not too much of reflecting elements can be used to achieve a good result.

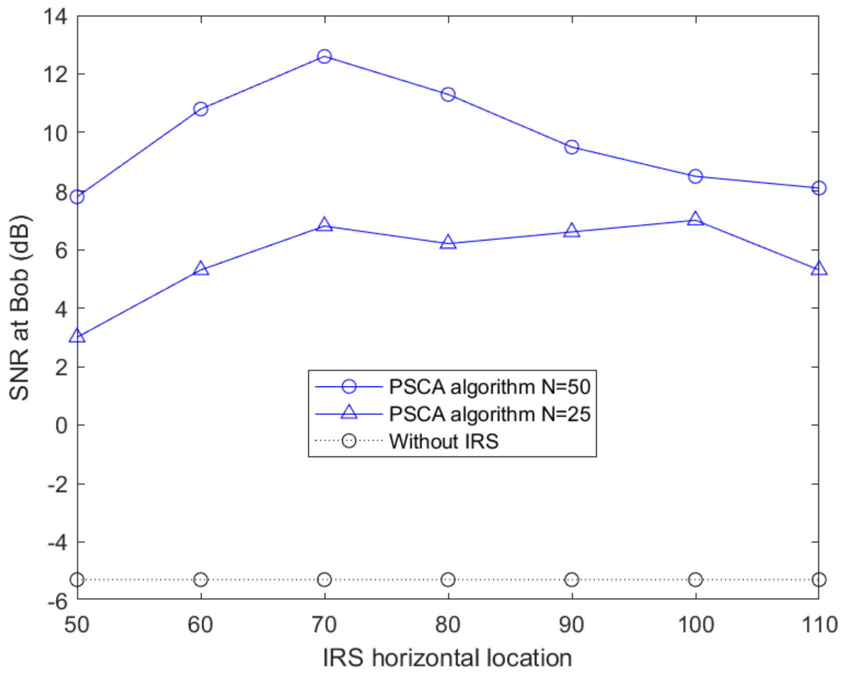


Figure 10: SNR at Bob Versus IRS horizontal location.

2) SNR at Bob Versus IRS horizontal location: In figure 10 we compare the relationship between SNR at Bob versus IRS horizontal location when number of reflecting elements $N = 25$ and 50 respectively. We can see that the results are all work better than that without IRS, and the overall performance of $N = 50$ is always better than when $N = 25$. When $N = 50$, the optimal solution will be closer to Bob while when $N = 25$, the optimal solution will be closer to Willie. This is because with bigger N the requirement is much easier to be satisfied.

6 Conclusions and Future Work

In this chapter, the results of this paper is being summarized and the directions of what we can do in the future is given.

6.1 Conclusion

From the above simulation result, we can see that IRS is a cost-effective material which can improve the influence of latency by adjusting the phase shift and the magnitude of the received signal and without increase the power consumption. As the material of IRS is of low cost, it can be widely used in out daily life without waste of money. In this paper, the IRS system model design and the model adding latency constraint are introduced.

As for the IRS system model beamforming design, we use two kinds of methods, one is semidefinite relaxation, the other is alternating optimization. The results show that when using IRS, the results are always better than that without using IRS and they are all of optimal solution. So IRS can design the direction of the beamforming to make it perform better. Under the condition of AP-IRS MRT it only performs better when the IRS is close to AP or user, for the AP-user MRT, it always perform better than without IRS but less better than the optimal one.

For the system adding latency constraint, two kinds of methods are being used, one is PSCA, the other is low-complexity algorithm. PSCA algorithm runs good while the low-complexity do not have a good results. From the result we can see that SNR at Bob is always better than the one without IRS, which means using IRS can solve the delay problem in some ways.

The methods listed above are all a little complex, in the future, more simple methods can be used to solve these problems.

6.2 Future Work

In the process of finishing the project, I have successfully achieved the goal and solved a lot of difficult question. In the future I can try to find and solve some existing problem about this project and find new methods to continue this study. The future works are listed as below:

- Using a new method to reduce the complexity of the algorithm. When programming the system with latency constraints, I spend a lot of time on coding and although the problem is being solved, the code is too complex. I have been provided with a simpler method by using python, but because of time limited, it is impossible for me to use this kind of method.

- In my simulation, the amplitude of IRS is fixed to be 1, in the future I can consider how to solve the problem when the amplitude is not equal to 1. In order to reduce the complexity of the simulation, I just do this approximation, but it is not accurate for the magnitude of the signal reflected by IRS cannot be totally used.
- In the project, I studied a lot of optimization methods like alternating optimization, etc., they are widely used in a lot of areas, the common things between them is that we should change the convex objects into non-convex ones to solve the problem. As they are ways to solve problems, in the future we can apply it to other problems and make them more widely used and concrete in our mind.
- Improve the performance of the two systems. Because of time limitation and lack of training there are still some ways by which I can improve the result. The result I have right now are some how not optimal enough especially some of which are far too worse. In the future, I can try to make appropriate adjustment to the methods I use and using more training data to improve the result of the two systems.

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