Artificial-Noise-Aided Secure MIMO Wireless Communications via Intelligent Reflecting Surface

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Abstract

This paper considers a MIMO secure wireless communication system aided by the physical layer security technique of sending artificial-noise (AN). To further enhance the system security, the advanced intelligent reflecting surface (IRS) is invoked in the AN-aided communication system, where the base station (BS), the legitimate information receiver (IR) and eavesdropper (Eve) are all equipped with multiple antennas. With the aim for maximizing the system secrecy rate (SR), the transmit precoding (TPC) matrix at the BS, the covariance matrix of AN and the phase shift coefficients at the IRS are jointly optimized subject to the constrains of transmit power limit and unit modulus of IRS phase shifts. Then, the secrecy rate maximization (SRM) problem is formulated and investigated, which is a non-convex problem with multiple coupled variables. To tackle it, we propose to employ the block coordinate descent (BCD) algorithm, which can alternatively update the TPC matrix, AN covariance matrix, and phase shifts while keeping the SR non-descending. Specifically, the optimal TPC matrix and AN covariance matrix are derived by Lagrangian multiplier method, and the optimal phase shifts are obtained by the Majorization-Minimization (MM) algorithm. Since all these variables can be calculated in closed form, the proposed algorithm is very efficient. Finally, simulation results validate the effectiveness of enhancing the system security via an IRS.

Index Terms

Intelligent Reflecting Surface (IRS), Large Intelligent Surface (LIS), Manifold Optimization, Multicell Communications, MIMO.

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I. INTRODUCTION

The next-generation (i.e., 6G) communication is expected to be a sustainable green, cost-effective and secure communication system [1]. In particular, secure communication is crucially important in 6G communication network since communication environments become increasingly complicated and the security of private information is imperative [2]. The information security using cryptographic encryption (in the network layer) is a conventional secure communication technique, which suffers from the vulnerabilities, such as secret key distribution, protection and management [3]. Unlike this network layer security approach, the physical layer security can guarantee good security performance bypassing the relevant manipulations on the secret key, thus is more attractive for the academia and industry [4]. There are various physical-layer secrecy scenarios. The first one is the classical physical-layer secrecy setting where there is one legitimate information receiver (IR) and one eavesdropper (Eve) operating over a single-input-single-output (SISO) channel (i.e., the so-called three-terminal SISO Gaussian wiretap channel) [5], [6]. The second one considers the physical-layer secrecy with an IR and Eve operating over a multiple-input-single-output (MISO) channel, which is called as three-terminal MISO Gaussian wiretap channel. The third one is a renewed and timely scenario with one IR and one Eve operating over a multiple-input-multiple-output (MIMO) channel, which is denoted as three-terminal MIMO Gaussian wiretap channel [7], [8] and is the focus of this paper. For MIMO systems, a novel idea in physical-layer secrecy is to transmit artificial noise (AN) from the base station (BS) to contaminate the Eve’s received signal [9]–[11]. For these AN-aided methods, a portion of transmit power is assigned to the artificially generated noise to interfere the Eve, which should be carefully designed. For AN-aided secrecy systems, while most of the existing AN-aided design papers focused on the MISO wiretap channel and null-space AN [7], [12], designing the transmit precoding (TPC) matrix together with AN covariance matrix for the MIMO wiretap channel is more challenging [13].

In general, the secrecy rate (SR) achieved by the mutual information difference between the legitimate IR and the Eve is limited by the channel difference between the BS-IR link and the BS-Eve link. The AN-aided method can further improve the SR, but it consumes the transmit power destined for the legitimate IR. When the transmit power is confined, the performance bottleneck always exists for the AN-aided secure communication. To conquer the dilemma, the recently proposed intelligent reflecting surface (IRS) technique can be exploited. Since higher SR can be achieved by enhancing the channel quality in the BS-IR link and degrading the channel
condition in the BS-Eve link, the IRS can serve as a powerful complement to AN-aided secure communication due to its capability of reconfiguring the wireless propagation environment. The IRS technique has been regarded as a revolutionary technique to control and programme the wireless environment [14], [15]. An IRS comprises an array of reflecting elements, which can reflect the incident electromagnetic (EM) wave passively, thus alter the phase shift of EM wave [16]. Hence, by smartly tuning the phase shifts with a preprogrammed controller, the direct signals from the BS and the reflected signals from the IRS can be combined constructively or destructively according to different requirements. In comparison to the existing related techniques which the IRS resembles, such as active intelligent surface [17], traditional reflecting surfaces [18], backscatter communication [19] and amplify-and-forward (AF) relay [20], the IRSs have the advantages of flexible reconfiguration on the phase shifts in real time, minor additional power consumption, easy installation with many reflecting elements, etc. Furthermore, due to the light weight and compact size, the IRS can be integrated into the traditional communication systems with minor modifications [21]. Because of these appealing virtues, IRS has introduced into various wireless communication systems., including the single-user case [22], [23], the downlink multiuser case [16], [24]–[27], mobile edge computing [28], wireless information and power transfer design [29], and the physical layer security design [30]–[33].

IRS is promising to strengthen the system security of wireless communication. In [30], [32], [34], the authors investigated the problem of maximizing the achievable SR in a secure MISO communication system aided by IRS, where both the legitimate user and eavesdropper are equipped with a single antenna. The TPC matrix at the BS and the phase shifts at the IRS were optimized by an alternate optimization (AO) strategy. To handle the nonconvex unit modulus constraint, the semidefinite relaxation (SDR) [35], majorization-minimization (MM) [16], [36], complex circle manifold (CCM) [37] techniques were proposed to optimize phase shifts. An IRS-assisted MISO secure communication with a single IR and single Eve was also considered in [31], but it was limited to a special scenario, where the Eve has a stronger channel than the IR, and the two channels from BS to Eve and IR are highly correlated. Under this assumption, the transmit beamforming and the IRS reflection beamforming are jointly optimized to improve the SR. Similarly, a secure IRS-assisted downlink MISO broadcast system was considered in [33], and it assumes that multiple legitimate IRs and multiple Eves are in the same directions to the BS, which implies that the IR channels are highly correlated with the Eve channels. [38] considered the transmission design for
an IRS-aided secure MISO communication with a single IR and single Eve, in which the system energy consumption is minimized under two assumptions that the channels of access point (AP)-IRS links are rank-one and full-rank. An IRS-assisted MISO network with cooperative jamming was investigated in [2]. The physical layer security in a simultaneous wireless information and power transfer (SWIPT) system was considered with the aid of IRS [39]. However, there are a paucity of papers considering the IRS-assisted secure communication with AN. A secure MISO communication system aided by the transmit jamming and AN was considered in [40], where a large number of Eves exist, and the AN beamforming vector and jamming vector were optimized to reap the additional degree of freedom (DoF) brought by the IRS. [41] investigated the resource allocation problem in an IRS-assisted MISO communication by jointly optimizing the beamforming vectors, the phase shifts of the IRS, and AN covariance matrix for secrecy rate maximization (SRM), but the direct BS-IRs links and direct BS-Eves link are assumed to be blocked.

Although a few papers have studied security enhancement for an AN-aided system through the IRS, the existing papers related to this topic either only studied the MISO scenario or assumed special settings to the channels. The investigation on the MIMO scenario with general channel settings is absent in the existing literature. Hence, we investigate this problem in this paper by employing an IRS in an AN-aided MIMO communication system for the physical layer security enhancement. Specifically, by carefully designing the phase shifts of the IRS, the reflected signals are combined with the direct signals constructively for enhancing the data rate at the IR and destructively for decreasing the rate at the Eve. As a result, the TPC matrix and AN covariance matrix at the BS can be designed flexibly with a higher DoF than the case without IRS. In this work, the TPC matrix, AN covariance matrix and the phase shift matrix are jointly optimized. Since these optimization variables are highly coupled, an efficient algorithm based on the block coordinate descent (BCD) and MM techniques for solving the problem is proposed.

We summarize our main contributions as follows:

1) This is the first research on exploiting an IRS to enhance security in AN-aided MIMO communication systems. Specifically, an SRM problem is formulated by jointly optimizing the TPC matrix and AN covariance matrix at the BS, together with the phase shifts of the IRS subject to maximum transmit power limit and the unit modulus constraint of the phase shifters. The objective function (OF) of this problem is the difference of two Shannon capacity expressions, thus is not jointly concave over the three highly-coupled variables. To
handle it, the popular minimum mean-square error (MMSE) algorithm is used to reformulate the SRM problem.

2) The BCD algorithm is exploited to optimize the variables alternately. Firstly, given the phase shifts of IRS, the optimal TPC matrix and AN covariance matrix are obtained in closed form by utilizing the Lagrangian multiplier method. Then, given the TPC matrix and AN covariance matrix, the optimization problem for IRS phase shifts is transformed by sophisticated matrix manipulations into a quadratically constrained quadratic program (QCQP) problem subject to unit modulus constraints. To solve it, the MM algorithm is utilized, where the phase shifts are derived in closed form iteratively. Based on the BCD-MM algorithm, the original formulated SRM problem can be solved efficiently.

3) The simulation results confirm that on the one hand, the IRS can greatly enhance the security of an AN-aided MIMO communication system; on the other hand, the phase shifts of IRS should be properly optimized. Simulation results also show that larger IRS element number and more transmit power is beneficial to the security. Moreover, properly-selected IRS location and good channel states of the IRS-related links are important to realize the full potential of IRS.

We organize the remainder of this paper as follows. Section II provides the signal model of an AN-aided MIMO communication system assisted by the IRS, and the SRM problem formulation. The SRM problem is reformulated in Section III, where the BCD-MM algorithm is proposed to optimize the TPC matrix, AN covariance matrix and phase shifts of IRS. In Section IV, numerical simulations are given to validate the algorithm efficiency and security enhancement. Section V concludes this paper.

**Notations:** Throughout this paper, boldface lower case, boldface upper case and regular letters are used to denote vectors, matrices, and scalars respectively. $\mathbf{X} \odot \mathbf{Y}$ is the Hadamard product of $\mathbf{X}$ and $\mathbf{Y}$. $\text{Tr}(\mathbf{X})$ and $|\mathbf{X}|$ denote the trace and determinant of $\mathbf{X}$ respectively. $\mathbb{C}^{M \times N}$ denotes the space of $M \times N$ complex matrices. $\text{Re}\{\cdot\}$ and $\text{arg}\{\cdot\}$ denote the real part of a complex value and the extraction of phase information respectively. $\text{diag}(\cdot)$ is the operator for diagonalization. $\mathcal{CN}(\mu, \mathbf{Z})$ represents a circularly symmetric complex gaussian (CSCG) random vector with mean $\mu$ and covariance matrix $\mathbf{Z}$. $(\cdot)^\top$, $(\cdot)^\text{H}$ and $(\cdot)^*$ denote the transpose, Hermitian and conjugate operators respectively. $(\cdot)^*$ stands for the optimal value, and $(\cdot)^\dagger$ means the pseudo-inverse.
II. SIGNAL MODEL AND PROBLEM FORMULATION

A. Signal Model

We consider an IRS-aided communication network shown in Fig. 1 that consists of a BS, a legitimate IR and an Eve, all of which are equipped with multiple antennas. The number of transmit antennas at the BS is $N_T \geq 1$, and the numbers of receive antennas at the legitimate IR and Eve are $N_I \geq 1$ and $N_E \geq 1$ respectively. To ensure secure transmission from the BS to the IR, the AN is sent from the BS to interfere the eavesdropper to achieve strong secrecy.

With above assumptions, the BS employed the TPC matrix to transmit data streams with AN. The transmitted signal can be modeled as

$$x = Vs + n,$$  \hspace{1cm} (1)

where $V \in \mathbb{C}^{N_T \times d}$ is the TPC matrix; the number of data streams is $d \leq \min(N_T, N_I)$; the transmitted data towards the IR is $s \sim \mathcal{CN}(0, I_d)$; and $n \in \mathcal{CN}(0, Z)$ represents the AN random vector with zero mean and covariance matrix $Z$.

Assuming that the wireless signals are propagated in a non-dispersive and narrow-band way, we model the equivalent channels of the BS-IRS link, the BS-IR link, the BS-Eve link, the IRS-IR link, the IRS-Eve link by the matrices $G \in \mathbb{C}^{M \times N_T}$, $H_{b,I} \in \mathbb{C}^{N_I \times N_T}$, $H_{b,E} \in \mathbb{C}^{N_E \times N_T}$, $H_{R,I} \in \mathbb{C}^{N_I \times M}$, $H_{R,E} \in \mathbb{C}^{N_E \times M}$, respectively. The phase shift coefficients of IRS are collected in a diagonal matrix defined by $\Phi = \text{diag}\{e^{j\theta_1}, \ldots, e^{j\theta_m}, \ldots, e^{j\theta_M}\}$, where $\theta_m \in [0, 2\pi]$ denotes the phase shift of the $m$-th reflection element. The multi-path signals that have been reflected...
by multiple times are considered to be absorbed and diffracted, then the signal received at the legitimate IR is given by
\[ y_I = (H_{b,I} + H_{R,I} \Phi G)x + n_I, \]  
(2)
where \( n_I \) is the random noise vector at IR obeying the distribution \( n_I \sim \mathcal{CN}(0, \sigma_I^2 I_{N_I}) \). The signal received at the Eve is
\[ y_E = (H_{b,E} + H_{R,E} \Phi G)x + n_E, \]  
(3)
where \( n_E \) is the Eve’s noise vector following the distribution \( n_E \sim \mathcal{CN}(0, \sigma_E^2 I_{N_E}) \).

Assume that the BS has acquired the prior information of all the channel state informations (CSIs). Then the BS takes the responsibility of optimizing the IRS phase shifts and feeding them back to the IRS controller. Upon substituting \( x \) into (2), \( y_I \) can be re-writtten as
\[ y_I = \hat{H}_I(Vs + n) + n_I = \hat{H}_I V s + \hat{H}_I n + n_I, \]  
(4)
where \( \hat{H}_I \triangleq H_{b,I} + H_{R,I} \Phi G \) is defined as the equivalent channel spanning from the BS to the legitimate IR. Then, the data rate (bit/s/Hz) achieved by the legitimate IR is given by
\[ R_I(V, \Phi, Z) = \log \left| I + \hat{H}_I V V^H \hat{H}_I^H J_I^{-1} \right|, \]  
(5)
where \( J_I \) is the interference-plus-noise covariance matrix given by \( J_I \triangleq \hat{H}_I Z \hat{H}_I^H + \sigma_I^2 I_{N_I} \).

Upon substituting \( x \) into (3), \( y_E \) can be re-writtten as
\[ y_E = \hat{H}_E(Vs + n) + n_E = \hat{H}_E V s + \hat{H}_E n + n_E, \]  
(6)
where \( \hat{H}_E \triangleq H_{b,E} + H_{R,E} \Phi G \) is defined as the equivalent channel spanning from the BS to the Eve. Then, the data rate (bit/s/Hz) achieved by the Eve is given by
\[ R_E(V, \Phi, Z) = \log \left| I + \hat{H}_E V V^H \hat{H}_E^H J_E^{-1} \right|, \]  
(7)
where \( J_E \) is the interference-plus-noise covariance matrix given by \( J_E \triangleq \hat{H}_E Z \hat{H}_E^H + \sigma_E^2 I_{N_E} \). The achievable secrecy rate is given by
\[ C_{AN}(V, \Phi, Z) = [R_I(V, \Phi, Z) - R_E(V, \Phi, Z)]^+ \]
\[ = \log \left| I + \hat{H}_I V V^H \hat{H}_I^H J_I^{-1} \right| - \log \left| I + \hat{H}_E V V^H \hat{H}_E^H J_E^{-1} \right| \]
\[ = \log \left| I + \hat{H}_I V V^H \hat{H}_I^H (\hat{H}_I Z \hat{H}_I^H + \sigma_I^2 I_{N_I})^{-1} \right| \]
\[- \log \left| I + \hat{H}_E V V^H \hat{H}_E^H (\hat{H}_E Z \hat{H}_E^H + \sigma_E^2 I_{N_E})^{-1} \right|. \]  
(8)
B. Problem Formulation

With the aim for maximizing SR, the TPC matrix $V$ at the BS, the AN covariance matrix $Z$ at the BS, and the phase shift matrix $\Phi$ at the IRS should be optimized jointly subject to the constraints of the maximum transmit power and unit modulus of phase shifts. Hence, we formulate the SRM problem as

\[
\begin{align*}
\max_{V, \Phi, Z} & \quad C_{AN}(V, \Phi, Z) \\
\text{s.t.} & \quad \text{Tr}(VV^H + Z) \leq P_T, \\
& \quad Z \succeq 0, \\
& \quad |\phi_m| = 1, m = 1, \ldots, M,
\end{align*}
\]  

(9a) 

(9b) 

(9c) 

(9d)

where $\phi_m = e^{j \theta_m}$, $\Phi = \text{diag}\{ \phi_1, \ldots, \phi_m, \ldots, \phi_M \}$ and $P_T$ is the maximum transmit power limit.

By variable substitution $Z = V_EV_E^H$, where $V_E \in \mathbb{C}^{N_T \times N_T}$, Problem (9) is equivalent to

\[
\begin{align*}
\max_{V, V_E, \Phi} & \quad C_{AN}(V, V_E, \Phi) \\
\text{s.t.} & \quad \text{Tr}(VV^H + V_EV_E^H) \leq P_T, \\
& \quad |\phi_m| = 1, m = 1, \ldots, M,
\end{align*}
\]  

(10a) 

(10b) 

(10c)

where $C_{AN}(V, V_E, \Phi) = C_{AN}(V, \Phi, Z) = [R_I(V, V_E, \Phi) - R_E(V, V_E, \Phi)]^+$ by substituting $Z = V_EV_E^H$ into (8). There are two difficulties in solving Problem (10), which lies in the OF of (10a) and the constraints of (10c). In (10a), the expression of OF is hard to tackle, and the variables of the TPC matrix $V$, $V_E$, and the phase shift matrix $\Phi$ are coupled with each other, which make Problem (10) difficult to solve. What’s more, the unit modulus constraint imposed on the phase shifts in (10c) compound the difficulty. In view of this, we divide and conquer Problem (10) by providing a low-complexity algorithm.

III. A LOW-COMPLEXITY ALGORITHM OF BCD-MM

Firstly, the OF of problem (10) is reformulated into a more tacklable expression equivalently. Then, the BCD-MM method is proposed for optimizing the TPC matrix $V$, $V_E$, and the phase shift matrix $\Phi$ alternatively.
A. Reformulation of the Original Problem

Firstly, the achievable SR $C_{AN}(V, V_E, \Phi)$ can be simplified as

$$C_{AN}(V, V_E, \Phi) = \log \left| I_{N_J} + \hat{H}_I V V^H \hat{H}_I^H (\hat{H}_I V E V_E^H \hat{H}_E^H + \sigma_E^2 I_{N_J})^{-1} \right|$$

$$+ \log \left| I_{N_E} + \hat{H}_E V E V^H \hat{H}_E^H (\sigma_E^2 I_{N_E})^{-1} \right|$$

$$- \log \left| I_{N_E} + \sigma_E^{-2} \hat{H}_E (V V^H + V_E V_E^H) \hat{H}_E^H \right|. \quad (11)$$

Secondly, we have

$$f_1 = \max_{U_I, W_I > 0} \log |W_I| - \text{Tr}(W_I E_I(U_I, V, V_E)) + d, \quad (12)$$

where $U_I \in \mathbb{C}^{N_J \times d}$ and $W_I \in \mathbb{C}^{d \times d}$ are the introduced auxiliary variables. The optimal $U_I^*, W_I^*$ to achieve the maximum value of (12) is given by

$$U_I^* = (\hat{H}_I V E V_E^H \hat{H}_I^H + \sigma_E^2 I_{N_J} + \hat{H}_I V V^H \hat{H}_I^H)^{-1} \hat{H}_I V, \quad (13)$$

$$W_I^* = [E_I^*(U_I^*, V, V_E)]^{-1}, \quad (14)$$

where $E_I^*$ is obtained by plugging the expression of $U_I^*$ into $E_I(U_I, V, V_E)$ as

$$E_I^*(U_I^*, V, V_E) = (U_I^{*H} \hat{H}_I V - I_d)(U_I^{*H} \hat{H}_I V - I_d)^H + U_I^{*H}(\hat{H}_I V V^H \hat{H}_I^H + \sigma_E^2 I_{N_J}) U_I^*. \quad (15)$$

We have

$$f_2 = \max_{U_E, W_E > 0} \log |W_E| - \text{Tr}(W_E E_E(U_E, V_E)) + N_t, \quad (16)$$

where $U_E \in \mathbb{C}^{N_E \times N_T}$ and $W_E \in \mathbb{C}^{N_T \times N_T}$ are the introduced auxiliary variables. The optimal $U_E^*$, $W_E^*$ to achieve the maximum value of (16) is given by

$$U_E^* = (\sigma_E^2 I_{N_E} + \hat{H}_E V E V_E^H \hat{H}_E^H)^{-1} \hat{H}_E V_E, \quad (17)$$

$$W_E^* = [E_E^*(U_E^*, V_E)]^{-1}, \quad (18)$$

where $E_E^*$ is obtained by plugging the expression of $U_E^*$ into $E_E(U_E, V_E)$ as

$$E_E^*(U_E^*, V_E) = (U_E^{*H} \hat{H}_E V_E - I_{N_T})(U_E^{*H} \hat{H}_E V_E - I_{N_T})^H + U_E^{*H}(\sigma_E^2 I_{N_E}) U_E^*. \quad (19)$$
We have

\[ f_3 = \max_{W_X > 0} \log |W_X| - \text{Tr}(W_X E_X(V, V_E)) + N_E, \]  

(20)

where \( W_X \in \mathbb{C}^{N_E \times N_E} \) are the introduced auxiliary variable. The optimal \( W_X^* \) to achieve the maximum value of (20) is given by

\[ W_X^* = [E_X(V, V_E)]^{-1}, \]  

(21)

where

\[ E_X(V, V_E) = I_{N_E} + \sigma_E^{-2} \hat{H}_E(VV^H + V_EV_E^H)\hat{H}_E^H. \]  

(22)

Finally, by substituting (12), (16), (20) into (11), Problem (10) is equivalently reformulated as

\[
\begin{align*}
\min_{V, V_E, \Phi} & - \text{Tr}(W_I V^H \hat{H}_I^H U_I) - \text{Tr}(W_I U_I^H \hat{H}_I V) + \text{Tr}(V^H H_V V) \\
& - \text{Tr}(W_E V_E^H \hat{H}_E^H U_E) - \text{Tr}(W_E U_E^H \hat{H}_E V_E) + \text{Tr}(V_E^H H_{V_E} V_E) \\
\text{s.t.} & \quad \text{Tr}(VV^H + V_EV_E^H) \leq P_T, \\
& \quad |\phi_m| = 1, m = 1, \cdots, M,
\end{align*}
\]

(23a)

\[
\begin{align*}
\min_{V, V_E} & - \text{Tr}(W_I V^H \hat{H}_I^H U_I) - \text{Tr}(W_I U_I^H \hat{H}_I V) + \text{Tr}(V^H H_V V) \\
& - \text{Tr}(W_E V_E^H \hat{H}_E^H U_E) - \text{Tr}(W_E U_E^H \hat{H}_E V_E) + \text{Tr}(V_E^H H_{V_E} V_E) \\
\text{s.t.} & \quad \text{Tr}(VV^H + V_EV_E^H) \leq P_T.
\end{align*}
\]

(26a)

It is obvious that Problem (23) is much easier to tackle than Problem (10) due to the convex quadratic OF in (23a). Now, we devote to solve Problem (23) equivalently instead of Problem (10), and the matrices \( V, V_E, \) and phase shift matrix \( \Phi \) will be optimized.

**B. Optimizing the Matrices \( V \) and \( V_E \)**

In this subsection, the TPC matrix \( V \) and matrix \( V_E \) are optimized by fixing \( \Phi \). Specifically, the unit modulus constraint on the phase shifts \( \Phi \) is removed, and the updated optimization problem reduced from Problem (23) is given by

\[
\begin{align*}
\min_{V, V_E} & - \text{Tr}(W_I V^H \hat{H}_I^H U_I) - \text{Tr}(W_I U_I^H \hat{H}_I V) + \text{Tr}(V^H H_V V) \\
& - \text{Tr}(W_E V_E^H \hat{H}_E^H U_E) - \text{Tr}(W_E U_E^H \hat{H}_E V_E) + \text{Tr}(V_E^H H_{V_E} V_E) \\
\text{s.t.} & \quad \text{Tr}(VV^H + V_EV_E^H) \leq P_T.
\end{align*}
\]

(26b)
The above problem is a convex QCQP problem, and the standard optimization packages, such as CVX [43] can be exploited to solve it. However, the calculation burden is heavy. To reduce the complexity, the near-optimal closed form expressions of the TPC matrix and noise covariance matrix are provided by applying the Lagrangian multiplier method.

Since Problem (26) is a convex problem, the Slater’s condition is satisfied, where the duality gap between Problem (26) and its dual problem is zero. Thus, Problem (26) can be solved by addressing its dual problem if the dual problem is easier. For this purpose, by introducing Lagrange multiplier $\lambda$ to combine the the constraint and OF of Problem (26), the Lagrangian function of Problem (26) is obtained as

$$L(V, V_E, \lambda) \triangleq -\text{Tr} \left( W_I V_H \hat{H}_I U_I \right) - \text{Tr} \left( W_I U_I^H \hat{H}_I V \right) + \text{Tr} \left( V_H H_I V \right) - \text{Tr} \left( W_E V_E^H \hat{H}_E^H U_E \right)$$
$$- \text{Tr} \left( W_E U_E^H \hat{H}_E V_E \right) + \text{Tr} \left( V_E^H H_E V_E \right) + \lambda \left[ \text{Tr} \left( V V_H + V_E V_E^H \right) - P_T \right]$$
$$= -\text{Tr} \left( W_I V_H \hat{H}_I^H U_I \right) - \text{Tr} \left( W_I U_I^H \hat{H}_I V \right) + \text{Tr} \left[ V_H (H_I + \lambda I) V \right]$$
$$- \text{Tr} \left( W_E V_E^H \hat{H}_E^H U_E \right) - \text{Tr} \left( W_E U_E^H \hat{H}_E V_E \right) + \text{Tr} \left[ V_E^H (H_{VE} + \lambda I) V_E \right] - \lambda P_T. \tag{27}$$

Then the dual problem of Problem (26) is

$$\max_{\lambda} h(\lambda) \tag{28a}$$
$$\text{s.t.} \quad \lambda \geq 0, \tag{28b}$$

where $h(\lambda)$ is the dual function given by

$$h(\lambda) \triangleq \min_{V, V_E} L(V, V_E, \lambda). \tag{29}$$

Note that Problem (29) is a convex quadratic optimization problem with no constraint, which can be solved in closed form. The optimal solution $V^*, V_E^*$ for Problem (29) is

$$[V^*, V_E^*] = \arg \min_{V, V_E} L(V, V_E, \lambda). \tag{30}$$

By setting the first-order derivative of $L(V, V_E, \lambda)$ w.r.t. $V$ to zero, we can obtain the optimal solution of $V$ as follows:

$$\frac{\partial L(V, V_E, \lambda)}{\partial V} = 0, \tag{31a}$$
$$\frac{\partial L(V, V_E, \lambda)}{\partial V_E} = 0. \tag{31b}$$
The left hand side of Equation (31a) can be expanded as
\[
\frac{\partial \mathcal{L}(V, V_E, \lambda)}{\partial V} = \frac{\partial \text{Tr} \left[ V^H (H_V + \lambda I) V \right]}{\partial V} - \left( W_I U_I^H \hat{H}_I \right)^H - \left( \hat{H}_I^H U_I W_I \right)
\]
\[
= (H_V + \lambda I) V + (H_V + \lambda I)^H V - \left( W_I U_I^H \hat{H}_I \right)^H - \left( \hat{H}_I^H U_I W_I \right)
\]
\[
= 2 (H_V + \lambda I) V - 2 \left( \hat{H}_I^H U_I W_I \right).
\]
(32)

The equation (31a) becomes
\[
(H_V + \lambda I) V = \left( \hat{H}_I^H U_I W_I \right).
\]
(33)

Then the optimal solution \( V^* \) for Problem (30) is
\[
V^* = (H_V + \lambda I)^\dagger \left( \hat{H}_I^H U_I W_I \right)
\]
\[
\triangleq \Theta_V(\lambda) \left( \hat{H}_I^H U_I W_I \right).
\]
(34)

Similarly, we solve Problem (30) by setting the first-order derivative of \( \mathcal{L}(V, V_E, \lambda) \) w.r.t. \( V_E \) to zero, which becomes
\[
2 (H_{VE} + \lambda I) V_E - 2 \hat{H}_E^H U_E W_E^H = 0.
\]
(35)

Then the optimal solution \( V_E^* \) for Problem (30) is
\[
V_E^* = (H_{VE} + \lambda I)^\dagger \hat{H}_E^H U_E W_E^H
\]
\[
\triangleq \Theta_{VE}(\lambda) \hat{H}_E^H U_E W_E^H.
\]
(36)

Once the optimal solution \( \lambda^* \) for Problem (28) is found, the final optimal \( V^*, V_E^* \) can be obtained. The value of \( \lambda^* \) should be chosen in order to guarantee the complementary slackness condition as follows:
\[
\lambda [\text{Tr}(V^* V^{*H} + V_E^* V_E^{*H}) - P_T] = 0.
\]
(37)

We define
\[
P(\lambda) \triangleq \text{Tr}(V^* V^{*H} + V_E^* V_E^{*H}) = \text{Tr}(V^* V^{*H}) + \text{Tr}(V_E^* V_E^{*H}),
\]
(38)

where
\[
\text{Tr} \left( V^* V^{*H} \right) = \text{Tr} \left( \Theta_V(\lambda) \hat{H}_I^H U_I W_I^H (\hat{H}_I^H U_I W_I^H)^H \Theta_V^H(\lambda) \right)
\]
\[
= \text{Tr} \left( \Theta_V^H(\lambda) \Theta_V(\lambda) (\hat{H}_I^H U_I W_I^H)(\hat{H}_I^H U_I W_I^H)^H \right),
\]
(39)
and

\[
\text{Tr} \left( V_E^H V_E^* \right) = \text{Tr} \left( \Theta_{VE}(\lambda) (\hat{H}_E^H U_E W_E^H)(\hat{H}_E^H U_E W_E^H)^H \Theta_{VE}^H(\lambda) \right) \\
= \text{Tr} \left( \Theta_{VE}^H(\lambda) \Theta_{VE}(\lambda) (\hat{H}_E^H U_E W_E^H)(\hat{H}_E^H U_E W_E^H)^H \right) . 
\] (40)

Then \( P(\lambda) \) becomes

\[
P(\lambda) = \text{Tr} \left( \Theta_E^n (\hat{H}_I^H U_I W_I^H)(\hat{H}_I^H U_I W_I^H)^H \right) + \text{Tr} \left( \Theta_{VE}^n (\hat{H}_E^H U_E W_E^H)(\hat{H}_E^H U_E W_E^H)^H \right) , 
\] (41)

where

\[
\Theta_E^n = \Theta_{VE}^H(\lambda) \Theta_{VE}(\lambda) = (H_V + \lambda I)^\dagger (H_V + \lambda I)^\dagger , 
\] (42)

and

\[
\Theta_{VE}^n = \Theta_{VE}^H(\lambda) \Theta_{VE}(\lambda) = (H_{VE} + \lambda I)^\dagger (H_{VE} + \lambda I)^\dagger . 
\] (43)

To find the optimal \( \lambda^* \geq 0 \), we first check whether \( \lambda = 0 \) is the optimal solution or not. If

\[
P(0) = \text{Tr} \left( V^{*H}(0) V^*(0) \right) + \text{Tr} \left( V_E^{*H}(0) V_E^*(0) \right) \leq P_T , 
\] (44)

then the optimal solutions are given by \( V^* = V(0) \) and \( V_E^* = V_E(0) \). Otherwise, the optimal \( \lambda^* > 0 \) is the solution of the equation \( P(\lambda) = 0 \).

It is ready to verify that \( H_V \) and \( H_{VE} \) is a positive definite matrix. Let us define the rank of \( H_V \) and \( H_{VE} \) as \( r_V = \text{rank}(H_V) \leq N_T \) and \( r_{VE} = \text{rank}(H_{VE}) \leq N_T \) respectively. By decomposing \( H_V \) and \( H_{VE} \) by using the singular value decomposition (SVD), we have

\[
H_V = [P_{V,1}, P_{V,2}] \Sigma_V [P_{V,1}, P_{V,2}]^H , \\
H_{VE} = [P_{VE,1}, P_{VE,2}] \Sigma_{VE} [P_{VE,1}, P_{VE,2}]^H ,
\] (45)

where \( P_{V,1} \) comprises the first \( r_V \) singular vectors associate with the \( r_V \) positive eigenvalues of \( H_V \), and \( P_{V,2} \) includes the last \( N_T - r_V \) singular vectors associate with the \( N_T - r_V \) zero-valued eigenvalues of \( H_V \). \( \Sigma_V = \text{diag} \left\{ \Sigma_{V,1}, 0_{(N_T-r_V) \times (N_T-r_V)} \right\} \) with \( \Sigma_{V,1} \) representing the diagonal submatrix collecting the first \( r_V \) positive eigenvalues. Similarly, the first \( r_{VE} \) singular vectors corresponding to the \( r_{VE} \) positive eigenvalues of \( H_{VE} \) are contained in \( P_{VE,1} \), while the last \( N_T - r_{VE} \) singular vectors corresponding to the \( N_T - r_{VE} \) zero-valued eigenvalues of \( H_{VE} \) are held in \( P_{VE,2} \). \( \Sigma_{VE} = \text{diag} \left\{ \Sigma_{VE,1}, 0_{(N_T-r_{VE}) \times (N_T-r_{VE})} \right\} \) is a diagonal matrix with \( \Sigma_{VE,1} \) representing
the diagonal submatrix gathering the first $r_{VE}$ positive eigenvalues. By defining $P_V \triangleq [P_{V,1}, P_{V,2}]$ and $P_{VE} \triangleq [P_{VE,1}, P_{VE,2}]$, and substituting (45) into (42) and (43), $P(\lambda)$ becomes

$$
P(\lambda) = \text{Tr} \left( [P_V \Sigma_V P_V^H + \lambda I]^{-1} [P_V \Sigma_V P_V^H + \lambda I]^{-1} (\hat{H}_I^H U_I W_I^H)(\hat{H}_I^H U_I W_I^H)^H \right)$$

$$+ \text{Tr} \left( [P_{VE} \Sigma_{VE} P_{VE}^H + \lambda I]^{-1} [P_{VE} \Sigma_{VE} P_{VE}^H + \lambda I]^{-1} (\hat{H}_E^H U_E W_E^H)(\hat{H}_E^H U_E W_E^H)^H \right)$$

$$= \text{Tr} \left( [P_V (\Sigma_V + \lambda I)^{-2} P_V^H] (\hat{H}_I^H U_I W_I^H)(\hat{H}_I^H U_I W_I^H)^H \right)$$

$$+ \text{Tr} \left( [P_{VE} (\Sigma_{VE} + \lambda I)^{-2} P_{VE}^H] (\hat{H}_E^H U_E W_E^H)(\hat{H}_E^H U_E W_E^H)^H \right)$$

$$= \text{Tr} \left( [\Sigma_V + \lambda I]^{-2} Z_V \right) + \text{Tr} \left( [\Sigma_{VE} + \lambda I]^{-2} Z_{VE} \right)$$

$$= \sum_{i=1}^{r_V} \left[ \frac{[Z_V]_{i,i}}{(\Sigma_V)_{i,i} + \lambda} \right]^2 + \sum_{i=1}^{r_{VE}} \left[ \frac{[Z_{VE}]_{i,i}}{(\Sigma_{VE})_{i,i} + \lambda} \right]^2 + \sum_{i=r_V+1}^{N_T} \left[ \frac{[Z_V]_{i,i}}{(\lambda)^2} \right] + \sum_{i=r_{VE}+1}^{N_T} \left[ \frac{[Z_{VE}]_{i,i}}{(\lambda)^2} \right],$$

(46)

where $Z_V = P_V^H (\hat{H}_I^H U_I W_I^H)(\hat{H}_I^H U_I W_I^H)^H P_V$ and $Z_{VE} = P_{VE}^H (\hat{H}_E^H U_E W_E^H)(\hat{H}_E^H U_E W_E^H)^H P_{VE}$. $[Z_V]_{i,i}$, $[Z_{VE}]_{i,i}$, $[\Sigma_V]_{i,i}$, and $[\Sigma_{VE}]_{i,i}$ represent the $i$th diagonal element of matrices $Z_V$, $Z_{VE}$, $\Sigma_V$, and $\Sigma_{VE}$, respectively. $P(\lambda)$ can be verified from (46) to be a monotonically decreasing function.

Then, the optimal $\lambda^*$ can be obtained by solving the following equation,

$$
\sum_{i=1}^{r_V} \left[ \frac{[Z_V]_{i,i}}{(\Sigma_V)_{i,i} + \lambda} \right]^2 + \sum_{i=1}^{r_{VE}} \left[ \frac{[Z_{VE}]_{i,i}}{(\Sigma_{VE})_{i,i} + \lambda} \right]^2 + \sum_{i=r_V+1}^{N_T} \left[ \frac{[Z_V]_{i,i}}{(\lambda)^2} \right] + \sum_{i=r_{VE}+1}^{N_T} \left[ \frac{[Z_{VE}]_{i,i}}{(\lambda)^2} \right] = P_T.
$$

(47)

To solve it, the bisection search method is utilized. Since $P(\infty) = 0$, the solution to Equation (47) must exist. The lower bound of $\lambda^*$ is a positive value approaching zero, while the upper bound of $\lambda^*$ is given by

$$
\lambda^* < \sqrt{\frac{\sum_{i=1}^{N_T} [Z_V]_{i,i} + \sum_{i=1}^{N_T} [Z_{VE}]_{i,i}}{P_T}} \triangleq \lambda^{ub}.
$$

(48)

which can be proved as

$$
p(\lambda^{ub}) = \sum_{i=1}^{N_T} \frac{[Z_V]_{i,i}}{(\Sigma_V)_{i,i} + \lambda^{ub})^2} + \sum_{i=1}^{N_T} \frac{[Z_{VE}]_{i,i}}{(\Sigma_{VE})_{i,i} + \lambda^{ub})^2} < \sum_{i=1}^{N_T} \frac{[Z_V]_{i,i}}{(\lambda^{ub})^2} + \sum_{i=1}^{N_T} \frac{[Z_{VE}]_{i,i}}{(\lambda^{ub})^2} = P_T.
$$

(49)

When the optimal $\lambda^*$ is found, the optimal matrix can be obtained by $V^* = V^*(\lambda^*)$ and $V_E^* = V_E^*(\lambda^*)$ in (34) and (36).
C. Optimizing the Phase Shifts $\Phi$

In this subsection, the phase shift matrix $\Phi$ is optimized by fixing $V$ and $V_E$. The transmit power constraint in Problem (23) is only related with $V$ and $V_E$, thus is removed. Then, the optimization problem for $\Phi$ reduced from Problem (23) is formulated as

$$
\min_{\Phi} g_0(\Phi) \triangleq -\text{Tr}(W_i V^H \hat{H}_i^H U_i) - \text{Tr}(W_i U_i^H \hat{H}_i V) + \text{Tr}(V^H H V)
- \text{Tr}(W E V_E^H \hat{H}_E^H U_E) - \text{Tr}(W E U_E^H \hat{H}_E V_E) + \text{Tr}(V_E^H H V_E V_E)
$$
\hspace{1cm} (50a)

s.t. $|\phi_m| = 1, m = 1, \cdots, M.$ \hspace{1cm} (50b)

By exploiting the matrix properties in [44, Eq. (1.10.6)], the trace operators can be removed, and the third and fourth terms in (51) become as

$$
\begin{align*}
&\text{Tr}(\Phi^H B_{VE} \Phi C_{VE}) = \phi^H (B_{VE} \odot C_{VE}^T) \phi, \\
&\text{Tr}(\Phi^H B_V \Phi C_V) = \phi^H (B_V \odot C_V^T) \phi,
\end{align*}
$$
\hspace{1cm} (53a-b)

where $\phi \triangleq [e^{j\theta_1}, \cdots, e^{j\theta_m}, \cdots, e^{j\theta_M}]^T$ is a vector holding the diagonal elements of $\Phi$.

Similarly, the trace operators can be removed for the first and second terms in (51), which become as

$$
\text{Tr}(\Phi^H D^H) = d^H(\phi^*), \text{Tr}(\Phi D) = \phi^T d,
$$
\hspace{1cm} (54)

where $d = [D]_{1,1}, \cdots, [D]_{M,M}]^T$ is a vector gathering the diagonal elements of matrix $D$. 

Hence, Problem (50) can be rewritten as

\[ \min_\theta \quad \phi^H \Xi \phi + \phi^T d + d^H (\phi^*) \]  
\[ \text{s.t.} \quad 0 \leq \theta_m \leq 2\pi, \ m = 1, \cdots, M, \]  

where \( \Xi = B_{VE} \odot C_{VE}^T + B_V \odot C_V^T \). \( \Xi \) is a semidefinite matrix, because it is a sum of two semidefinite matrices, both of which are Hadamard products of two semidefinite matrices. It is observed from (52b), (52c), (52d) and (52e) that \( B_{VE}, C_{VE}^T, B_V \) and \( C_V^T \) are semidefinite matrices. Then, the Hadamard products of \( B_{VE} \odot C_{VE}^T \) and \( B_V \odot C_V^T \) are semidefinite according to the Property (9) on Page 104 of [44].

Since \( \phi = [\phi_1, \cdots, \phi_M]^T \), and \( \phi_m = e^{j\theta_m}, \forall m \), Problem (55) can be further simplified as

\[ \min_\phi \quad f(\phi) \overset{\Delta}{=} \phi^H \Xi \phi + 2 \text{Re} \{ \phi^H (d^*) \} \]  
\[ \text{s.t.} \quad |\phi_m| = 1, \ m = 1, \cdots, M. \]  

The Problem (56) can be solved efficiently by the MM algorithm as [21]. Details are omitted for simplicity.

D. Overall Algorithm to Solve Problem (10)

To sum up, the detailed execution of the overall BCD-MM algorithm proposed for solving Problem (10) is provided in Algorithm 1. The MM algorithm is exploited for solving the optimal phase shifts \( \Phi^{(n+1)} \) of Problem (56) in Step 5. The iteration process in MM algorithm ensures that the OF value of Problem (56) decreases monotonically. Moreover, the BCD algorithm also guarantees that the OF value of Problem (23) monotonically decreases in each step and each iteration of Algorithm 1. Since the OF value in (23a) has a lower bound with the power limit, the convergence of Algorithm 1 is guaranteed.

Based on the algorithm description, the complexity analysis of the proposed BCD-MM algorithm is performed. In Step 3, computing the decoding matrices \( U_{I}^{(n)} \) and \( U_{E}^{(n)} \) costs the complexity of \( O(N_T^3) + O(N_E^3) \), while calculating the auxiliary matrices \( W_{I}^{(n)}, W_{E}^{(n)}, \) and \( W_{X}^{(n)} \) consumes the complexity of \( O(d^3) + O(N_T^3) + O(N_E^3) \). The complexity of calculating the TPC matrix \( V^{(n+1)} \) and AN covariance matrix \( V_{E}^{(n+1)} \) in Step 4 can be analyzed according to the specific process of Lagrangian multiplier method based on the fact that the complexity of computing product \( XY \) of complex matrices \( X \in \mathbb{C}^{m \times n} \) and \( Y \in \mathbb{C}^{n \times p} \) is \( O(mnp) \). By assuming that
Algorithm 1 BCD-MM Algorithm

1: Parameter Setting. Set the maximum number of iterations $n_{\text{max}}$ and the first iterative number $n = 1$; Give the error tolerance $\varepsilon$.
2: Variables Initialization. Initialize the variables $V^{(1)}$, $V_E^{(1)}$ and $\Phi^{(1)}$ in the feasible region; Compute the OF value of Problem (10) as $\text{OF}(V^{(1)}, V_E^{(1)})$;
3: Auxiliary Variables Calculation. Given $V^{(n)}$, $V_E^{(n)}$, $\Phi^{(n)}$, compute the optimal matrices $U_I^{(n)}$, $W_I^{(n)}$, $U_E^{(n)}$, $W_E^{(n)}$, $W_X^{(n)}$ according to (13), (14), (17), (18), (21) respectively;
4: Matrices Optimization. Given $U_I^{(n)}$, $W_I^{(n)}$, $U_E^{(n)}$, $W_E^{(n)}$, $W_X^{(n)}$, solve the optimal PTC matrix $V^{(n+1)}$ and AN covariance matrix $V_E^{(n+1)}$ of Problem (30) with the Lagrangian multiplier method;
5: Phase Shifts Optimization. Given $U_I^{(n)}$, $W_I^{(n)}$, $U_E^{(n)}$, $W_E^{(n)}$, $W_X^{(n)}$, $V^{(n)}$, $V_E^{(n)}$, solve the optimal phase shifts $\Phi^{(n+1)}$ of Problem (56) with the MM algorithm;
6: Termination Check. If $|\text{OF}(V^{(n+1)}, V_E^{(n+1)}, \Phi^{(n+1)}) - \text{OF}(V^{(n)}, V_E^{(n)}, \Phi^{(n)})|/\text{OF}(V^{(n+1)}, V_E^{(n+1)}, \Phi^{(n+1)}) < \varepsilon$ or $n \geq n_{\text{max}}$, terminate. Otherwise, update $n \leftarrow n + 1$ and jump to step 2.

$N_T > N_I$ (or $N_E > d$, the complexity of computing the matrices $\{H_V, H_{VE}\}$ in (24) and (25) is $\mathcal{O}(N_T^3) + \mathcal{O}(2N_T^2d) + \mathcal{O}(2N_T^2N_E)$; while the complexity of computing $V^*$, $V_E^*$ in (34) and (36) is $\mathcal{O}(2N_T^3)$. The SVD decomposition of $\{H_V, H_{VE}\}$ require the computation complexity of $\mathcal{O}(2N_T^3)$, meanwhile calculating $\{Z_V\}$ and $\{Z_{VE}\}$ require the complexity of $\mathcal{O}(N_T^2N_I) + \mathcal{O}(2N_T^3)$. The complexity of finding the Lagrangian multipliers $\{\lambda\}$ is negligible. Thus, the overall complexity for $V^{(n+1)}$, $V_E^{(n+1)}$ is about $\mathcal{O}(\max\{2N_T^3, 2N_T^2N_E\})$. In step 5, obtaining optimal $\Phi^{(n+1)}$ by the MM algorithm need a complexity of $C_{MM} = \mathcal{O}(M^3 + T_{MM}M^2)$, where $T_{MM}$ is the iteration number for convergence. Based on the complexity required in Step 3, 4 and 5, the overall complexity $C_{\text{BCD-MM}}$ of the BCD-MM algorithm can be evaluated by

$$C_{\text{BCD-MM}} = \mathcal{O}(\max\{2N_T^3, 2N_T^2N_E, C_{MM}\}).$$  

IV. SIMULATION RESULTS

In this section, numerical simulations are carried out to evaluate the assistance of the IRS on the AN-aided MIMO secure communication system. We focus on the scenario of the standard three-terminal MIMO Gaussian wiretap channel shown in Fig. 2, where there are one BS, one legitimate IR and one Eve, all with multiple antennas. The distance from the BS to the IRS is
Fig. 2. The three-terminal MIMO communication scenario in simulation.

$d_{BR} = 50$ m. We assume that the line connecting the IR and Eve is parallel to the line connecting the BS and the IRS, and that the vertical distance between them is $d_v = 2$ m. The large-scale path loss is modeled as $PL = PL_0 - 10\alpha \log_{10} \left( \frac{d}{d_0} \right)$, where $PL_0$ is the path loss at the reference distance $d_0 = 1$ m, $\alpha$ is the path loss exponent, $d$ is the link distance. In our simulations, we set $PL_0 = -30$ dB. The path loss exponents of the links from BS to Eve, from BS to IR, from IRS to Eve and from IRS to IR are $\alpha_{BE} = 3.5$, $\alpha_{BI} = 3.5$, $\alpha_{RE} = 2.5$ and $\alpha_{RI} = 2.5$ respectively. The path-loss exponents of the link from BS to IRS is set to be $\alpha_{BR} = 2.2$, which means that the IRS is well-located, and the path loss is negligible in this link. For the small-scale fading, all channels are modeled as Rayleigh fading.

If not specified, the simulation parameters are set as follows. The IR’s noise power and the Eve’s noise power are $\sigma_I^2 = -75$ dBm and $\sigma_E^2 = -75$ dBm. The numbers of BS antennas, IR antennas, and Eve antennas are $N_T = 4$, $N_I = 2$, and $N_E = 2$ respectively. There are $d = 2$ data streams and $M = 50$ IRS reflection elements. The transmit power limit is $P_T = 15$ dBm, and the error tolerance is $\varepsilon = 10^{-6}$. The horizontal distance between the BS and the Eve is $d_{BE} = 44$ m. The horizontal distance between the BS and the IR is selected from $d_{BI} = [10 \text{ m}, 70 \text{ m}]$. The channels are realized 200 times independently to average the simulation results.

A. Convergence Analysis

The convergence performance of the proposed BCD-MM algorithm is investigated. The iterations of the BCD algorithm are termed as outer-layer iterations, while the iteration of the MM algorithm are termed as the inner-layer iterations. Fig. 3 shows three examples of convergence behaviour for $M = 10$, 20 and 40 phase shifts of IRS. In Fig. 3, the SR increases versus the iteration
number, and finally reaches a stable value. It is shown that the algorithm converges quickly, almost with 20 iterations, which demonstrate the efficiency of the proposed algorithm. Moreover, larger converged SR value is reached with a larger $M$, which means that better security can be obtained by using more IRS elements. However, more IRS elements brings heavier computation, which is demonstrated in Fig. 3 in the form of a slower convergence speed with more phase shifters.

![Fig. 3. Convergence behaviour of the BCD algorithm.](image)

Specifically, we evaluate the convergence performance of the MM algorithm used for solving the optimal IRS phase shifts. The inner-layer iterative process of the MM algorithm in the first iteration of the BCD algorithm is shown in Fig. 4. The SR value increases as the iteration number

![Fig. 4. Convergence behaviour of the MM algorithm.](image)
increases, and finally converges to a stable value. According with the convergency performance in the out-layer iteration, similar conclusions can be drawn for the inner-layer iteration, which is that higher converged SR value can be obtained with more phase shifts but at the cost of lower convergence speed. The reason for the lower convergence speed with larger $M$ value is that more optimization variables are introduced, which require more computation complexity.

B. Performance Evaluation

In this subsection, our proposed algorithm is evaluated by comparing the simulation results to two schemes of

1) **RandPhase**: The phase shifts of the IRS are randomly selected from $[0, 2\pi]$. In this scheme, the MM algorithm is skipped, and only the TPC matrix and AN noise covariance matrix are optimized.

2) **No-IRS**: Without the IRS, the channel matrices of IRS related links become zero matrices, which is $\mathbf{H}_{R,I} = 0$, $\mathbf{H}_{R,E} = 0$ and $\mathbf{G} = 0$. This scheme results a conventional AN-aided communication system, and only the TPC matrix and AN noise covariance matrix need to be optimized.

1) **Impact of Transmit Power**: To evaluate the impact of the transmit power limit $P_T$, the average SR versus the transmit power limit for various schemes are given in Fig. 5, which demonstrates that the achieved SRs of three schemes increase as the power limit $P_T$ increases. It is observed
that the BCD-MM algorithm significantly outperforms the other two benchmark schemes over the entire range of transmit power limits. By comparing the RandPhase scheme to the No-IRS scheme, we find that the RandPhase scheme is better than the No-IRS scheme for obtaining higher SR, and that the SR gap increases with the power limit $P_T$. The reason is that, for the RandPhase scheme, the IR is closer to the IRS than the Eve is, and more signal power from the IRS can be acquired by the IR than that by the Eve, while for the No-IRS scheme, the IR is further from the BS than the Eve is, and less signal power from the BS can be acquired by the IR than that by the Eve. This comparison result signifies that even the phase shifts of IRS is random, the IRS can enhance the system security. In comparison to the no-IRS scheme, the SR gain achieved by the proposed algorithm is very obvious, and increases greatly with the power limit $P_T$, which confirms the effectiveness and benefits of employing the IRS. By comparing the proposed scheme and the RandPhase scheme, we find that the security gain obtained for the proposed scheme is much greater than that for the RandPhase scheme. That's because the phase shifts of IRS are properly designed to enhance the signal received at the IR more constructively, and weaken the signal received at the Eve more destructively. This comparison result emphasizes that optimizing the phase shifts of IRS is important and necessary.

2) Impact of the Phase Shifts Number: The averaged SR performance of three schemes with various phase shifts number $M$ is shown in Fig. 6, which demonstrates that the proposed BCD-MM algorithm is significantly superior to the other two schemes. We observe that the SR achieved by
the BCD-MM scheme obviously increases with $M$, while the RandPhase scheme only shows slight improvement as $M$ increases, and the No-IRS scheme has very low SRs irrelative with $M$. Larger the element number $M$ of IRS is, more significant the performance gain obtained by the proposed algorithm is. For example, when $M$ is small as $M = 10$, the SR gain of the BCD-MM relative to the No-IRS is only 1.3 bit/s/Hz, while this SR gain becomes 9.5 bit/s/Hz when $M$ increases to $M = 100$. The performance gain for the proposed algorithm originates from two perspectives. On the one hand, a higher array gain can be obtained by increasing $M$, since more signal power can be received at the IRS with larger $M$. On the other hand, a higher reflecting beamforming gain can be obtained by increasing $M$, which means that the sum of coherently adding the reflected signals at the IRS elements increases with $M$ by appropriately designing the phase shifts. However, only the array gain can be exploited by the RandPhase scheme, thus the SRs for it increase very slowly, and remain at much lower values than those for the proposed algorithm. These results further confirm that more security improvement can be archived by using a large IRS with more reflect elements and optimizing the phase shifts properly, however there may bring the computation complexity problem.

3) Impact of the relative location of IRS: Fig. 7 illustrates the achieved SRs for three schemes with various BS-IR horizontal distance $d_{BI}$, where the BS-Eve distance is fixed to be $d_{BE} = 44$ m. It is observed that the proposed BCD-MM algorithm is the best among the three schemes for obtaining the highest SR value. When the IR moves far away from the BS, the SRs decrease for the
three schemes, however, the SRs achieved for the RandPhase and proposed BCD-MM algorithm increase greatly when the IR approaches the IRS. The achieved SRs at different BS-IR distances of the RandPhase scheme and the no-IRS scheme are almost the same, except for $d_{BI} \in [40m, 50m]$, in which case the IRS brings prominent security enhancement when IR becomes close to it even with random IRS phase shifts. For other BS-IR distances where the IR is far from the IRS, the SRs of RandPhase scheme are similar with those of the No-IRS scheme due to the not fully explored potential of IRS. By optimizing the phase shifts of IRS, the SRs are enhanced at different BS-IRS distances. And the SR gain of the proposed BCD-MM algorithm over the RandPhase scheme increases when the IR moves close to the IRS ($d_{BI} \in [40m, 50m]$). This signifies that as long as the IRS is deployed close to the IR, significant security enhancement can be achieved by the IRS in an AN-aided MIMO communication system. Moreover, it is highly recommended that the IRS phase shifts should be optimized to prevent the system security degrading into the level of No-IRS scheme.

4) Impact of the Path Loss Exponent of IRS-related Links: In the above simulations, the path loss exponents of the IRS-related links (including the BS-IRS link, IRS-IR link and IRS-Eve link) are set to be low by assuming that the IRS is properly located to obtain clean channels without heavy blockage. Practically, such kind of settings may not always be sensible due to real-field environment. Thus, it is necessary to investigate the security gain brought by the IRS and our proposed algorithm with higher value of IRS-related path loss exponents. For the sake of analysis, we assume the path-loss exponents of the links from BS to IRS, from IRS to IR and from IRS to Eve are the same as $\alpha_{BR} = \alpha_{RI} = \alpha_{RE} \triangleq \alpha_{IRS}$. Then, the achieved SR versus the path-loss exponent $\alpha_{IRS}$ of IRS-related links are shown in Fig. 8, which demonstrates that the SR obtained by the BCD-MM algorithm decreases as $\alpha_{IRS}$ increases, and finally drops to the same SR value which is achieved by the RandPhase and No-IRS schemes. The reason is that larger $\alpha_{IRS}$ means more severe signal attenuation in the IRS-related links, and more weakened signal received and reflected at the IRS. On the contrary, the performance gains brought by our proposed algorithm over the RandPhase and No-IRS schemes is significant with a small $\alpha_{IRS}$. Specifically, for $\alpha_{IRS} = 2$ (almost ideal channels), the security gain is up to 9.6 bit/s/Hz over the No-IRS scheme, and 6.8 bit/s/Hz over the RandPhase scheme. Therefore, the security gain of IRS-assisted systems depends on the channel conditions of the IRS-related links. This suggests that it is much preferred to deploy the IRS with fewer obstacles, in which case, the performance gain brought by the IRS can be explored
thoroughly. Fig. 8 also shows that when $\alpha_{\text{IRS}}$ is small, the RandPhase scheme can obtain security gain over the No-IRS scheme, but this security gain decreases to zero when $\alpha_{\text{IRS}}$ becomes large. However, the SR gain of the RandPhase scheme over the No-IRS scheme is almost negligible in comparison to the SR gain of the proposed scheme over the No-IRS scheme, which demonstrates that the necessity of jointly optimizing the TPC matrix, AN covariance matrix, and the phase shifts at the IRS.

![Graph showing Achievable SR versus path loss exponent](image)

Fig. 8. Achievable SR versus the path loss exponent of IRS-related links.

V. CONCLUSIONS

In this paper, we propose to enhance the security of AN-aided MIMO secure communication systems by exploiting an IRS. With the assist of IRS, the signal received at the legitimate IR can be enhanced while the signal received at the Eve can be weakened. To exploit the IRS sufficiently, we formulate a SRM problem by jointly optimizing the TPC matrix at the BS, the covariance matrix of AN and phase shifts at the IRS with the constraints of transmit power limit and unit-modulus of phase shifts. To solve this non-convex problem, we propose to use the BCD algorithm to decouple the optimization variables, and optimize them iteratively. The optimal TPC matrix and AN covariance matrix were obtained in closed form by the Lagrange multiplier method, and the phase shifts at the IRS were obtained in closed form by an efficient MM algorithm. Various simulations validated that significant security gains can be achieved by the proposed algorithm with IRS. Furthermore, useful suggestions for choosing and deploying the IRS are provided.
The derivation in (62) can be used for the second and third parts of (61). By defining \( V_X = (V^HV + V_EV_E^H) \) and \( M_I = U_IW_IW_I^H \), the first part of (61) can be derived as

\[
\begin{align*}
\text{Tr} \left[ \hat{H}_I (V^HV + V_EV_E^H) \hat{H}_I^H \right] &= \text{Tr} \left[ \hat{H}_IW_I^H \right] \\
&= \text{Tr} \left[ (H_{b,I} + H_{R,I}\Phi G) V_X \left( H_{b,I}^H + G^H\Phi^H H_{R,I}^H \right) M_I \right] \\
&= \text{Tr} \left[ (H_{b,I}V_XH_{b,I}^H + H_{b,I}V_XG^H\Phi^H H_{R,I}^H + H_{R,I}\Phi GV_XH_{b,I}^H + H_{R,I}\Phi GV_XG^H\Phi^H H_{R,I}^H) M_I \right] \\
&= \text{Tr} \left[ H_{b,I}V_XH_{b,I}^H M_I + H_{b,I}V_XG^H\Phi^H H_{R,I}^H M_I + H_{R,I}\Phi GV_XH_{b,I}^H M_I \right. \\
&\quad + \left. H_{R,I}\Phi GV_XG^H\Phi^H H_{R,I}^H M_I \right].
\end{align*}
\]

The derivation in (62) can be used for the second and third parts of (61).
Based on the derivation in (62), it is obvious that the second part of (61) can be derived as
\[
\sigma_E^{-2} \text{Tr} \left[ \hat{H}_E (VV^H + VEV_E^H) \hat{H}_E W_X \right]
\]
\[
= \sigma_E^{-2} \text{Tr} [H_{b,E} V_X H_{b,E}^H W_X + H_{b,E} V_X G^H \Phi H_{R,E}^H W_X + H_{R,E} \Phi GV_X H_{b,E}^H W_X
\]
\[
+ H_{R,E} \Phi GV \Phi H_{R,E}^H W_X].
\]  
(63)

Based on the derivation in (62) and by defining \( M_E = U_E W_E U_E^H \), it is obvious that the third part of (61) can be derived as
\[
\text{Tr} \left[ \hat{H}_E V_E V_E^H \hat{H}_E U_E W_E U_E^H \right]
\]
\[
= \text{Tr} \left[ \hat{H}_E (V_E V_E^H) \hat{H}_E M_E \right]
\]
\[
= \text{Tr} [H_{b,E} V_E V_E^H H_{b,E} M_E + H_{b,E} V_E V_E^H G \Phi H_{R,E}^H M_E + H_{R,E} \Phi GV_E V_E^H H_{b,E}^H M_E
\]
\[
+ H_{R,E} \Phi GV \Phi H_{R,E}^H M_E].
\]  
(64)

By adding (62), (63) and (64), and gathering constant terms irreverent with \( \Phi \), Equation (61) becomes
\[
\text{Tr} (V^H H_Y V) + \text{Tr} (V_E^H H_Y V_E)
\]
\[
= \text{Tr} [\Phi (H_{R,I}^H M_I H_{b,I} V_X G^H + \sigma_E^{-2} H_{R,E}^H W_X H_{b,E} V_X G^H + H_{R,E}^H M_E H_{b,E} V_E V_E^H G^H)]
\]
\[
+ \text{Tr} [\Phi (GV_X H_{b,I}^H M_I H_{R,I} + \sigma_E^{-2} GV_X H_{b,E}^H W_X H_{R,E} + GV_E V_E^H H_{b,E}^H M_E H_{R,E})]
\]
\[
+ \text{Tr} [\Phi GV_X G^H \Phi H_{R,I}^H M_I H_{R,I} + \sigma_E^{-2} H_{R,E}^H W_X H_{R,E})]
\]
\[
+ \text{Tr} [\Phi GV_E V_E^H G^H \Phi H_{R,E}^H M_E H_{R,E}] + C_{t_1},
\]  
(65)

where
\[
C_{t_1} = \text{Tr} [H_{b,I} V_X H_{b,I}^H M_I] + \sigma_E^{-2} \text{Tr} [H_{b,E} V_X H_{b,E}^H W_X] + \text{Tr} [H_{b,E} V_E V_E^H H_{b,E}^H M_E] .
\]  
(66)

The first term of \( g_0(V, V_E, \Phi) \) is derived as
\[
\text{Tr} \left( W_I V^H \hat{H}_I^H U_I \right) = \text{Tr} \left( U_I W_I^H V^H \hat{H}_I^H \right) = \text{Tr} [U_I W_I^H V^H \hat{H}_I^H] + \text{Tr} [H_{R,I}^H U_I W_I^H V^H G \Phi H_I^H].
\]  
(67)

The second term of \( g_0(V, V_E, \Phi) \) is derived as
\[
\text{Tr} \left( W_I U_I^H \hat{H}_I V \right) = \text{Tr} \left( \hat{H}_I V W_I U_I^H \right) = \text{Tr} [(H_{b,I} + H_{R,I} \Phi G) V W_I U_I^H]
\]
\[
= \text{Tr} [H_{b,I} V W_I U_I^H] + \text{Tr} [\Phi GV W_I U_I^H H_{R,I}].
\]  
(68)
The fourth term of $g_0(V, V_E, \Phi)$ is derived as

$$
\text{Tr} \left( W_E V_E^H \hat{H}_E^H U_E \right) = \text{Tr} \left( U_E W_E^H V_E^H \hat{H}_E^H \right)$$

$$= \text{Tr} \left[ U_E W_E^H V_E^H H_{b,E}^H \right] + \text{Tr} \left[ H_{R,E}^H U_E W_E^H V_E^H G^H \Phi^H \right]. \tag{69}
$$

The fifth term of $g_0(V, V_E, \Phi)$ is derived as

$$
\text{Tr} \left( W_E U_E^H \hat{H}_E V_E \right) = \text{Tr} \left( \hat{H}_E V_E W_E U_E^H \right)$$

$$= \text{Tr} \left[ H_{b,E} V_E W_E U_E^H \right] + \text{Tr} \left[ \Phi G V_E W_E U_E^H H_{R,E} \right]. \tag{70}
$$

By including the first term in (67), the second term in (68), the fourth term in (69), the fifth term in (70), and the sum of the third and six terms in (65) of $g_0(V, V_E, \Phi)$ and gathering constant terms irreverent with $\Phi$, we have

$$g_0(\Phi) = -\text{Equation (67)} - \text{Equation (68)} - \text{Equation (69)} - \text{Equation (70)} + \text{Equation (65)}$$

$$= \text{Tr} \left[ \Phi^H \left( H_{R,I}^H M_I H_{b,I} V_X G^H + \sigma_E^2 H_{R,E}^H W_X H_{b,E} V_X G^H + H_{R,E}^H M_E H_{b,E} V_E V_E^H G^H \right) \right]$$

$$+ \text{Tr} \left[ \Phi \left( G V_X H_{b,I} M_I H_{R,I} + \sigma_E^2 G V_X H_{b,E} W_X H_{R,E} + G V_E V_E^H H_{b,E} M_E H_{R,E} \right) \right]$$

$$+ \text{Tr} \left[ \Phi G V^H \Phi^H \left( H_{R,I}^H M_I H_{R,I} + \sigma_E^2 H_{R,E}^H W_X H_{R,E} + H_{R,E}^H M_E H_{R,E} \right) \right]$$

$$+ \text{Tr} \left[ \Phi G V^H \Phi^H \left( H_{R,I}^H M_I H_{R,I} + \sigma_E^2 H_{R,E}^H W_X H_{R,E} \right) + C_t \right]$$

$$= \text{Tr} \left[ \Phi^H \left( H_{R,I}^H M_I H_{b,I} V_X G^H + \sigma_E^2 H_{R,E}^H W_X H_{b,E} V_X G^H + H_{R,E}^H M_E H_{b,E} V_E V_E^H G^H \right) \right]$$

$$+ \text{Tr} \left[ \Phi \left( G V_X H_{b,I} M_I H_{R,I} + \sigma_E^2 G V_X H_{b,E} W_X H_{R,E} + G V_E V_E^H H_{b,E} M_E H_{R,E} \right) \right]$$

$$+ \text{Tr} \left[ \Phi G V^H \Phi^H \left( H_{R,I}^H M_I H_{R,I} + \sigma_E^2 H_{R,E}^H W_X H_{R,E} \right) + C_t \right], \tag{71}
$$

where

$$C_t = C_{t_1} + C_{t_2} + C_{t_3} + C_{t_4} + C_{t_5}. \tag{72}$$
Then $g_0(\Phi)$ becomes

$$g_0(\Phi) = \text{Tr} (\Phi^H D^H) + \text{Tr} (\Phi D) + \text{Tr} \left[ \Phi C_{VE} \Phi^H B_{VE} \right] + \text{Tr} \left( \Phi C_V \Phi^H B_V \right) + C_t$$

$$= \text{Tr} (\Phi^H D^H) + \text{Tr} (\Phi D) + \text{Tr} \left[ \Phi^H B_{VE} \Phi C_{VE} \right] + \text{Tr} \left( \Phi^H B_V \Phi C_V \right) + C_t,$$

where

$$D = G V_{X} H_{b,I}^{H} M_{I} H_{R,I} + \sigma_{E}^{-2} G V_{X} H_{b,E}^{H} W_{X} H_{R,E} + G V_{E} V_{H}^{H} H_{b,E}^{H} M_{E} H_{R,E}$$

$$- G V W_{I} U_{I}^{H} H_{R,I} - G V_{E} W_{E} U_{E}^{H} H_{R,E},$$

$$C_{VE} = G V_{E} V_{E}^{H} G^{H},$$

$$C_{V} = G V_{V}^{H} G^{H},$$

$$B_{VE} = \left( H_{R,I}^{H} U_{I} W_{I} U_{I}^{H} H_{R,I} + \sigma_{E}^{-2} H_{R,E}^{H} W_{X} H_{R,E} + H_{R,E}^{H} U_{E} W_{E} U_{E}^{H} H_{R,E} \right),$$

$$B_{V} = \left( H_{R,I}^{H} U_{I} W_{I} U_{I}^{H} H_{R,I} + \sigma_{E}^{-2} H_{R,E}^{H} W_{X} H_{R,E} \right).$$

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