Enhanced Secure Wireless Transmission Using IRS-Aided Directional Modulation

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Abstract—As an excellent aided communication technology, intelligent reflecting surface (IRS) can make a significant rate enhancement and coverage extension. In this paper, we present an investigation on beamforming in an IRS-aided directional modulation (DM) network. To fully explore the advantages of IRS, two beamforming methods with enhanced secrecy rate (SR) performance are proposed. The first method of maximizing secrecy rate (Max-SR) alternately optimizes confidential message (CM) beamforming vector, artificial noise (AN) beamforming vector and phase shift matrix. The first optimization vector is directly computed by the Rayleigh ratio and the last two are solved with generalized power iteration (GPI). This method is called Max-SR-GPI. To reduce the computational complexity, a new method of maximizing receive power with zero-forcing constraint (Max-RP-ZFC) of only reflecting CM and no AN is proposed. Simulation results show that the proposed two methods have about 30 percent rate gains over the cases of random-phase IRS and no IRS, and the proposed Max-SR-GPI performs slightly better than the Max-RP-ZFC in terms of SR, particularly in the small-large IRS.

Index Terms—Intelligent reflecting surface (IRS), directional modulation (DM), artificial noise (AN), confidential message (CM), secrecy rate (SR).

I. INTRODUCTION

With the development of the sixth generation mobile communications, intelligent reflecting surface (IRS) has been promising as a key technology to enhance rate, extend coverage and remove blind areas [1], [2], [3], [4]. IRS is becoming increasingly important in such diverse communication areas as multiple input and multiple output (MIMO) [5], spatial and directional modulation networks [6], relay [7], and covert [8]. To utilize IRS to increase the achievable rate of the downlink, the joint optimization of transmitter beamforming, IRS phase shift, IRS orientation, and position in IRS-assisted multiple-input-single-output (MISO) free-space wireless transmission systems was proposed in [4]. In [5], the authors confirmed that the cell-edge user performance can be improved by using IRS in the case of downlink multi-user MISO. In [6], IRS was proposed to assist spatial modulation that maximizes secrecy rate (SR) by adjusting the switching state of the IRS reflecting elements by power control. However, the transmission behavior of the transmitter, once detected by a malicious node, exposed the network to a security risk, in [8], and IRS-assisted artificial noise (AN) enhanced wireless covert communication was designed to achieve covert transmission rate multiplication, and more importantly the existence of perfect covertness was proved under perfect channel state information (CSI). In [9], considering a more realistic scenario without Eve’s CSI, which jointly optimized beamforming and interference to satisfy the quality of service for Bob and emitted AN to interfere with Eve. In [10], the authors demonstrated that using AN can be an effective way to help improve the security. In addition to using one IRS, the authors used two or more IRSs to further enhance the system performance in [11], [12]. In IRS-assisted secure communication [13], the authors alternately optimized the CM beamforming vector and the IRS phase shift matrix with the goal of maximizing SR.

As an advanced physical layer security technique, directional modulation (DM) was well suited for line-of-sight channel and implemented secure precise wireless transmission with the help of AN, random subcarrier selection, and beamforming in [14], [15]. In [16], with the aid of IRS, DM can achieve two-way independent confidential message (CM) streams from Alice to Bob in multipath channel, here IRS may control the phases of path gains. However, the proposed two methods require high computational amounts. In [17], a single CM symbol was transmitted from Alice to Bob using two symbol periods, which results in a significant SR loss. In [18], the authors used a semi-positive definite relaxation method to alternately optimize the CM, AN beamforming vectors and the IRS phase shift matrix with the objective of maximizing SR. In this paper, we focus on a single CM stream transmission with transmitting one CM symbol per symbol period. Two low-complexity methods are proposed to strike a good balance between complexity and performance. The main contributions of this paper are summarized as follows:

1) A system of combining IRS and DM is established to realize an enhanced single CM stream by fully making use of the advantages of DM and IRS. To achieve an improved SR performance, a method of maximizing the SR is proposed to alternately optimize the CM beamforming vector, AN beamforming vector, and IRS phase shift matrix. The first optimization vector (OV) is computed by the Rayleigh ratio, and the last two OVs are solved by converting their optimization problems into the GPI canonical forms. Since the OVs of the latter two are solved by the GPI algorithm, and the optimal SR obtained by this method is related to the initial value. Therefore, this method has high complexity.

2) To reduce the high complexity of the above method, the new method is proposed as follows: IRS only reflects CM but no AN. In other words, the AN beamforming vector is orthogonal to both channels from Alice to IRS and from Alice to Bob. Under the zero-forcing constraint (ZFC), maximizing the receive power (Max-RP) is proposed to compute the CM beamforming vector, IRS phase matrix and AN beamforming vector, respectively. Hence, this method is called Max-RP-ZFC. Compared with Max-SR-GPI method, the proposed Max-RP-ZFC method has lower computational complexity. However, the latter performs slightly worse than the former in the case of small-scale IRS. As the number of IRS elements tends to large-scale, the SR difference between them becomes trivial.

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presents the system

\[ g_{ae} = g_{e} g_{a} \]

denotes the equivalent path loss coefficient of Alice-to-IRS channel and IRS-to-Eve channel, \( h_{ae} \in \mathbb{C}^{N_{a} \times 1} \) represents the Alice-to-Eve channel, \( h_{ie} \in \mathbb{C}^{N_{i} \times 1} \) represents the IRS-to-Eve channel, and \( n_{e} \sim \mathcal{CN}(0, \sigma_{e}^{2}) \) is the AWGN at Eve. The normalized steering vector \( h(\theta) \) is defined as

\[ h(\theta) = \frac{1}{\sqrt{N}} \left[ e^{j2\pi \phi_{a}(1)}, \ldots, e^{j2\pi \phi_{a}(n)}, \ldots, e^{j2\pi \phi_{a}(N)} \right]^H, \]

where the phase shift \( \Phi_{a}(\theta) \) is defined as

\[ \Phi_{a}(\theta) = -\frac{d}{\lambda} \left( n - \frac{N+1}{2} \right) \cos \theta, \quad n = 1, \ldots, N, \]

where \( \lambda \) is the wavelength, \( n \) is the antenna index, \( d \) represents the element spacing in the transmit antenna array, and \( \theta \) is the direction of departure.

The signal received at IRS can be expressed as

\[ y_{b} = h_{a} x + h_{b} x = h_{a} x \left( \frac{\beta_{1} P_{b} v_{a} s_{a}}{\sqrt{\beta_{1} P_{b} v_{a} v_{AN} z}} \right). \]

In terms of (2), the signal-to-interference and noise ratio (SINR) at Bob is

\[ \gamma_{b} = \frac{\beta_{1} P_{b} \left| \frac{\sqrt{\beta_{1}} P_{b} v_{a} s_{a}}{\sqrt{\beta_{1} P_{b} v_{a} v_{AN} z}} \right|^{2}}{\beta_{2} P_{b} \left| \frac{\sqrt{\beta_{2}} P_{b} v_{AN} z}{\sqrt{\beta_{2} P_{b} v_{AN} z}} \right|^{2} + \sigma_{e}^{2}}. \]

From (3), the SINR of Eve is

\[ \gamma_{e} = \frac{\beta_{2} P_{b} \left| \frac{\sqrt{\beta_{2}} P_{b} v_{AN} z}{\sqrt{\beta_{2} P_{b} v_{AN} z}} \right|^{2}}{\beta_{1} P_{b} \left| \frac{\sqrt{\beta_{1}} P_{b} v_{a} s_{a}}{\sqrt{\beta_{1} P_{b} v_{a} v_{AN} z}} \right|^{2} + \sigma_{e}^{2}}. \]

The corresponding rates at Bob and Eve are as follows

\[ R_{b} = \log_{2} (1 + \gamma_{b}), \]

and

\[ R_{e} = \log_{2} (1 + \gamma_{e}), \]

respectively, which directly gives the secrecy rate as

\[ R_{s} = [R_{b} - R_{e}]^{+} = \log_{2} \left( 1 + \frac{\gamma_{e}}{1 + \gamma_{e}} \right), \]

where \([x]^{+} = \max\{0, x\} \) is the additive white Gaussian noise (AWGN) at Bob.

Similarly, the signal received at Eve can be expressed as

\[ y_{e} = \left( \frac{\sqrt{\gamma_{ae}} P_{e} h_{ae}^{H} v_{a} s_{a}}{\sqrt{\gamma_{ae}} P_{e} h_{ae}^{H} v_{a} v_{AN} z} \right) x + n_{e} \]

II. SYSTEM MODEL

In Fig. 1, an IRS-aided DM network is shown. Alice is equipped with \( N_{a} \) antennas, IRS has \( N_{r} \) reflecting elements, and Bob and Eve are employed with single antenna. The transmit baseband signal is in the form

\[ x = \sqrt{\beta_{1} P_{b}} v_{a} s_{a} + \sqrt{\beta_{2} P_{b}} v_{AN} z, \]

where \( P_{b} \) is the transmit power, \( s_{a} \) denotes the CM with a constraint \( E[|s_{a}|^{2}] = 1 \), and \( \beta_{1} \) is the AN with the average power constraint \( E[|z|^{2}] = 1 \). \( v_{a} \) denotes the precoding vector of CM, and \( v_{AN} \) is the precode vector of AN with \( v_{a} \in \mathbb{C}^{N_{a} \times 1} \) and \( \sqrt{\beta_{1} P_{b}} v_{AN} \in \mathbb{C}^{N_{a} \times 1} \). \( \beta_{1} \) and \( \beta_{2} \) respectively represent the power allocation (PA) factors of CM and AN with \( \beta_{1} + \beta_{2} = 1 \). The signal received at Bob can be represented as

\[ y_{b} = \left( \frac{\sqrt{\gamma_{ab} h_{ab}^{H}}}{\gamma_{ab}} \right) x + n_{b} \]

\[ = \sqrt{\gamma_{ab} P_{b} h_{ab}^{H} v_{a} s_{a}} + \sqrt{\gamma_{ab} P_{b} h_{ab}^{H} \Theta H_{a} v_{a} s_{a}} + n_{b} \]

\[ \sqrt{\gamma_{ab} P_{b} h_{ab}^{H} v_{AN} z} + \sqrt{\gamma_{ab} P_{b} h_{ab}^{H} \Theta H_{a} v_{AN} z}, \]

where \( y_{b} \) is the path loss coefficient between Alice and Bob, and \( g_{ab} = g_{a} g_{b} \) denotes the equivalent path loss coefficient of Alice-to-IRS channel and IRS-to-Bob channel. The path loss coefficient is calculated as: \( y_{b} = \alpha/d_{b} \), where \( \alpha = 10^{-2}, c = 2 \) and \( d_{b} \) is the reference distance. \( h_{ab} \in \mathbb{C}^{N_{a} \times 1} \) represents the Alice-to-IRS channel, \( h_{ie} \in \mathbb{C}^{N_{i} \times 1} \) represents the IRS-to-Bob channel, \( \Theta = \text{diag}(e^{j\phi_{1}}, \ldots, e^{j\phi_{N_{r}}}) \) is a diagonal matrix with the phase shift \( \phi_{n} \) incurred by the m-th reflecting element of the IRS, \( \Theta = \text{diag}(\theta) \) with \( \theta \in \mathbb{C}^{N_{i} \times 1} \). \( H_{a} = h(\theta_{a}, r) h^{H}(\theta_{a}, i) \) is the equivalent channel, and \( n_{b} \sim \mathcal{CN}(0, \sigma_{e}^{2}) \) is the additive white Gaussian noise (AWGN) at Bob.

Similarly, the signal received at Eve can be expressed as

\[ y_{e} = \left( \frac{\sqrt{\gamma_{ae} h_{ae}^{H}}}{\gamma_{ae}} \right) x + n_{e} \]

III. TWO PROPOSED BEAMFORMING METHODS WITH ENHANCED PERFORMANCE

In what follows, to harvest the SR performance gain available by IRS, two iterative methods, called Max-SR-GPI and Max-RP-ZFC, are proposed. The former is of high-performance while the latter is of low-complexity.
A. Proposed Max-SR-GPI

The optimization problem of maximizing the SR can be cast as

$$\begin{align}
\max_{v_a, v_{AN}, \Theta} R_b (v_a, v_{AN}, \Theta) & \\
\text{s.t. } v_ahv_a = 1, v_{AN}v_{AN} = 1, |\Theta| = 1.
\end{align}$$

(12a)

The rate of Bob in the above SR can be rewritten as

$$R_b = \log_2 \left( 1 + \frac{v_a^H R^b v_a}{v_{AN}^H G_g v_{AN} + \sigma_b^2} \right),$$

(13)

where

$$R^b = \left( \sqrt{\beta_1 p_{t} g_{a} h_{ab}^H} + \sqrt{\beta_2 p_{t} g_{a} h_{ab}^H} \Theta \Theta_h \right),$$

$$G_g = \left( \sqrt{\beta_2 p_{t} g_{a} h_{ab}^H} + \sqrt{\beta_1 p_{t} g_{a} h_{ab}^H} \Theta \Theta_h \right).$$

Similarly, the rate of Eve can be rewritten as

$$R_e = \log_2 \left( 1 + \frac{v_a^H pp^H v_a}{v_{AN}^H qq^H v_{AN} + \sigma_e^2} \right),$$

(15)

where

$$p^H = \left( \sqrt{\beta_1 p_{t} g_{a} h_{ab}^H} + \sqrt{\beta_2 p_{t} g_{a} h_{ab}^H} \Theta \Theta_h \right),$$

$$q^H = \left( \sqrt{\beta_2 p_{t} g_{a} h_{ab}^H} + \sqrt{\beta_1 p_{t} g_{a} h_{ab}^H} \Theta \Theta_h \right).$$

According to (13) and (15), given $\Theta$ and $v_{AN}$, the optimization problem in (12) is converted into

$$\max_{v_a} v_a \left( (a + \sigma_e^2) I_N + pp^H \right) v_a,$$

$$v_{AN}^H v_{AN} = 1,$$

where $a = v_{AN}^H gg^H v_{AN}$, and $b = v_{AN}^H qq^H v_{AN}$ due to the fact that the logarithm function is a monotonically increasing function.

Therefore, using the Rayleigh-Ritz ratio theorem, $v_a$ is the eigenvector corresponding to the largest eigenvalue of the following matrix

$$\left( (b + \sigma_e^2) I_N + pp^H \right)^{-1} \left( (a + \sigma_e^2) I_N + pp^H \right).$$

Similarly, given $\Theta$ and $v_a$, the optimization problem in (12) is converted into

$$\max_{v_{AN}} \frac{v_a^H v_{AN} F_{v_{AN}}}{v_{AN}^H v_{AN} F_{v_{AN}}} \times \frac{v_a^H M v_{AN}}{v_{AN}^H N v_{AN}} \text{ s.t. } v_{AN}^H v_{AN} = 1,$$

(19)

where

$$c = v_a^H pp^H v_a, d = v_a^H pp^H v_a, E = (c + \sigma_e^2) I_N + gg^H, F = \sigma_e^2 I_N + gg^H, M = \sigma_e^2 I_N + qq^H, N = (d + \sigma_e^2) I_N + qq^H.$$

Therefore, $v_{AN}$ can be solved by using GPl algorithm in [19].

Given $v_{AN}$ and $v_a$, let us define a new optimization variable $\overline{\Theta} = [1 \ \Theta^H]^H$. Accordingly, the rate of Bob can be rewritten as

$$R_b = \log_2 \left( 1 + \frac{P_t (\overline{\Theta}^H w w^H r)}{P_t (\overline{\Theta}^H v v^H r) + \sigma_e^2} \right),$$

(21)

where

$$w^H = \left[ \sqrt{\beta_1 g_{a} h_{ab}^H} v_a \ \sqrt{\beta_2 g_{a} h_{ab}^H} \Theta_h \right],$$

$$v^H = \left[ \sqrt{\beta_2 g_{a} h_{ab}^H} v_{AN} \ \sqrt{\beta_1 g_{a} h_{ab}^H} \Theta_h \right].$$

Algorithm 1: Proposed Max-SR-GPI Method.

1: Set initial solution $\Theta^{(0)}$, $v_{AN}^{(0)}$, and $v_a^{(0)}$. Randomly take the value of $\Theta$, and calculate the initial $R_b^{(0)}$ multiple times based on formula (11).

2: Set $p=0$, threshold $\epsilon$.

3: repeat

4: Given $(\Theta^{(p)}, v_{AN}^{(p)})$, according to (18) to get $v_a^{(p+1)}$.

5: Given $(\Theta^{(p)}, v_a^{(p+1)})$, according to (19) to get $v_{AN}^{(p+1)}$.

6: Given $(v_a^{(p+1)}, v_{AN}^{(p+1)})$, according to (25) to get $\Theta^{(p+1)}$.

7: Compute $R_b^{(p+1)}$ using $v_a^{(p+1)}$, $v_{AN}^{(p+1)}$, and $\Theta^{(p+1)}$.

8: $p=p+1$;

9: until $R_b^{(p)} - R_b^{(p-1)} \leq \epsilon$, and record the maximum SR value $R_b^{(p)}$.

Similarly, the rate of Eve can be rewritten as

$$R_e = \log_2 \left( 1 + \frac{P_t \left( \overline{\Theta}^H m m^H \overline{\Theta} \right)}{P_t \left( \overline{\Theta}^H n n^H \overline{\Theta} \right) + \sigma_e^2} \right),$$

(23)

where

$$m^H = \left[ \sqrt{\beta_1 g_{a} h_{ab}^H} v_a \ \sqrt{\beta_2 g_{a} h_{ab}^H} \Theta_h \right]$$

$$n^H = \left[ \sqrt{\beta_2 g_{a} h_{ab}^H} \Theta_h \right] \left[ \sqrt{\beta_2 g_{a} h_{ab}^H} \Theta_h \right] v_{AN} \ \left[ \sqrt{\beta_2 g_{a} h_{ab}^H} \Theta_h \right] \left[ \sqrt{\beta_2 g_{a} h_{ab}^H} \Theta_h \right] \Theta_h.$$}

Therefore, the optimization problem is converted into

$$\max_{\overline{\Theta}} \frac{\overline{\Theta}^H Q \overline{\Theta}}{\overline{\Theta}^H K \overline{\Theta}} \times \frac{\overline{\Theta}^H T \overline{\Theta}}{\overline{\Theta}^H R \overline{\Theta}} \text{ s.t. } \overline{\Theta}^H \overline{\Theta} = N_e + 1,$$

(25)

where

$$Q = \left( P_t v v^H + P_t w w^H + \frac{I_{N_e+1}}{N_e+1} \sigma_e^2 \right),$$

$$K = \left( P_t v v^H + \frac{I_{N_e+1}}{N_e+1} \sigma_e^2 \right) \left( P_t n n^H + \frac{I_{N_e+1}}{N_e+1} \sigma_e^2 \right),$$

$$R = \left( P_t n n^H + P_t m m^H + \frac{I_{N_e+1}}{N_e+1} \sigma_e^2 \right).$$

Finally, $\overline{\Theta}$ in (25) can be solved via GPl in [19]. The whole procedure is summarized in the following Table.

The complexity of Max-SR-GPI method is $O(L_1 (L_2 (3 (N_e + 1)^2) + L_3 (3N_e^2)))$ flop-point operations (FLOPs), where $L_1$, $L_2$, and $L_3$ denote the iterative numbers of optimization variables $v_a$, $v_{AN}$, and $\Theta$.

B. Proposed Max-RP-ZFC

In the previous subsection, the optimization variables $v_{AN}$ and $\Theta$ are computed by the iterative method GPl. The computational complexity is a linearly increasing function of their numbers of iterations. To reduce the complexity, a low-complexity method Max-RP-ZFC is proposed to solve $v_{AN}$ by ZF and $\Theta$ by maximizing receive power (RP) in closed-form.

First, we optimize the AN beamforming vector $v_{AN}$, which is independent of $\Theta$ and $v_a$. Below, maximizing the receive AN power along the direct channel from Alice to Eve at Eve with respect to $v_{AN}$ is formulated as

$$\max_{v_{AN}} \frac{v_{AN}^H h_{ab} h_{ab}^H v_{AN}}{v_{AN}^H v_{AN}} \text{ s.t. } (h_{ab} H_{ab}^H)^H v_{AN} = 0, v_{AN}^H v_{AN} = 1.$$
Define \( G = (h_{ab} H_b^H)^H \), \( T_{-a} = [I_{N_a} - G H (GG^H)^H] \), and \( v_{AN} = T_{-a}^{-1} u_{AN} \), then, problem (27) can be simplified as
\[
\max_{u_{AN}} u_{AN}^H T_{-a}^{-1} h_{ab}^H T_{-a}^{-1} u_{AN} \quad \text{s.t.} \quad u_{AN}^H u_{AN} = 1. \quad (28)
\]
which directly gives
\[
v_{AN} = T_{-a}^{-1} h_{ab} \quad \text{s.t.} \quad \| T_{-a}^{-1} h_{ab} \| \quad (29)
\]
due to the fact that matrix \( T_{-a}^{-1} h_{ab} h_{ab}^H T_{-a}^{-1} \) is a rank-one matrix. Now, we establish a joint two-variable \((v_a, \theta)\) optimization problem of maximizing RP at Bob as follows
\[
\max_{v_a, \theta} v_a^H (h_{ib}^H \Theta H_{ai} + h_{ib}^H) v_a \quad \text{s.t.} \quad h_{ii}^H v_a = 0, \quad v_a^H v_a = 1, \quad \theta^H \theta = N_r. \quad (30a)
\]

Similar to (28), fixing \( \Theta \), we have
\[
v_a = P (h_{ib}^H \Theta H_{ai} + h_{ib}^H) / \| P (h_{ib}^H \Theta H_{ai} + h_{ib}^H)^H \|, \quad (31)
\]
where \( P = I_{N_a} - h_{ab} h_{ab}^H \). Then, fixing \( v_a \), (30) can be rewritten as
\[
\max_{\theta} u_{b}^H P (h_{ib}^H \Theta H_{ai} + h_{ib}^H) (h_{ib}^H \Theta H_{ai} + h_{ib}^H) P u_{b} \quad \text{s.t.} \quad \theta^H \theta = N_r, \quad (32a)
\]
where \( u_{b}^H u_{b} = 1 \). The objective function of (32) can be expressed in the following quadratic form
\[
\theta^H \text{diag} \left( u_{b}^H P (h_{ib}^H H_{ai} + h_{ib}^H) \right) h_{ib} h_{ib}^H \text{diag} \left( H_{ai} P u_{b} \right) \theta + \theta^H \text{diag} \left( u_{b}^H P (h_{ib}^H H_{ai} + h_{ib}^H) \right) h_{ib} h_{ib}^H \text{diag} \left( H_{ai} P u_{b} \right) \theta + u_{b}^H P h_{ib} h_{ib}^H P u_{b}. \quad (33)
\]
Substituting the above expression in (32) yields
\[
\max_{\theta} \theta^H A \theta + \theta^H b + b^H \theta + C \quad \text{s.t.} \quad \theta^H \theta = N_r, \quad (34)
\]
where
\[
A = \text{diag} \left( u_{b}^H P (h_{ib}^H H_{ai} + h_{ib}^H) \right) h_{ib} h_{ib}^H \text{diag} \left( H_{ai} P u_{b} \right),
\]
\[
b = \text{diag} \left( u_{b}^H P (h_{ib}^H H_{ai} + h_{ib}^H) \right) h_{ib} h_{ib}^H P u_{b},
\]
\[
C = u_{b}^H P h_{ib} h_{ib}^H P u_{b}. \quad (35)
\]
The Lagrangian function of (34) can be expressed as
\[
f(\theta, \lambda) = \theta^H A \theta + \theta^H b + b^H \theta + C + \lambda \left( \theta^H \theta - N_r \right). \quad (36)
\]
whose partial derivative with respect to \( \theta \) is set to 0 to obtain the following equation
\[
\frac{\partial f(\theta, \lambda)}{\partial \theta} = A \theta + \lambda \theta + b = 0. \quad (37)
\]
which generates
\[
\theta = -(A + \lambda I_{N_r})^{-1} b. \quad (38)
\]
Since \( A = a a^H \) is a matrix of rank-one, using the Sherman-Morrison formula, the constraint of (34) can be expressed as
\[
b^H \left( \frac{1}{\lambda^2} + \frac{(aa^H)^2}{\lambda^2} - 2aa^H \right) \lambda \left( a + a^H a \right) b = N_r. \quad (39)
\]
which can be simplified as
\[
N_r \lambda^4 + 2N_r a_i \lambda^3 + (N_r a_i^2 - b_i) \lambda^2 - 2c_1 \lambda - c_2 = 0, \quad (40)
\]
where
\[
a_1 = a^H a, \quad b_1 = b^H b, \quad c_1 = b^H ba^H a,
\]
\[
d_1 = b^H aa^H aa^H b, \quad e_1 = b^H aa^H b. \quad (41)
\]
Observing (40), it is a fourth-order polynomial and has a set \( S \) of four roots denoted as: \( \lambda_1, \lambda_2, \lambda_3, \) and \( \lambda_4 \). Finding the optimal value of \( \lambda_0 \) modelled as the following maximum problem
\[
\lambda_0 = \arg \max_{\lambda \in S} (36). \quad (42)
\]
Alternating iteration between \( v_a \) and \( \theta \) are repeated until \( H_s^{(p)} - H_s^{(p-1)} \leq \epsilon \).

The complexity of Max-RP-ZFC is \( O(L_a(N^3 + 14N^2)) \) FLOPs, where \( L_a \) denotes the alternating iterative number between \( v_a \) and \( \theta \).

**IV. SIMULATION RESULTS AND DISCUSSIONS**

In this section, numerical simulation results are presented to evaluate the SR and convergent performance of our proposed methods. Simulation parameters are set as follows: \( P_s = 30 \text{ dBm}, \sigma_n^2 = \sigma_z^2 = -40 \text{ dBm}, \) and \( N_a = 16 \). The distances and angles are set as \( d_{ab} = 20 \text{ m}, d_{ah} = 40 \text{ m}, d_{ah} = 20.0547 \text{ m}, d_{ah} = 33.6765 \text{ m}, d_{ah} = 50 \text{ m}, \theta_a = 29\pi/60, \theta_{ab} = 1\pi/2, \) and \( \theta_{ah} = 23\pi/36 \).

Fig. 2 shows the SR versus the number of iterations for three typical numbers of elements of IRS as follows: 32, 128, and 1024. It is seen that the proposed two methods rapidly converge to the SR cell with...
only $4 \sim 5$ iterations. Also, the SR performance gain achieved by IRS is very attractive.

Fig. 3 depicts the computational complexity versus $N_r$. It is obvious that the two methods proposed in this paper are low-complexity with order $O(N_r^3)$ FLOPs whereas the AO and SDR complexities are the orders $O(N_r^{1.5})$ and $O(N_r^{4.5})$ FLOPs, respectively.

Fig. 4 plots the SR versus $N_r$. The SR performance of the proposed two methods is better than those of no IRS, random phase and existing method in [13] and [17]. In summary, the two proposed methods strike a good balance between SR and complexity.

V. CONCLUSION

In this paper, we have investigated of the IRS-assisted DM networks with single-CM-stream transmission. In order to improve the SR performance, two high-performance methods Max-SR-GPI and Max-RP-ZFC were proposed. Simulation results showed that the proposed two methods harvest obvious SR performance gains over no IRS, random phase and existing method [17], especially in the case of large-scale IRS. Additionally, they can converge rapidly. The Max-RP-ZFC was far lower than Max-SR-GPI in terms of computational complexity, particularly in the case of large-scale IRS.

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