SCIENCE CHINA
Information Sciences

• RESEARCH PAPER •

March 2023, Vol. 66 132303:1–132303:10
https://doi.org/10.1007/s11432-021-3494-2

Joint information transmission design for intelligent reflecting surface aided system with discrete phase shifts

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Received 23 December 2021/Revised 5 April 2022/Accepted 11 May 2022/Published online 17 February 2023

Abstract This paper studies the joint information transmission for an intelligent reflecting surface (IRS) aided single-input multiple-output (SIMO) communication system, where an IRS with discrete phase shifts aims to deliver its sensed information to the receiver, besides being a helper of the SIMO system. We first introduce the joint modulation strategy for delivering the information of the primary system and the IRS simultaneously. Specifically, the IRS modulates its information in both the received antenna index and the received signal phase; while the transmitter limits its signal phase in a small region to avoid aliasing. For the proposed strategy, we devise an IRS discrete phase shifts design to maximize the received signal power at the selected receive antenna. Then, we propose a receiver constellation design for the IRS's and transmitter's information to maximize the minimum Euclidean distance. Numerical results show that the proposed scheme is practical and competitive compared with several existing joint information transmission schemes.

Keywords IRS, information transmission, constellation design

Citation Tao Q, Zhang S W, Zhong C J, et al. Joint information transmission design for intelligent reflecting surface aided system with discrete phase shifts. Sci China Inf Sci, 2023, 66(3): 132303, https://doi.org/10.1007/s11432-021-3494-2

1 Introduction

Intelligent reflecting surface (IRS) is a technology that can reflect the incident signal with adjustable amplitude/phase shift, so as to reconfigure the wireless environment to favor communication at a low cost. As such, IRS has emerged as an attractive approach for beyond 5G wireless networks and is gaining increasing research interest [1–7]. For example, the authors in [8] proposed to utilize the IRS to assist a multiple-input single-output (MISO) system by jointly optimizing the base station (BS) active beamforming and IRS passive beamforming. For a multiple-input multiple-output (MIMO) system, Ref. [9] found that by jointly designing the BS and IRS beamforming, the user can achieve substantially increased capacity compared with the traditional system without the IRS. Focusing on non-orthogonal multiple access (NOMA) networks, Refs. [10, 11] improved the sum rate of the system by properly designing the IRS.

The above studies all treated the IRS as a helper for the primary communication system. However, in practice, the IRS may be equipped with sensors and required to upload the sensed data occasionally. Due to the lack of baseband processing capability and passiveness of the IRS, IRSs cannot perform baseband modulation and actively emit the modulated signals. Therefore, new mechanisms need to be designed to send the IRS's information. Along this line, Refs. [12–14] studied the IRS-based transmitter, where the IRS and an RF generator are together treated as transmitter so as to reduce the hardware cost. As proved, the phase shift keying modulation, quadrature amplitude modulation, receive quadrature reflecting modulation and receive antenna index modulation (IM) can be realized at the transmitter.

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Considering an IRS-aided single-input multiple-output (SIMO) communication system [15], to jointly transfer the IRS and transmitter information, the IM on the receive antenna was proposed for IRS information, while the amplitude/phase modulation was used for the transmitter signal. The authors in [16] proposed to transfer the IRS binary symbol by selecting two reflection matrices, at the same time enhancing the quadrature phase shift keying modulated MISO system. For the above methods, the number of information bits contained in each IRS symbol is strictly limited to the small number of receive antennas or small modulation order. To improve the data rate of IRS, the authors in [17] proposed to modulate the IRS information in the index of the large number of reflecting patterns, whose number is usually much larger than the number of receive antennas, and then, ML detection is performed to recover the information. The authors in [18,19] studied the joint information transmission schemes for single-user and multi-user scenarios, respectively, where the information of the IRS is encoded in the index of IRS elements, and several algorithms are proposed to separate the IRS's and transmitter's symbols, such as the bilinear generalized approximate message passing (BiG-AMP) and the generalized approximate message passing (GAMP) algorithms. However, the complexity of these algorithms is very high, which is practically unaffordable.

To further increase the data rate of IRS and reduce the detection complexity, in this study, we propose a novel joint modulation strategy to deliver the information from the primary communication system and the IRS simultaneously. Specifically, the IRS modulates its information in both the receive antenna index and the received signal phase; while the transmitter of the primary system encodes its information in both the amplitude and phase of the received signal, where the phase is limited to a small region to avoid aliasing with the IRS's information phase. Note that with our proposed scheme, only one single RF chain is needed at the receiver. To assist the transmitter's information transmission, the IRS with discrete phase shift is designed to maximize the received signal power at the chosen antenna. Then, based on the proposed modulation strategy, the receiver constellation is designed to maximize the minimum Euclidean distance (MED). Finally, numerical results show that our proposed scheme outperforms other benchmark schemes in various aspects.

2 System model

We consider a SIMO communication system with one single-antenna transmitter and one multi-antenna receiver equipped with $M > 1$ antennas, as illustrated in Figure 1. The direct channel between the transmitter and the receiver is assumed to be blocked due to obstacles; thus an IRS with N reflecting elements is deployed to create an additional communication link for the primary SIMO system. Moreover, the IRS is assumed to be equipped with sensors and is required to deliver the sensed information (e.g., temperature) to the receiver. Particularly, we consider a low-rate control and feedback link between the transmitter and the IRS, through which the transmitter sends control signals to the IRS for tuning the reflection coefficients.

Let $h_{1,n}$ denote the channel coefficient from the transmitter to the n-th IRS reflecting element, and $h_{2,m,n}$ denote the channel coefficient from the n-th IRS reflecting element to the m-th antenna at the receiver. Let $\theta_n \in [0, 2\pi)$ denote the phase shift at the n-th IRS reflecting element, and $\boldsymbol{\Theta} = \left[\mathrm{e}^{\mathrm{j}\theta_1}, \ldots, \mathrm{e}^{\mathrm{j}\theta_n}, \ldots, \mathrm{e}^{\mathrm{j}\theta_N}\right]$ denote the reflection vector at the N elements, by considering a common unit reflection amplitude at all the elements due to the difficulty in jointly tuning the phase shift and reflection amplitude at the IRS. In majority of the existing studies, the phase shifts θ_n's are assumed to be continuous variables that can be arbitrarily selected from $[0, 2\pi)$, which incurs large hardware cost and complexity [20]. In this study, we consider discrete phase shifts at the IRS, where each phase shift θ_n can be selected from a finite set with $B \geqslant 1$ phase shift levels uniformly distributed within $[0, 2\pi)$, denoted as $\mathcal{B} = \{0, \frac{2\pi}{B}, \ldots, \frac{2\pi(B-1)}{B}\}$.

To simultaneously deliver the information of the transmitter and the IRS to the receiver, we need to carefully design an efficient joint modulation strategy. First, we propose a novel strategy for modulating the information from the IRS in both the received antenna index and the received signal phase. Specifically, on one hand, the IRS adjusts its phase shifts to maximize the noise-free received signal power at the m-th receive antenna, where the selected antenna index m is determined by $\log_2 M$ information bits.

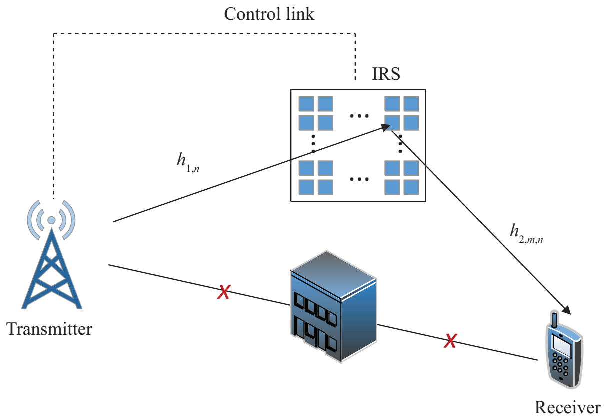

Figure 1 An IRS-aided SIMO communication system.

The problem of designing the IRS phase shifts is formulated as

$$\text{(P1)} \quad \max_{\boldsymbol{\Theta}} \quad \left| \sum_{n=1}^{N} h_{1,n} h_{2,m,n} e^{j\theta_n} \right|^2 \tag{1}$$

$$\text{s.t.} \quad \theta_n \in \mathcal{B}, \ \forall n. \tag{2}$$

In Section 3, we will propose an efficient algorithm for finding the optimal solution to (P1), which is denoted as $\boldsymbol{\Theta}_m^* = [e^{j\theta_{m,1}^*}, \ldots, e^{j\theta_{m,n}^*}, \ldots, e^{j\theta_{m,N}^*}]$. With $\boldsymbol{\Theta}_m^*$, the equivalent channel from the transmitter to the m-th receive antenna is expressed as $H_{\text{opt}}^m = \sum_{n=1}^{N} h_{1,n} h_{2,m,n} e^{j\theta_{m,n}^*}$. On the other hand, based on the optimal phase shifts $\boldsymbol{\Theta}_m^*$, the IRS further adds a common phase shift $\frac{2\pi b}{B}$ to all the elements, where $b \in \{0, 1, \ldots, B-1\}$ is determined by $\log_2 B$ information bits. As such, the IRS is able to deliver $\log_2 M + \log_2 B$ information bits in each symbol interval, which is $\log_2 B$ bits larger than the index modulation scheme proposed in [15].

Next, we introduce the modulation strategy of the transmitter's information. With the IRS configurations described above, the baseband received signal at the m-th antenna is given by

$$y_m = \sqrt{E} H_{\text{opt}}^m e^{j\frac{2\pi b}{B}} s + z_m, \tag{3}$$

where E denotes the average transmit power; $s \in \mathbb{C}$ denotes the information symbol for the transmitter, with average power $\mathbb{E}[|s|^2] = 1$; $z_m \sim \mathcal{CN}(0, \sigma^2)$ denotes the circularly symmetric complex Gaussian (CSCG) noise at the m-th receive antenna with zero mean and variance σ^2. For the simplicity of notation, let $\delta = \arg\{H_{\text{opt}}^m\}$, $\phi_I = \frac{2\pi b}{B}$ and $\phi_T = \arg\{s\}$, thus the phase of the noise-free received signal is $\vartheta \triangleq \arg\{H_{\text{opt}}^m e^{j\frac{2\pi b}{B}} s\} = \delta + \phi_I + \phi_T$, from which it is difficult to separate ϕ_I and ϕ_T due to their aliasing. To recover ϕ_I and ϕ_T from the received signal, we propose to limit the value of ϕ_T in the region $[0, \frac{2\pi}{B})$, since the possible values of ϕ_I differ by at least $\frac{2\pi}{B}$. As such, the phase of the transmitter and the IRS can be uniquely determined as $\phi_T = (\vartheta - \delta) \mod \frac{2\pi}{B}$ and $\phi_I = \vartheta - \delta - \phi_T$, respectively, based on the knowledge of δ which can be obtained via channel estimation at the transmitter and feedback to the receiver. For example, assuming $\delta = 0$, $\phi_I \in \left\{0, \frac{\pi}{2}, \pi, \frac{3\pi}{2}\right\}$, and ϕ_T is limited to the set $\left\{0, \frac{\pi}{8}, \frac{\pi}{4}, \frac{3\pi}{8}\right\}$. If $\vartheta = \frac{\pi}{4}$, we have $\phi_I = 0$ and $\phi_T = \frac{\pi}{4}$; if $\vartheta = \frac{5\pi}{8}$, we have $\phi_I = \frac{\pi}{2}$ and $\phi_T = \frac{\pi}{8}$.

Note that with B possible values of b in the IRS's remaining information and Q possible values for the transmitter's information signal s, the normalized noise-free received signal $e^{j\frac{2\pi b}{B}} s$ has BQ possible values. This thus motivates us to design a BQ-ary constellation that carries both the information of the IRS and the transmitter. Specifically, we propose to consider the amplitude and phase shift keying (APSK) constellation where the symbols are located on multiple concentric rings, since: (1) each symbol $e^{j\frac{2\pi b}{B}} s$ is the phase-rotated version of the transmitter's signal, with the phase rotation determined by the IRS's information, which can be easily characterized under the APSK framework; (2) the APSK constellation yields lower detection error compared to quadrature amplitude modulation (QAM) constellations under the same power constraint. In Section 3, we will investigate the APSK constellation optimization.

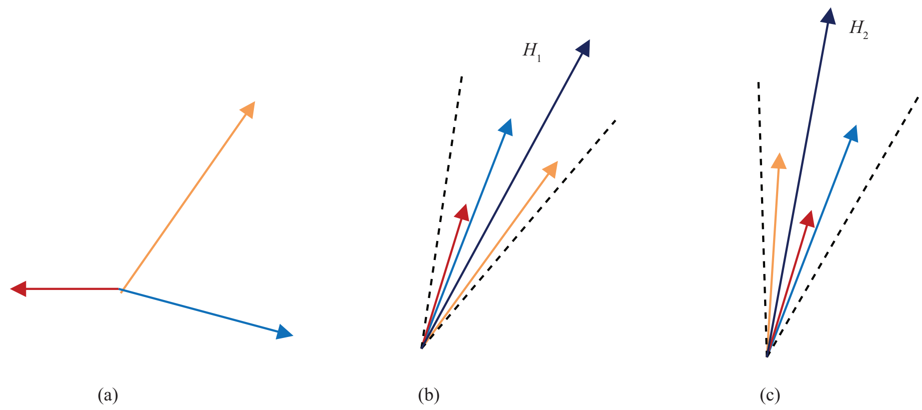

Figure 2 Illustration of the proposed design for the IRS discrete phase shifts. (a) The original cascaded reflected channels; (b) rotating all channels into a $2\pi/B$ region; (c) adding $2\pi/B$ phase shift to the channel with the smallest phase.

3 Optimization of IRS discrete phase shifts

In this section, we will solve the problem (P1) to find the IRS discrete phase shift design for maximizing the received signal power at the selected m-th receive antenna. Note that due to the discrete phase shift constraint in (2), (P1) is much more difficult to solve as compared to the case with continuous phase shifts, for which a closed-form optimal solution can be readily derived. In the literature, there are three major types of methods for solving discrete phase shift design problems:

- Exhaustive search over all possible discrete phase shift sets, which finds the optimal solution with a prohibitive complexity of $\mathcal{O}(B^N)$.
- Branch-and-bound (e.g., [20]), which finds the optimal solution with worst-case complexity exponential over N.
- Projection method by first obtaining the optimal continuous phase shifts, and then projecting them to the nearest discrete phase shifts. However, this approach is heuristic and may lead to significantly compromised performance.

Motivated by the limitations of the existing methods, in this study, we propose a novel method that finds the optimal solution to the problem (P1) based on constraining the cascaded channel phases in a small region, with complexity $\mathcal{O}(N)$.

The main idea of our proposed algorithm is illustrated in Figure 2 (by taking the example of $N=3$) and described as follows. The original cascaded reflected channels $h_{1,n}h_{2,m,n}$ for $n=1$ to 3 are plotted by solid lines with different colors, as shown in Figure 2(a). Obviously, to maximize $|\sum_{n=1}^{N} h_{1,n}h_{2,m,n}\mathrm{e}^{\mathrm{j}\theta_n}|^2$, we need to let the phase all cascaded reflected channels $h_{1,n}h_{2,m,n}\mathrm{e}^{\mathrm{j}\theta_n}$s fall into the smallest region. First, we tune the discrete phase shifts θ_ns such that the phase of all channels $h_{1,n}h_{2,m,n}\mathrm{e}^{\mathrm{j}\theta_n}$s fall into a region of $\frac{2\pi}{B}$, which is the minimum range can always ensure all channels are inside, e.g., $\left[0, \frac{2\pi}{B}\right)$, where the boundaries are plotted by dash lines in Figure 2(b). Then, we propose to add a $\frac{2\pi}{B}$ phase shift to the channel with the smallest phase successively, checking the resulted objective value of problem (P1), and finally select the set of phase shifts with the largest objective value. Specifically, note that adding $\frac{2\pi}{B}$ phase shift to the cascaded reflected channel with the smallest phase will guarantee that all the channels are still in the range of $\frac{2\pi}{B}$, and can experience all possible equivalent states, thus ensuring the optimality of the proposed algorithm. The detailed algorithm is given in Algorithm 1.

4 Optimization of the APSK constellation

At the receiver, the index of the selected antenna m is first detected via a one-dimensional search, i.e.,

$$m = \arg \max_{\bar{m}=1,\ldots,M} |y_{\bar{m}}|^2, \tag{4}$$

which represents $\log_2 M$ bits from the IRS. Then, the receiver recovers the remaining $\log_2 B$ bits from the IRS and $\log_2 Q$ bits from the transmitter based on y_m. In the following, we aim to optimize the BQ-ary APSK constellation for the noise-free signal in y_m, i.e., $\mathrm{e}^{\mathrm{j}\frac{2\pi b}{B}}s$, to minimize the symbol error rate (SER), P_s. Specifically, since the exact SER has a complicated expression which is difficult to

Algorithm 1 IRS discrete phase shift design for (P1)

Input: $\{h_{1,n}h_{2,m,n}\}_{n=1}^N, B$;
Output: $\boldsymbol{\Theta}_{m,0}^*$, $H_{\rm opt}^m$;
1: Set $\{p_n\}_{n=1}^N = \lfloor -\arg\{h_{1,n}h_{2,m,n}\}\frac{B}{2\pi} - \frac{1}{2}\rfloor$, $\{S_n\}_{n=1}^N = \arg\{h_{1,n}h_{2,m,n}\} + \frac{2\pi p_n}{B}$, $\{\theta_{m,n,0}^*\}_{n=1}^N = \frac{2\pi p_n}{B}$;
2: Sort $\{S_n\}_{n=1}^N$ in ascending order and store the corresponding index in vector $\boldsymbol{I} = [I_1, \ldots, I_N]$, hence, we have $S_{I_1} < S_{I_2} < \cdots < S_{I_N}$;
3: **for** $i = 1$ to N
4: Set $S_{I_i} = S_{I_i} + \frac{2\pi}{B}$ and $H_i = \sum_{n=1}^N |h_{1,n}h_{2,m,n}|{\rm e}^{{\rm j}S_n}$;
5: **end**
6: Let $[H_{\max}^m, {\rm IDX}] = \max\{|H_n|\}_{n=1}^N$, where index represents the index of maximum value in $\{|H_n|\}_{n=1}^N$;
7: Set $\Theta_0^* = \arg(H^*)$;
8: **for** $i = 1$ to IDX
9: Set $\theta_{m,I_i,0}^* = \frac{2\pi(p_{I_i}+1)}{B}$;
10: **end**
11: Let $\boldsymbol{\Theta}_{m,0}^* = [{\rm e}^{{\rm j}\theta_{m,1,0}^*}, \ldots, {\rm e}^{{\rm j}\theta_{m,n,0}^*}, \ldots, {\rm e}^{{\rm j}\theta_{m,N,0}^*}]$ and $H_{\rm opt}^m = \sum_{n=1}^N h_{1,n}h_{2,m,n}{\rm e}^{{\rm j}\theta_{m,n,0}^*}$.

handle, we consider minimizing the union bound of SER [21], i.e., $P_s \leqslant \bar{P}_s = (BQ-1)\mathcal{Q}(\frac{d_{\min}}{\sqrt{2}\sigma})$[1)], where $d_{\min}$ represents the MED and $\mathcal{Q}(\cdot)$ is the Q-function. Since $\mathcal{Q}(\cdot)$ is a decreasing function, minimizing $\bar{P}_s$ is equivalent to maximizing $d_{\min}$. In the rest of this section, we will study the APSK constellation optimization to minimize $d_{\min}$.

4.1 Problem formulation

We consider $L \geqslant 1$ rings on the BQ-ary APSK constellation, where the radius of each l-th ring is denoted as R_l, with $0<R_1\leqslant R_2\leqslant\cdots\leqslant R_L$. We assume that A_l points are uniformly located on each l-th ring, with $\sum_{l=1}^L A_l = BQ$, for which the reference phase of the first point is denoted as $\omega_l \in [0, \frac{2\pi}{A_l})$ without loss of generality. Different from the traditional APSK constellation design where each symbol only carries the information of transmitter [22], each symbol in our considered APSK constellation carries information from both the transmitter and the IRS, thus resulting some new constraints in the optimization problem. Specifically, note that the IRS's information is only embedded in the symbol phases and is independent of the transmitter's information. Thus, we need to ensure that the number of points on each ring is an integer multiple of B, i.e., $A_l = K_lB$, where $l = 1, \ldots, L$ and $K_l \in \mathbb{N}^+$, so as to represent the B possible realizations of the IRS's information. On the other hand, to fully deliver the transmitter's information, there should be Q points in each phase region $[\frac{2\pi b}{B}, \frac{2\pi(b+1)}{B})$ in the constellation. Fortunately, this can be automatically satisfied with $\sum_{l=1}^L A_l = BQ$ and $A_l = K_lB$, for which the proof is simple and thus omitted. Let d_l denote the MED among the points on the l-th ring, which can be expressed as $d_l = R_l\sqrt{2[1-\cos(\frac{2\pi}{A_l})]}$ if $A_l \geqslant 2$ and $d_l = \infty$ otherwise. Let $d_{l,h}$ denote the MED among the points on rings l and h, which is expressed as $d_{l,h} = \sqrt{R_l^2 + R_h^2 - 2R_lR_hC_{l,h}(A_l, A_h, \omega_l, \omega_h)}$ for $l \neq h$, where $C_{l,h}(A_l, A_h, \omega_l, \omega_h) = \max_{t_l,t_h}\cos(\omega_l - \omega_h + \frac{2\pi t_l}{A_l} - \frac{2\pi t_h}{A_h})$, and $t_l \in \{0, 1, \ldots, A_l - 1\}, t_h \in \{0, 1, \ldots, A_h - 1\}$. The overall MED of the constellation is thus given by $d_{\min} = \min_{l,h\in\{1,\ldots,L\},l\neq h}\{d_l, d_{l,h}\}$. Hence, the constellation optimization problem is formulated as

$$
\begin{aligned}
({\rm P2}) \quad \max_{\substack{\{A_l\}_{l=1}^L,\\ \{\omega_l\}_{l=1}^L, \{R_l\}_{l=1}^L}} \quad & d_{\min} && (5)\\
{\rm s.t.} \quad & \sum_{l=1}^L A_l = BQ, && (6)\\
& \frac{A_l}{B} \in \{1, \ldots, Q\},\ l = 1, \ldots, L, && (7)\\
& \omega_l \in \left[0, \frac{2\pi}{A_l}\right), && (8)\\
& 0 < R_1 \leqslant \cdots \leqslant R_L, && (9)\\
& d_{\min} \leqslant \min_{l,h\in\{1,\ldots,L\},l\neq h}\{d_l, d_{l,h}\}, && (10)
\end{aligned}
$$

1) Note that we assume an equiprobable signal set for both the IRS's information in ${\rm e}^{{\rm j}\frac{2\pi b}{B}}$ and the transmitter's information in s, thus each symbol in the BQ-ary APSK occurs with equal probability.

$$\frac{1}{BQ}\sum_{l=1}^{L} A_l R_l^2 \leqslant 1, \tag{11}$$

where Eq. (11) denotes the unit average power constraint of the constellation.

4.2 Optimal solution to problem (P2)

Note that problem (P2) is a non-convex optimization problem due to the integer constraints on A_l and the non-convex constraints in (10). In this subsection, we focus on the practical case of $L = 2$, and propose an effective method to solve the problem (P2)[2)].

First, note that since $R_1 \leqslant R_2$, it can be shown that the optimal solution to A_1 and A_2 should satisfy $A_1 \leqslant A_2$[3)]. Moreover, since $A_1+A_2=BQ$, we can obtain the feasible set of A_1 as $A_1 \in \{B, 2B, \ldots, \frac{Q}{2}B\}$. The optimal solution to A_1 can be found by solving problem (P2) with all feasible A_1's, and finding the value of A_1 with the maximum MED $d_{\min}$. In the following, we focus on solving problem (P2) with given A_1.

With given A_1, we first note that the optimization of ω_1 and ω_2 is independent of R_1 and R_2, which can be expressed as the following problem:

$$\text{(P2-1)} \quad \min_{\omega_1,\omega_2} \quad C_{1,2}(A_1, A_2, \omega_1, \omega_2) \tag{12}$$

$$\text{s.t.} \quad \omega_l \in \left[0, \frac{2\pi}{A_l}\right), \; l = 1, 2. \tag{13}$$

Problem (P2-1) is in a similar form as problem (P2.1) in [22], for which the optimal solution can be efficiently obtained via Algorithm 1 in [22], which is omitted here for brevity. Let ω_1^* and ω_2^* denote the optimal solution to problem (P2-1) and $C_{1,2}^*$ denote its optimal value.

Next, we optimize R_1 and R_2 with the obtained ω_1^* and ω_2^*. Note that under the average power constraint in (11), the optimal radius of the outer ring R_2, can be expressed as $R_2 = \sqrt{\frac{BQ}{A_2} - \frac{A_1}{A_2}R_1^2}$ such that the equality holds in (11), otherwise, the MED can be further enlarged since d_2 and $d_{1,2}$ are both increasing functions of R_2. Therefore, the optimization of R_1 and R_2 is reduced to the optimization of only R_1. In addition, considering the constraint in (15), we obtain the feasible region of R_1 as $0<R_1\leqslant 1$. As such, the problem (P2) reduces to the following problem:

$$\text{(P2-2)} \quad \max_{R_1} \quad \min\left\{d_1(R_1), d_2(R_1), d_{1,2}(R_1)\right\} \tag{14}$$

$$\text{s.t.} \quad 0 < R_1 \leqslant 1, \tag{15}$$

where $d_l(R_1), l = 1, 2$ denotes the MED among the points on ring l, and $d_{1,2}(R_1)$ is the MED among the points on ring 1 and 2 by taking R_1 as variable. Then, we have the following theorem.

Theorem 1. The optimal solution and optimal value for the problem (P2-2) can be obtained as

- $R_1^* = R_{1,2}$, $d_{\min} = d_2(R_{1,2})$, if $d_{1,2}(R_{1,2}) > d_1(R_{1,2})$,
- $R_1^* = R_{1,3}$, $d_{\min} = d_1(R_{1,3})$, if $d_{1,2}(R_{1,2}) \leqslant d_1(R_{1,2})$,

where $R_{1,2}$ is the intersection between $d_1(R_1)$ and $d_2(R_1)$, which is given by

$$R_{1,2} = \sqrt{\frac{[1-\cos(\frac{2\pi}{A_2})]\frac{BQ}{A_2}}{1-\cos(\frac{2\pi}{A_1}) + \frac{A_1}{A_2}[1-\cos(\frac{2\pi}{A_2})]}},$$

and $R_{1,3}$ is the intersection between $d_1(R_1)$ and $d_{1,2}(R_1)$, which is obtained as

$$R_{1,3} = \sqrt{\frac{-b-\sqrt{b^2-4ac}}{2a}},$$

$a=[2\cos(\frac{2\pi}{A_1}) - 1 - \frac{A_1}{A_2}]^2 + 4\frac{A_1}{A_2}{C_{1,2}^*}^2$, $b=\frac{BQ}{A_2}[4\cos(\frac{2\pi}{A_1}) - 2 - 2\frac{A_1}{A_2} - 4{C_{1,2}^*}^2]$ and $c=(\frac{BQ}{A_2})^2$.

2) The extension of our methods to the case with $L > 2$ is left as our future work.

3) The proof is similar to that in [22] and is omitted her for brevity.

Proof. To find the optimal solution to problem (P2-2), we need to analyze the relationship among $d_1(R_1), d_2(R_1)$ and $d_{1,2}(R_1)$. It is easy to see that $d_1(R_1)$ monotonically increases with R_1, while $d_2(R_1)$ monotonically decreases with R_1, and we can prove that $d_{1,2}(R_1)$ decreases with R_1 in the region $R_1 \to 0^+$ and has at most one pole in the feasible domain. The proof is omitted due to the limited space. Observing the relationship among $d_1(R_1), d_2(R_1)$, and $d_{1,2}(R_1)$, we have the following findings: when $R_1 = 0$, we have $d_1(R_1) < d_2(R_1)$ and $d_1(R_1) < d_{1,2}(R_1)$; when $R_1 = R_2 = 1$, we have $d_1(R_1) > d_2(R_1) > d_{1,2}(R_1)$. As such, the relationship between $d_{1,2}(R_{1,2})$ and $d_1(R_{1,2})$ can be divided into two cases, and the results in Theorem 1 can be obtained accordingly.

Based on the optimal solution to the problem (P2) with fixed A_1 derived above, by performing one-dimension searching over all possible A_1s, the optimal solution to the problem (P2) can be obtained. The overall algorithm is summarized in Algorithm 2.

Algorithm 2 Algorithm for solving the problem (P2) when $L = 2$

Input: B and Q.
Output: A_1^*, A_2^*, ω_1^*, ω_2^*, R_1^*, R_2^*.
1: **for** K_1 from 1 to $\frac{Q}{2}$
2: Let $A_1 = BK_1$, $A_2 = BQ - A_1$;
3: Obtaining ω_1^* and ω_2^* by utilizing the method proposed in [22];
4: Obtaining R_1^*, R_2^* and $d_{\min}(K_1)$ by utilizing Algorithm 2;
5: **end**
6: Output the values of A_1^*, A_2^*, ω_1^*, ω_2^*, R_1^*, R_2^* corresponding to the maximum value among $\max[d_{\min}(1), d_{\min}(2), \ldots, d_{\min}(\frac{Q}{2})]$.

5 Summary of the proposed protocol

In this section, we summarize the protocol of the proposed scheme, which can be divided into two phases.

• Preparatory phase. During this phase, the transmitter designs the optimal receiver constellation by utilizing Algorithm 2 and feeds the values of A_1^*, A_2^*, R_1^*, R_2^*, ω_1^*, and ω_2^* to the receiver. Next, at each channel coherence interval, the transmitter estimates the cascaded channel state information (CSI) by using the methods proposed in [23], designs the IRS reference phase shift $\mathbf{\Theta}_{m,0}^*$ for $m = 1, 2, \ldots, M$ via Algorithm 1, and feeds $\mathbf{\Theta}_{m,0}^*$ to the IRS via the control link and $\{H_{\rm opt}^m\}_{m=1}^M$ to the receiver for coherent detection.

• Information transmission phase. During this phase, the IRS and transmitter send their information independently. On receiving the signal, the receiver first decodes $\log_2 M$ bits from the IRS by detecting the selected antenna m via (16), and then decodes $\log_2 B + \log_2 Q$ bits from both the IRS and the transmitter via ML detection of the designed BQ-ary APSK constellation, i.e.,

$$q = \arg \min_{\bar{q}=1,\ldots,BQ} \left| \frac{y_m}{|H_{\rm opt}^m|} - C_{\bar{q}} \right|^2, \tag{16}$$

where $\frac{y_m}{|H_{\rm opt}^m|}$ is the normalization process and $C_{\bar{q}}$ represents the complex coordinates of the point with index $\bar{q}$. The remaining task is to project q into $\log_2 B$ bits IRS symbol and $\log_2 Q$ bits transmitter symbol. By considering the joint modulation strategy, the coding structure of the points on the constellation is special. Specifically, in each phase region $[\frac{2\pi b}{2}, \frac{2\pi(b+1)}{4}), b = 0, 1, \ldots, B$, the IRS symbols are binary form of b, while the transmitter symbols vary in base Q. For example, assuming $B = 4$, $Q = 4$, $L = 2$, $A_1 = 4$ and $A_2 = 12$, the corresponding constellation is presented as in Figure 3, where the number 1–16 represents the index q of each point. Let the structure of the symbols be "abcd", where "ab" represents 2 bits IRS symbol, while "cd" is 2 bits transmitter symbol, the index q and its corresponding symbol of the points are listed in Table 1, based on which, the IRS and transmitter symbols can be decoded.

6 Numerical result

In this section, we provide numerical results to evaluate the performance of the proposed scheme. We consider a Rayleigh fading channel model, with $h_{1,n} \sim \mathcal{CN}\left(0, C_0 d_1^{-\alpha_1}\right)$ and $h_{2,m,n} \sim \mathcal{CN}\left(0, C_0 d_2^{-\alpha_2}\right)$ for all m, n, where $C_0 = -30$ dB denotes the path loss at the reference distance 1 meter (m); $\alpha_1 = 2$ and $\alpha_2 = 3.5$ denote the path loss exponents for the transmitter-IRS and IRS-user links, respectively;

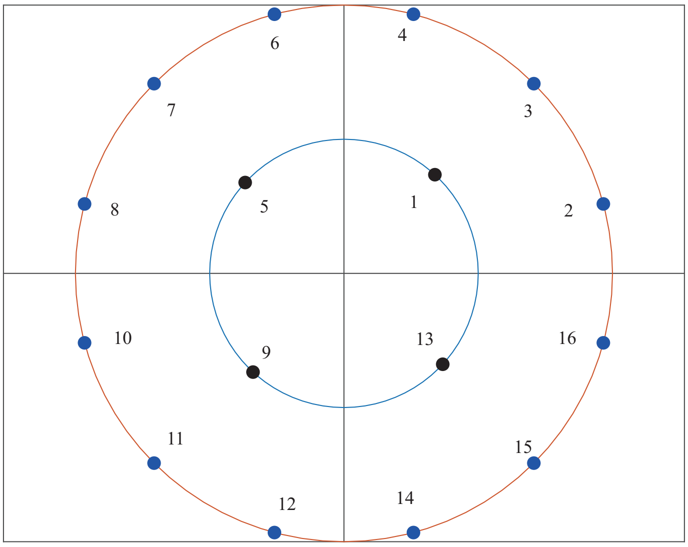

Figure 3 The constellation.

Table 1 The coding table

Index	Symbol	Index	Symbol	Index	Symbol	Index	Symbol
1	0000	5	0100	9	1100	13	1000
2	0001	6	0101	10	1101	14	1001
3	0011	7	0111	11	1111	15	1011
4	0010	8	0110	12	1110	16	1010

$d_1 = 2$ m and $d_2 = 80$ m denote the transmitter-IRS and IRS-user distances, respectively. Note that we have neglected the sizes of the IRS and the receiver antenna array since the transmitter-IRS and IRS-user link distances are sufficiently large. Unless otherwise specified, we set $N = 32$, $B = 4$, $M = 4$, $Q = 4$, and the bandwidth of the system as 180 kHz with the noise power spectral density being -173 dBm/Hz.

In Figure 4, we evaluate the SER performance of the transmitter and IRS symbols under perfect and imperfect CSI. The estimated cascaded CSI is modeled as $\widehat{h}_{m,n} = h_{1,n}h_{2,m,n} + e_{m,n}$, $m = 1, 2, \ldots, M, n = 1, 2, \ldots, N$, where $\widehat{h}_{m,n}$ and $e_{m,n}$ represent the estimated cascaded channel and estimation error associated with the m-th antenna through the n-th IRS element, respectively. In addition, $e_{m,n}$ is modeled as a complex Gaussian random variable with zero mean and variance σ_e^2. We set the variance of the estimation error as $\sigma_e^2 = \eta C_0^2 d_1^{-\alpha_1} d_2^{-\alpha_2}$, where η represents the ratio of error variance to the actual channel power. It is observed that the SER of IRS/transmitter with imperfect CSI decreases with increasing estimation accuracy (which corresponds to smaller η). Moreover, it is observed that $\eta = 0.01$ can already achieve satisfactory performance with only a small gap with the case with perfect CSI, thus validating the effectiveness of the proposed scheme under imperfect CSI.

In Figure 5, we compare the performance of the proposed scheme with two benchmark schemes.

- Benchmark 1. IRS-assisted spatial modulation is proposed in [15], where the IRS modulates its information only in the index of receive antenna.
- Benchmark 2. Passive beamforming and information transfer scheme proposed in [18], where the IRS transmits its information via spatial modulation over the index of the IRS elements. Note that we compare the detectors with the best detection performance proposed in [18], i.e., the BiG-AMP detector is used for the transmitter information and the GAMP detector is used for the IRS information.

For fairness of comparison, we consider the same total number of information bits. Specifically, for benchmark 1, let each transmitter symbol contain $\log_2 Q + \log_2 B$ bits to make up for the additional $\log_2 B$ bits of IRS information in our proposed scheme. For benchmark 2, let the symbol ratio between the transmitter and the IRS be 8, i.e., the transmitter transmits 8 symbols during each IRS symbol time. In addition, since the number of information bits contained in the transmitter/IRS symbols may be different for each scheme, we consider the bit error rate (BER) for both IRS and transmitter as the performance metric.

It is observed from Figure 5 that the BER of the transmitter information for the proposed scheme

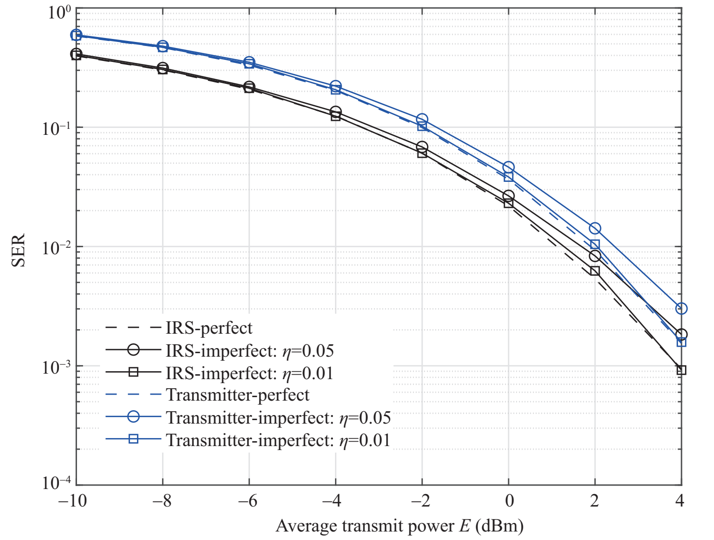

Figure 4 Performance of the proposed scheme with perfect and imperfect CSI.

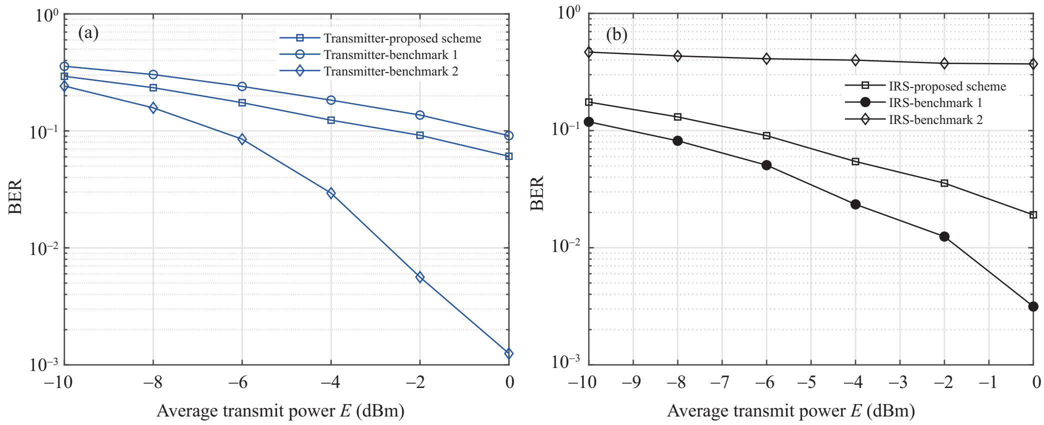

Figure 5 BER comparison between the proposed scheme with benchmarks. (a) BER of transmitter information; (b) BER of IRS information.

outperforms that of the benchmark 1. Moreover, the BER of the IRS information for benchmark 2 suffers from a severe error floor, since the number of receive antennas is relatively small compared with the number of IRS reflecting elements ($N = 32$), thus the index of the IRS reflecting element cannot be accurately detected. On the other hand, the BER of the transmitter's information for benchmark 2 outperforms that for our proposed scheme, but at a cost of high complexity. In conclusion, the proposed scheme is competitive compared with these existing schemes regarding either BER or complexity.

7 Conclusion

The paper studies an IRS-aided SIMO system, where the IRS assists the primary system as well as delivers its own information. To enhance the received signal power of the primary system, we proposed a discrete phase shift design with $\mathcal{O}(N)$ complexity. To convey the information of the transmitter and the IRS simultaneously, we proposed a novel modulation strategy to avoid signal aliasing. Based on the proposed strategy, we derived the optimal constellation design to minimize the MED based on the APSK structure. Numerical results validated the effectiveness of the proposed scheme with imperfect CSI, and showed that the proposed scheme is competitive compared with several benchmark schemes.

Acknowledgements This work was supported in part by National Key R&D Program of China (Grant No. 2019YFB1803400) and Zhejiang Provincial Natural Science Foundation of China (Grant No. LD21F010001).

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