Computer Life
ISSN: 1819-4818 | Vol. 11, No. 2, 2023

Research on resource allocation of IRS-assisted NOMA system

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Abstract: The transmission scheme of intelligent reflecting surface (IRS) assisted non-orthogonal multiple access (NOMA) downlink multi-cluster system is studied. In this paper, the non-convex optimization problem of maximizing the sum rate of each cluster of users is solved by jointly optimizing the successive interference cancellation (SIC) decoding sequence, power allocation and IRS phase shift, while considering the quality of service of each user. Firstly, the SIC decoding order is determined according to the difference of the combined channel gain. Then, the optimization problem is decoupled into two sub-problems, and the Lagrange function is alternately used for power allocation and semidefinite relaxation (SDR) for phase shift optimization until the optimization objective converges. The simulation results show that compared with other transmission schemes of IRS-NOMA system, the proposed optimization scheme can significantly improve the sum rate of the system.

Keywords: Intelligent reflect surface; Non-orthogonal multiple access; Power allocation; Phase shift optimization.

1. Introduction

Intelligent reflective surface (IRS) is an emerging radio frequency wireless communication technology. IRS is composed of a large number of passive reflection units. These units can change their electromagnetic characteristics according to the input signal, so as to realize the effective control and reconstruction of electromagnetic waves. Unlike traditional programmable antennas and smart surfaces, IRS has the ability of autonomous decision-making, adaptive adjustment and autonomous learning. By adjusting the parameters such as reflection coefficient, phase and amplitude, IRS can achieve accurate adjustment and control of electromagnetic waves, thereby optimizing the quality, capacity and coverage of wireless signal transmission [1]-[4]. IRS has become the focus of wireless communication research to alleviate the challenges encountered in wireless networks.

Non-orthogonal multiple access (NOMA) technology can allocate resources non-orthogonally, which can expand the connection scale and improve the spectrum utilization. By applying superposition coding and successive interference cancellation (SIC) at the receiver, multiple users can be multiplexed on the same sub-channel [5]-[8]. In [9], the sub-channel energy efficiency optimization and power allocation problem in the downlink of NOMA heterogeneous networks are studied. In [10], a power allocation strategy with energy efficiency optimization is proposed under the condition of satisfying the user's service quality.

NOMA can use the user's dynamic channel conditions to improve spectrum efficiency and user fairness. IRS can reconstruct the propagation environment of mobile users, so that the user 's data rate and reception reliability are significantly improved. Therefore, the integration of the two technologies has attracted the attention of academia and industry. In [11], the problem of maximizing the sum rate of the uplink IRS-NOMA system is studied. By establishing the relationship between the user 's transmission power and the IRS phase shift, the semi-definite relaxation (SDR) is used to solve the problem. At the same time, the influence of the number of IRS reflection elements on the sum rate is revealed. In [12], the problem of minimizing the downlink transmission power of an IRS-assisted NOMA system is studied. A new difference-of-convex (DC) programming algorithm is proposed. By transforming the quadratic programming into a rank-1 constrained matrix optimization problem, and further expressing the non-convex rank function as a DC function, the obtained non-convex quadratic programming can be effectively solved. In Reference [13], researchers proposed a sub-optimal algorithm combining block coordinate descent (BCD) and SDR to jointly optimize the power allocation of BS and the phase shift of IRS, so as to maximize the minimum signal-to-interference-plus-noise-ratio (SINR) of the decoding signal at the user side. In order to better integrate IRS and NOMA technology, the research work of this paper is as follows:

For the IRS-assisted NOMA downlink multi-user transmission model, considering the minimum rate requirement of each cluster user and the maximum transmission power constraint of BS, a non-convex problem based on the sum rate maximization under the model is established;

In order to solve the non-convex optimization problem of coupling optimization variables. The optimization problem is divided into two stages: the first stage uses the semi-definite relaxation method to determine the SIC decoding order according to the difference of the combined channel gain; in the second stage, the problem is decomposed into two sub-problems: IRS phase optimization and BS power allocation. The objective function and constraints are transformed by Lagrange function, relaxation variable, SCA and other methods. The sub-problem is transformed into a convex problem to solve;

Numerical simulations show that we verify the effectiveness of the proposed joint SIC decoding order, BS power allocation, and IRS phase shift optimization algorithm. At the same time, compared with other benchmark schemes, the proposed algorithm can effectively improve the sum rate of the system.

2. System Model and Problem Modeling

The research scenario in this paper is the IRS-assisted NOMA downlink system. The model is shown in the figure. The system has a single antenna BS and 2K (K > 2) single antenna users. The users are divided into K clusters, and each cluster contains two users. BS transmits signals to 2K users through direct channel and IRS reflection channel on the same time-frequency resource block. Assuming that the distance from BS to user k is $d_{B,k}$, the distance from BS to IRS is $d_{B,R}$, and the distance from IRS to user k is $d_{R,k}$, BS can simultaneously receive the information transmitted by the direct channel and the reflected channel.

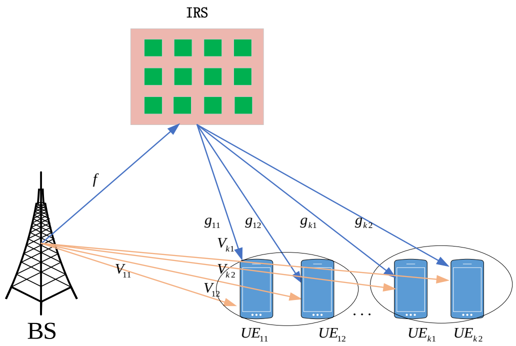

Figure1: System model diagram

Assuming that the channel state information (CSI) is known, the channel model consists of two parts: path loss and small-scale fading. Assuming that the direct channel with BS users is modeled as Rayleigh fading, the channel between BS and user k can be expressed as:

$$V_k = d_{B,k}^{-\alpha_{B,k}/2} b_1 \tag{1}$$

where $\alpha_{B,k}$ denotes the path loss exponent between the base station and user k, b_1 denotes the small-scale fading which obeys a complex Gaussian distribution with mean 0 and variance 1 denoted as $b_1 \sim \mathcal{CN}(0,1)$.

It is assumed that there is a Line-of-Sight (LoS) link between BS and RIS and between RIS and user k. Therefore, the channel can be modeled as Rician fading. Then the channel vector between BS and RIS and the channel amount between RIS and user k can be expressed as:

$$f = d_{B,R}^{-\alpha_{B,R}/2}\left(\sqrt{\frac{K_{B,R}}{1+K_{B,R}}} b_2^{LoS} + \sqrt{\frac{1}{1+K_{B,R}}} b_2\right) \tag{2}$$

$$g_k = d_{R,k}^{-\alpha_{R,k}/2}\left(\sqrt{\frac{K_{R,k}}{1+K_{R,k}}} b_3^{LoS} + \sqrt{\frac{1}{1+K_{R,k}}} b_3\right) \tag{3}$$

where $f \in \mathbb{C}^{M\times 1}$, $g_k \in \mathbb{C}^{N\times 1}$, $\alpha_{B,R}$ and $\alpha_{R,k}$ are the path loss exponents from BS to IRS and from IRS to user k, respectively. $K_{B,R}$ and $K_{R,k}$ are the Rician factors, b_2^{LoS} and b_3^{LoS} are the LoS components, and b_2 and b_3 are the non-line-of-sight link components of the channel.

Due to the huge path loss caused by multiple reflections, this paper only considers the signal reflected by the IRS once, so the user UE_{k1} and UE_{k2} receive signals as follows:

$$y_{k1} = \left(g_{k1}^H \Theta f + V_{k1}\right)\left(\sqrt{p_{k1}} x_{k1} + \sqrt{p_{k2}} x_{k2}\right) + n_{k1} \tag{4}$$

$$y_{k2} = \left(g_{k2}^H \Theta f + V_{k2}\right)\left(\sqrt{p_{k2}} x_{k2} + \sqrt{p_{k1}} x_{k1}\right) + n_{k2} \tag{5}$$

x_{k1} and x_{k2} are the expected received signals of users UE_{k1} and UE_{k2}, respectively. The power allocated by the base station for the Kth cluster is P_k, and the power allocated for users UE_{k1} and UE_{k2} is p_{k1} and p_{k2}. Θ are the diagonal phase shift matrix of IRS, $\Theta = \mathrm{diag}\{w_1, \cdots w_n, \cdots w_N\}$, where $w_n = e^{j\theta}$ and $\theta_n \in [0, 2\pi)$ represent the phase shift of the nth reflection unit. n_{k1} and n_{k2} are the Gaussian white noise received by users UE_{k1} and UE_{k2}, respectively, and their unilateral power spectral density is N_0.

The user UE_{k1} first detects the expected signal UE_{k2} of x_{k2} and eliminates the interference of x_{k2} to y_{k1}, and then detects its own expected signal x_{k1}, UE_{k1} decoding x_{k2} signal-to-interference-plus-noise ratio is:

$$r_{k1\to k2} = \frac{p_{k2}\left|g_{k1}^H \Theta f + V_{k1}\right|^2}{p_{k1}\left|g_{k1}^H \Theta f + V_{k1}\right|^2 + \sigma_{k1}^2} \tag{6}$$

After removing the interference of x_{k2} to y_{k1}, the signal-to-interference-plus-noise ratio of decoding x_{k1} is:

$$r_{k1\to k1} = \frac{p_{k1}\left|g_{k1}^H \Theta f + V_{k1}\right|^2}{\sigma_{k1}^2} \tag{7}$$

UE_{k2} is a long-distance user, which directly decodes its own expected signal x_{k2}, and the signal-to-interference-plus-noise ratio is:

$$r_{k2\to k2} = \frac{p_{k2}\left|g_{k2}^H \Theta f + V_{k2}\right|^2}{p_{k1}\left|g_{k2}^H \Theta f + V_{k2}\right|^2 + \sigma_{k2}^2} \tag{8}$$

The rates of users UE_{k1} and UE_{k2} are:

$$R_{k1} = B\log_2\left(1 + \frac{p_{k1}\left|g_{k1}^H \Theta f + V_{k1}\right|^2}{\sigma_{k1}^2}\right) \tag{9}$$

$$R_{k2} = B\log_2\left(1 + \frac{p_{k2}\left|g_{k2}^H \Theta f + V_{k2}\right|^2}{p_{k1}\left|g_{k2}^H \Theta f + V_{k2}\right|^2 + \sigma_{k2}^2}\right) \tag{10}$$

The power allocation optimization problem model of maximizing the total rate is established. The power allocation objective is to maximize the total rate of the system under the given total power and the rate requirement of each user:

$$
\begin{aligned}
P(1) \quad & \max_{P_k,\Theta} \sum_{k=1}^{K} \left(R_{k1} + R_{k2} \right) \\
s.t \quad & C1: \sum_{k=1}^{K} \left(p_{k1} + p_{k2} \right) = P_{\max} \\
& C2: R_{k1} \ge r_{k1\min}, \forall k \\
& C3: R_{k2} \ge r_{k2\min}, \forall k \\
& C4: r_{k1\to k2} \ge r_{k2\to k2}, \forall k \\
& C5: \left| w_n \right| = 1, 1 \le n \le N, w_n = e^{j\theta} \\
& C6: s(k) \in \Omega
\end{aligned} \tag{11}
$$

C1 denotes the total power of the base station is $P_{\max}$, C2 denotes the minimum rate requirement of the user UE_{k1}, C3 denotes the minimum rate requirement of the user UE_{k2}, C4 denotes the signal-to-interference-plus-noise ratio requirement of the user UE_{k1} decoding x_{k2}, C5 denotes the modulus of the limited phase shift w_n is 1, and Ω in C6 denotes the set of SIC decoding sequences.

Since the optimization problem P (1) is non-convex and the optimization variables are coupled with each other, it is difficult to solve directly. Therefore, the objective optimization is divided into two stages: the first stage is to determine the SIC decoding order based on maximizing the combined channel gain difference; in the second stage, according to the decoding order determined in the first stage, the problem is decoupled into two sub-problems to be solved alternately.

3. Algorithm Design

3.1. SIC decoding order determination algorithm

In the NOMA system, the channel difference of the users in the sub-channel determines the upper limit of the NOMA performance. The greater the channel gain difference between the two users, the more obvious the throughput of the NOMA system relative to the OMA system. In the IRS-assisted NOMA system, the channel gain difference between two users in the sub-channel depends not only on the channel gain of the direct link, but also on the channel gain of the reflected link controlled by the IRS. In order to avoid the high complexity caused by exhaustive search, the difference method of maximizing the channel gain between all packet users is adopted. The established problem is:

$$
\begin{aligned}
P(2) \quad & \max_{\Theta} \sum_{k=1}^{K} \left| \left| g_{k1}^{H} \Theta f + V_{k1} \right|^2 - \left| g_{k2}^{H} \Theta f + V_{k2} \right|^2 \right| \\
s.t \quad & C1: \left| \left| g_{k1}^{H} \Theta f + V_{k1} \right|^2 - \left| g_{k2}^{H} \Theta f + V_{k2} \right|^2 \right| \ge r \\
& C2: \left| w_n \right| = 1, 1 \le n \le N, w_n = e^{j\theta}
\end{aligned} \tag{12}
$$

The channel gain is identically transformed $w = diag\{\Theta\} = \left[w_1, w_2, \cdots w_N \right]$,let $h_{k1} = diag\left\{ g_{k1}^{H} \right\} f$, $h_{k2} = diag\left\{ g_{k2}^{H} \right\} f$, $\left| g_{k1}^{H} \Theta f + V_{k1} \right|^2 = \left| h_{k1}^{H} w + V_{k1} \right|^2$, $\left| g_{k2}^{H} \Theta f + V_{k2} \right|^2 = \left| h_{k2}^{H} w + V_{k2} \right|^2$,

Definition $w = \left[w \ \ 1 \right]^T$, $h_{k1} = \left[h_{k1} \ \ V_{k1}^{*} \right]^T$, $h_k = \left[h_{k2} \ \ V_{k2}^{*} \right]^T$ The auxiliary matrix B_1 , B_2 W is introduced, $B_1 = \sum_{k=1}^{K} h_{k1} h_{k1}^{\ \ H}$, $B_2 = \sum_{k=1}^{K} h_{k2} h_{k2}^{\ \ H}$, Thus the P (2) problem is converted to P (3) :

$$
\begin{aligned}
P(3) \quad & \max_{\Theta} \left[w^{H} B_1 w - w^{H} B_2 w \right] \\
s.t \quad & C1: \left[w^{H} B_1 w - w^{H} B_2 w \right] \ge r \\
& C2: \left| w_n \right| = 1, 1 \le n \le N, w_n = e^{j\theta}
\end{aligned} \tag{13}
$$

Since the constant modulus constraint P (3) of a specific element is still a non-convex problem, the P (3) problem is solved, and the constraint C2 is relaxed to the sphere constraint $\left| w \right|_2 = N$. The P (3) problem is transformed into P (4):

$$
\begin{aligned}
P(4) \quad & \max_{\Theta} \left[w^{H} \left(B_1 - B_2 \right) w \right] \\
s.t \quad & C1: \left[w^{H} \left(B_1 - B_2 \right) w \right] \ge r \\
& C2: \left| w \right|_2 = N
\end{aligned} \tag{14}
$$

Here $\left| w \right|_2$ denotes the Euclidean norm of w , so that the $W = ww^{H}$, P(4) problem can be converted to P (5):

$$
\begin{aligned}
P(5) \quad & \max_{\Theta} Tr\left[\left(B_1 - B_2 \right) W \right] \\
s.t \quad & C1: Tr\left[\left(B_1 - B_2 \right) W \right] \ge r \\
& C2: diag\left\{ W \right\} = 1 \\
& C3: W \ ? \ 0 \quad 0 \ \ 0 \\
& C4: 0ank\left(W \right) = 10
\end{aligned} \tag{15}
$$

C3 indicates that W is a symmetric positive semidefinite matrix, and C4 indicates that the rank limit of W is one. By relaxing the constraint C3, the semidefinite programming (SDR) problem can be solved by using the CVX toolbox. However, the obtained W^{*} may not satisfy the rank 1 constraint. Therefore, the Gaussian random technique is used to obtain its approximate solution, and then W^{*} is obtained.

3.2. Optimal power allocation

Given the IRS phase shift matrix Θ , the power optimization allocation is carried out. The optimization problem is:

$$P(6) \quad \max_{P_k} \sum_{k=1}^{K} \left(R_{k1} + R_{k2} \right)$$

$$\begin{aligned} s.t \quad & C1: \sum_{k=1}^{K} \left(p_{k1} + p_{k2} \right) \le P_{\max} \\ & C2: R_{k1} \ge r_{k1\min}, \forall k \\ & C3: R_{k2} \ge r_{k2\min}, \forall k \\ & C4: r_{k1 \to k2} \ge r_{k2 \to k2}, \forall k \end{aligned} \tag{16}$$

From C2 and C3 in P (6), we can obtain:

$$p_{k1} \ge \frac{a_{k1}\sigma_{k1}^2}{\left| g_{k1}^H \Theta f + V_{k1} \right|^2} \tag{17}$$

$$p_{k2} \ge a_{k2} p_{k1} + \frac{a_{k2}\sigma_{k2}^2}{\left| g_{k2}^H \Theta f + V_{k2} \right|^2} \tag{18}$$

where $a_{k1} = 2^{\frac{r_{k1\min}}{B}} - 1$, $a_{k2} = 2^{\frac{r_{k2\min}}{B}} - 1$:

When (18) holds, C4 in P (6) must hold, so when p_{k1} and p_{k2} satisfy C2, C3, C4 in (17) and (18) P (6) must hold.

Let p_{k0} denote the minimum power required for the K cluster.

$$p_{k0} \ge \frac{a_{k1}\sigma_{k1}^2}{\left| g_{k1}^H \Theta f + V_{k1} \right|^2} + \frac{a_{k1}a_{k2}\sigma_{k1}^2}{\left| g_{k1}^H \Theta f + V_{k1} \right|^2} + \frac{a_{k2}\sigma_{k2}^2}{\left| g_{k2}^H \Theta f + V_{k2} \right|} \tag{19}$$

Each cluster has a minimum power, and the total power $P_{\max}$ is required to meet:

$$P_{\max} \ge \sum_{k=1}^{K} \left[\frac{a_{k1}\sigma_{k1}^2}{\left| g_{k1}^H \Theta f + V_{k1} \right|^2} + \frac{a_{k1}a_{k2}\sigma_{k1}^2}{\left| g_{k1}^H \Theta f + V_{k1} \right|^2} + \frac{a_{k2}\sigma_{k2}^2}{\left| g_{k2}^H \Theta f + V_{k2} \right|} \right] \tag{20}$$

Since $P_k = p_{k1} + p_{k2}$, when p_{k2} takes the minimum value that satisfies the condition, p_{k1} is the largest.

$$p_{k1} = \frac{p_k - \dfrac{a_{k2}\sigma_{k2}^2}{\left| g_{k2}^H \Theta f + V_{k2} \right|^2}}{1 + a_{k2}} \tag{21}$$

$$p_{k2} = \frac{a_{k2} p_k + \dfrac{a_{k2}\sigma_{k2}^2}{\left| g_{k2}^H \Theta f + V_{k2} \right|^2}}{1 + a_{k2}} \tag{22}$$

Let $b_k = \left| g_{k1}^H \Theta f + V_{k1} \right|^2 \left| g_{k2}^H \Theta f + V_{k2} \right|$, $c_k = \sigma_{k1}^2 \left(\left| g_{k2}^H \Theta f + V_{k2} \right|^2 + a_{k2} \left| g_{k1}^H \Theta f + V_{k2} \right|^2 \right)$, $d_k = a_{k2}\sigma_{k2}^2 \left| g_{k1}^H \Theta f + V_{k1} \right|$, Then the optimization problem becomes:

$$P(7) \quad \max_{P_k, \Theta} \sum_{k=1}^{K} \left[B\log_2 \left(1 + \frac{b_k p_k - d_k}{c_k} \right) + B\log_2 \left(1 + a_{k2} \right) \right] \tag{23}$$

$$\begin{aligned} s.t \quad & C1: \sum_{k=1}^{K} \left(p_{k1} + p_{k2} \right) = P_{\max} \\ & C2: p_{k0} \ge \frac{a_{k1}\sigma_{k1}^2}{\left| g_{k1}^H \Theta f + V_{k1} \right|^2} + \frac{a_{k1}a_{k2}\sigma_{k1}^2}{\left| g_{k1}^H \Theta f + V_{k1} \right|^2} + \frac{a_{k2}\sigma_{k2}^2}{\left| g_{k2}^H \Theta f + V_{k2} \right|}, \forall \end{aligned}$$

The constraint condition C1 indicates that the total power of the base station is $P_{\max}$; the constraint condition C2 represents that the power allocated to the k-th cluster cannot be lower than the minimum power required by the cluster, otherwise it cannot meet the rate requirements of users in the k-th cluster.

The P (8) problem is solvable when (20) is satisfied, otherwise the P (8) problem is not solvable. Constructing Lagrange function：

$$F\left(p_k, \lambda, k = 1, 2, \cdots K \right) = \sum_{k=1}^{K} \left[= B\log_2 \left(1 + \frac{b_k p_k - d_k}{c_k} \right) + B\log_2 \left(1 + a_{k2} \right) \right] + \lambda \left(\sum_{k=1}^{K} p_k - P_{\max} \right) \tag{24}$$

The first-order partial derivatives of $\left(p_k, \lambda, k = 1, 2, \cdots K \right)$ with respect to p_k and λ are calculated respectively, and let them be equal to $k = 0, 1, 2, \cdots K$

$$\frac{b_k}{\left(c_k + b_k p_k - d_k \right) \ln 2} + \lambda = 0 \tag{25}$$

$$\sum_{k=1}^{K} p_k - P_{\max} = 0 \tag{26}$$

Thus, p_{k1} and p_{k2} can be obtained：

$$p_{k1} = \frac{P_{\max} - \sum_{i=2}^{K} \left(\dfrac{-b_i d_1 + b_1 d_i - b_1 c_i + b_i c_1}{b_1 b_i} \right)}{K} \tag{27}$$

$$p_{k2} = p_{k1} + \frac{-b_i d_1 + b_1 d_i - b_1 c_i + b_i c_1}{b_1 b_i} \tag{28}$$

3.3. Phase optimization

After obtaining the optimal solution p_k through power allocation optimization, the power allocated to the user is fixed and the IRS phase shift matrix is optimized. At this time, the optimization problem becomes：

$$P(8) \quad \max_{P_k, \Theta} \sum_{k=1}^{K} \left(R_{k1} + R_{k2} \right)$$

$$\begin{aligned} s.t \quad & C1: R_{k1} \ge r_{k1\min}, \forall k \\ & C2: R_{k2} \ge r_{k2\min}, \forall k \\ & C3: r_{k1 \to k2} \ge r_{k2 \to k2}, \forall k \\ & C4: \left| w_n \right| = 1, 1 \le n \le N, w_n = e^{j\theta} \end{aligned} \tag{29}$$

Obviously, due to the non-convexity of the objective function and the specific element constant modulus constraint, the P (8) problem is still a non-convex problem and difficult to solve. Let

$w=\left[e^{j\theta_1},e^{j\theta_2},\cdots e^{j\theta_N}\right]^H$, $\bar{w}=\left[1,w^H\right]^H$, $W=\bar{\mathrm{w}}\bar{\mathrm{w}}^H$, $\lambda_{k1}=1/\sigma_{k1}^2$, $\lambda_{k1}=1/\sigma_{k2}^2$, At the same time, the channel gain is identically deformed：

$$H_{k1}=\begin{bmatrix} V_{k1} \\ diag\left\{g_{k1}^H\right\}f \end{bmatrix}\begin{bmatrix} V_{k1} \\ diag\left\{g_{k1}^H\right\}f \end{bmatrix}^H,$$

$$H_{k2}=\begin{bmatrix} V_{k2} \\ diag\left\{g_{k2}^H\right\}f \end{bmatrix}\begin{bmatrix} V_{k2} \\ diag\left\{g_{k2}^H\right\}f \end{bmatrix}^H$$

We can get：

$$\bar{r}_{k1}=\log_2\left[1+\lambda_{k1}p_{k1}Tr\left(H_{k1}W\right)\right] \tag{30}$$

$$\bar{r}_{k2}=\ln\left[\lambda_{k2}\left(p_{k1}+p_{k2}\right)Tr\left(H_{k2}W\right)+1\right]-\ln\left[\lambda_{k2}p_{k1}Tr\left(H_{K2}W\right)+1\right] \tag{31}$$

The first-order Taylor expansion of (31) is carried out, and its global linear lower bound is obtained :

$$\ln\left[\lambda_{k2}\left(p_{k1}+p_{k2}\right)Tr\left(H_{k2}W\right)+1\right]-\left\{\ln\left[\lambda_{k2}p_{k1}Tr\left(H_{K2}\tilde{W}\right)+1\right]+\frac{\lambda_{k2}p_{k1}Tr\left[H_{k2}\left(W-\tilde{W}\right)\right]}{\lambda_{k2}p_{k1}Tr\left(H_{k2}\tilde{W}\right)+1}\right\}$$

Resolving the constraint C2 can be obtained：

$$\lambda_{k2}p_{k2}Tr\left(H_{k2}W\right)\geq\left(2^{r_{k2\min}}-1\right)\left[\lambda_{k2}p_{k1}Tr\left(H_{k2}W\right)+1\right] \tag{33}$$

Transform the constraint C3：

$$\ln\left[\lambda_{k2}p_{k1}Tr\left(H_{k2}W\right)+1\right]\geq\ln\left[\lambda_{k2}p_{k2}Tr\left(H_{k2}W\right)\right]+\ln\left[\lambda_{k1}p_{k1}Tr\left(H_{k1}W\right)+1\right] \tag{34}$$

(34) is still not a convex set. Using the first-order Taylor approximation in SCA to solve this problem, it can be obtained that：

$$\begin{aligned}&\ln\left[\lambda_{k1}p_{k2}Tr\left(H_{k1}W\right)\right]+\ln\left[\lambda_{k2}p_{k1}Tr\left(H_{k2}W\right)+1\right]\geq\ln\left[\lambda_{k2}p_{k2}Tr\left(H_{k2}\tilde{W}\right)\right]+\\&\ln\left[\lambda_{k1}p_{k1}Tr\left(H_{k1}\tilde{W}\right)+1\right]+\frac{Tr\left[H_{k2}\left(W-\tilde{W}\right)\right]}{Tr\left(H_{k2}\tilde{W}\right)}+\frac{\lambda_{k1}p_{k1}Tr\left[H_{k1}\left(W-\tilde{W}\right)\right]}{\lambda_{k1}p_{k1}Tr\left[H_{k1}\tilde{W}\right]+1}\end{aligned} \tag{35}$$

The optimization problem becomes：

$$\begin{aligned}P(9)\quad &\max_{W}\sum_{k=1}^{K}\left(\bar{r}_{k1}+\bar{r}_{k2}\log_2^e\right)\\ s.t\quad &(33),(35)\\ &C1:\bar{R}_{k1}\geq r_{k1\min},\forall k\\ &C2:diag\left(W\right)=1\\ &C3:W\pm 0\\ &C4:rank\left(W\right)=1\end{aligned} \tag{36}$$

By relaxing the constraint C3, the SDR problem can be solved by using the CVX toolbox. The specific algorithm is as follows

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Algorithm 1: Alternating Optimization for Solving

Solving the problem P (5), the decoding order Ω is obtained.

Input: Channel matrices $\mathbf{g}_{11},\cdots,\mathbf{V}_{11},\cdots,$ and $\mathbf{f}$;maximum transmit power $P_{\max}$;noise power σ^2 ;loss factor κ .

Output: optimized RIS-NOMA Precoding matrix $\Theta^{(t)}$; Optimal power allocation p_k .

1:Initialize $\Theta^{(0)}$, $P_K^{(0)}$,Maximum number of iterations $T_{\max}$,convergence precision $\varepsilon>0$, iteration times $t=0$, and sum rate $R_{sum}^{t=0}$.

2: Repeat

3:with given Θ^{opt} ,find the optimal $P_K^{(t)}$ according to P(3).

4:with given $P_K^{(t)}$,find the optimal $\Theta^{(t)}$ according to P(6).

5.Updata $t=t+1$.

6:until $\left|R_{sum}^{t}-R_{sum}^{t-1}\right|\leq\varepsilon$,or $t>T_{\max}$.

7:reture Θ^{opt} , P_K^{opt} .

“显示文本”不能横跨多行!

4. Simulation Results Analysis

In this section, we give the simulation results to show the proposed transmission scheme. Assuming that BS is located at (-3,0,0) of the three-dimensional plane, four clusters of eight users are randomly distributed in a circle with a radius of four centers (-3,10). The other parameters are set as follows: noise power of σ^2=144dBm, The Rylance factor is 4, precision $\varepsilon=10^{-4}$.

figure 2 shows the change of sum achievable rate performance relative to the number of IRS reflection units. The maximum transmission power $P_{\max}=15dbm$ of BS and the minimum communication SINR of each user are set as $r_k^{\min}=5dB$ .From Figure 2, it can be seen that the sum achievable rate value increases with the increase of the number of reflection elements of IRS, which is because more reflection units can generate stronger channel gain, thereby improving the system sum rate. The proposed IRS-NOMA transmission scheme is superior to other reference schemes. Compared with the non-IRS-assisted NOMA and IRS-assisted OMA schemes, the transmission scheme proposed in this paper can achieve better sum achievable rate performance, which shows the importance of deploying IRS in the communication system and optimizing the phase shift of IRS. The sum achievable rate performance gap between IRS-NONA and IRS-OMA schemes is because NOMA has higher spectral efficiency than OMA.

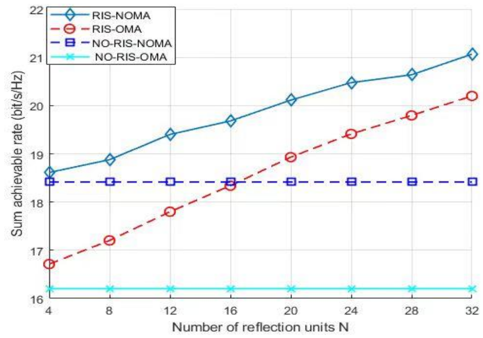

Fig2: Relationship between sum achievable rate and reflection unit

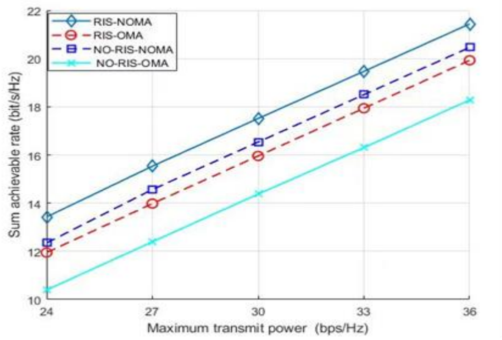

Fig3: Comparison of transmit power and sum achievable rate

Figure 3 shows the transformation between the transmit power and the sum achievable rate. The number of IRS reflection units is set to $N = 12$. It can be seen from Fig.4 that the IRS-NOMA transmission scheme proposed in this paper is always superior to the other three schemes. In the IRS-NOMA and IRS-OMA schemes, the optimized IRS improves the channel gain, and does not and will not reduce the system sum rate while improving the minimum rate of weak users, which greatly ensures that the QoS of weak users in each cluster is satisfied.

5. Conclusion

Aiming at the IRS-assisted NOMA system, this paper proposes a system sum rate maximization problem that jointly optimizes IRS phase shift and power allocation. Considering the maximum transmit power of the base station, IRS phase shift and SIC decoding sequence constraints, a system sum rate maximization model is established. In order to solve the non-convex optimization problem, the SIC decoding order is first determined based on the maximum gain difference of the combined channel, and then the optimization problem is decoupled into two sub-problems for alternating optimization. The original problem is transformed into a convex problem and solved by using relaxation variables, SCA and other methods. The simulation results show that the proposed algorithm can effectively improve the sum rate of the system compared with the benchmark scheme.

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