Sum Rate Maximization for Multiuser MISO Downlink with Intelligent Reflecting Surface

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Abstract—The intelligent reflecting surface (IRS) is a planar array with massive reconfigurable passive elements, which can align the reflecting signals at the receivers via controlling the phase shifts at each element independently. Since IRS can be implemented without RF chain, it is seen as a cost-effective solution for boosting the spectral efficiency for the future communication system. This work considers the IRS-aided multiuser multi-input single-output downlink transmission, and our goal is to maximize the sum rate by jointly optimizing the transmit beamformer at the base station and the (continuous/discrete) phase shifts at the IRS. This sum rate maximization (SRM) problem is challenging, especially for the discrete-phase case. To tackle it, we first derive a more tractable equivalent formulation of the SRM problem with structured convex constraints and a smooth objective. From there, a custom-derived block-coordinated accelerated projected gradient algorithm is developed. Simulation results demonstrate that the proposed design outperforms state-of-the-art design in both sum rate performance and the running time, when the size of the IRS is large.

Index Terms—Intelligent reflecting surface, passive beamforming, accelerated projected gradient

I. INTRODUCTION

Low-cost, energy efficient and high spectral efficiency transmission techniques are indispensable for future communication system. To this end, the intelligent reflecting surface (IRS) has recently been put forward and gained considerable attention. The idea of IRS is to deploy a large planar array to intentionally create additional reflecting paths for the receivers. More specifically, the IRS consists of a large number of passive elements, each of which is capable of independently reflecting the incident electromagnetic wave with certain phase shifts. Therefore, by intelligently adjusting the reflection phase shift of each element, IRS is able to adapt to wireless channels, so that the reflected signals can be constructively (resp. destructively) aligned at the desired (resp. non-intended) receiver.

While currently IRS is still in its infant stage, there has been a flow of researches on incorporating IRS into the existing communication systems [1]–[8]. In particular, the works [1] and [2] considered the joint active and passive beamforming to minimize the transmit power at the based station (BS). The sum rate maximization problem is studied in [7] and [8], where the former focused on the single-user case, and the latter on the multiple users’ case. Besides the conventional communications, the use of IRS in improving the physical-layer security has also been considered in [3] and [4]. We should mention that early works on IRS usually assume that the reflection elements are implemented with infinite resolution phase shifters, that is, the phase shifts of the reflection elements can vary continuously from 0 to 2π. However, in practice it could be costly to achieve a continuous phase shifter due to the hardware limitations. As such, some recent studies on IRS have taken into account the finite-resolution phase shifters. The works [5] and [6] considered the transmit power minimization problem for IRS with finite discrete-phase shifts and employed the alternating optimization technique to obtain an approximate solution.

In this work, we consider the IRS-aided multiuser multi-input and single-output (MISO) downlink; see Fig. 1 where the system sum rate is maximized by jointly optimizing the transmit beamformer at the BS and the (continuous/discrete) phase shifts at the IRS. This sum rate maximization (SRM) problem is nonconvex, and furthermore a mixed integer nonlinear program (MINP) when the discrete-phase shifts are concerned. To tackle it, we first derive an equivalent formulation of the SRM problem by using the recent advances in discrete-phase optimization [9]. The advantage of the reformulation is that it has simple structured convex constraints and a smooth objective function. Based on the reformulation, a block-coordinated accelerated projected gradient (APG) algorithm is proposed. Simulation results demonstrate that the proposed algorithm is able to achieve better sum rate with lower complexity than state-of-the-art method, when the size of the IRS is large.

II. SYSTEM MODEL AND PROBLEM STATEMENT

Consider a multiuser MISO downlink transmission with the aid of an IRS; see Fig. 1. The BS equipped with $M$
transmit antennas unicasts $K$ independent information to $K$ single-antenna users. The IRS with $N$ reflecting elements is deployed to assist the downlink communications. Let $s_k$ with $\mathbb{E}\{|s_k|^2\} = 1$ and $v_k \in \mathbb{C}^M$ be the data symbol and the associated transmit beamformer for the $k$th user, resp. The transmit signal $x \in \mathbb{C}^M$ at the BS and the reflected signal $z_r \in \mathbb{C}^N$ at the IRS are resp. given by

$$x = \sum_{k \in \mathcal{K}} v_k s_k \quad \text{and} \quad z_r = \Theta^H G x,$$

where $G \in \mathbb{C}^{N \times M}$ is the channel from the BS to the IRS, and $\Theta = \sqrt{\eta} \text{Diag}(\theta)$ denoting the reflecting matrix with $\text{Diag}(\theta)$ being a diagonal matrix with diagonal entries $\theta$, $\eta \in [0, 1]$ and $\theta = [\theta_1, \ldots, \theta_N]^T$ being the reflection coefficient and the phase shift at each element of the IRS, resp. In this work, we will consider both continuous-phase shift and discrete-phase shift; that is,

$$\theta_n \in F_{\text{CP}} \triangleq \{ \theta \in \mathbb{C} \mid |\theta| = 1 \}, \forall n = 1, \ldots, N,$$

for the continuous-phase shift, and

$$\theta_n \in F_{\text{DP}} \triangleq \{1, \omega, \ldots, \omega^{L-1}\}, \forall n = 1, \ldots, N,$$

for the discrete-phase shift, where $\omega = e^{2\pi i/L}$ and $L$ is the number of realizable phase angles.

Let $h_{d,k} \in \mathbb{C}^M / h_{r,k} \in \mathbb{C}^N$ be the channel from the BS/IRS to the $k$th user. From (1), the received signal at the $k$th user may be expressed as

$$y_k = h_{d,k}^H x + h_{r,k}^H \Theta^H G x + n_k,$$

$$= (h_{d,k}^H + h_{r,k}^H \Theta^H G) \sum_{\ell \in \mathcal{K}} v_{\ell} s_{\ell} + n_k$$

$$= (h_{d,k} + H_{r,k} \theta)^H \sum_{\ell \in \mathcal{K}} v_{\ell} s_{\ell} + n_k,$$

where $H_{r,k} \triangleq \sqrt{\eta} G^H \text{Diag}(h_{r,k})$ and $n_k \sim \mathcal{C}\mathcal{N}(0, \sigma_k^2)$ is additive white Gaussian noise with mean zero and variance $\sigma_k^2$. Accordingly, the received SINR at the user $k$ is

$$\text{SINR}_k = \frac{|| (h_{d,k} + H_{r,k} \theta)^H v_k ||^2}{\sigma_k^2 + \sum_{j \neq k} || (h_{d,j} + H_{r,j} \theta)^H v_j ||^2}, \forall k \in \mathcal{K}. \quad (2)$$

Upon the above model, our problem of interest is to jointly design the transmit beamformers $\{v_k\}_{k \in \mathcal{K}}$ at the BS and the phase shift $\theta$ at the IRS, so that the sum rate of all the users is maximized, viz.,

$$\max_{\theta, \{v_k\},} \sum_{k \in \mathcal{K}} \log(1 + \text{SINR}_k), \quad (3a)$$

subject to

$$\sum_{k \in \mathcal{K}} \| v_k \|_2^2 \leq P_{\text{max}}, \quad (3b)$$

$$\theta_n \in F, \forall n = 1, \ldots, N, \quad (3c)$$

where $P_{\text{max}}$ is the total transmit power budget at the BS, and $F$ represents either $F_{\text{CP}}$ or $F_{\text{DP}}$.

The difficulty of solving problem (3) lies in the tightly coupled variables in the sum rate expression and the nonconvex phase-shift constraints (3c). Furthermore, when the discrete-phase shift is concerned, problem (3) is essentially a mixed integer nonlinear program, which is generally NP-hard. As such, in the ensuing section, we will focus on developing an efficient approximate approach to problem (3). The crux of our approach is a more tractable reformulation of problem (3) and the block-coordinated APG method.

### III. A TRACTABLE APPROACH TO PROBLEM (3)

Let us first derive an equivalent formulation of problem (3), which facilitates the subsequent algorithm design.

**Theorem 1** There exists some $\bar{\lambda} > 0$ such that for any $\lambda > \bar{\lambda}$, problem (4) is equivalent to problem (3):

$$\min_{\theta, \{u_k, v_k, w_k\}_{k \in \mathcal{K}}} \phi_\lambda(V, u, w, \theta) \quad (4a)$$

subject to

$$\sum_{k \in \mathcal{K}} \| v_k \|_2^2 \leq P_{\text{max}}, \quad (4b)$$

$$w_k \geq 0, \forall k \in \mathcal{K}, \quad (4c)$$

$$\theta_n \in F, \forall n = 1, \ldots, N, \quad (4d)$$

where $F$ denotes the convex hull of $\mathcal{F}$ and

$$\phi_\lambda(V, \theta, u, w) = \sum_{k \in \mathcal{K}} (w_k e_k(u_k, V) - \log(w_k)) - \lambda \|\theta\|_2^2 \quad (5)$$

$$e_k(u_k, V) = \sigma_k^2 |u_k|^2 + |1 - u_k (h_{d,k} + H_{r,k} \theta)^H v_k|^2$$

$$+ \sum_{j \neq k} |u_k (h_{d,j} + H_{r,j} \theta)^H v_j|^2 \quad (6)$$

**Proof.** See the Appendix.

The main advantage of the reformulation (4) is that all the constraints are convex with a nice geometry structure, particularly for the set $F$. However, the tightly coupled nonconvex objective still poses a challenge for problem (4). In the following, we will take a divide-and-conquer approach to alternately optimize problem (4) with respect to $\{u_k, v_k, w_k\}_{k \in \mathcal{K}}$ and $\theta$. Algorithm 1 summarizes the main procedure of our approach.

**Algorithm 1.** A divide-and-conquer approach to problem (4)

1. Initialize $(V^0, u^0, w^0, \theta^0)$ and set $t = 0$.
2. repeat
3. $$(V^{t+1}, u^{t+1}, w^{t+1}) = \arg \min_{V, u, w} \phi_\lambda(V, u, w, \theta^t)$$

$$\text{s.t.} \quad (4b) - (4c) \quad (7)$$

4. $\theta^{t+1} = \arg \min_{\theta} \phi_\lambda(V^{t+1}, u^{t+1}, w^{t+1}, \theta)$$

$$\text{s.t.} \quad (4d) \quad (8)$$

5. $t = t + 1$

6. until some stopping criterion is satisfied.

### A. Solution to problem (7)

With fixed $\theta$, it is easy to see that problem (7) is exactly the same as the weighted MMSE (WMMSE) problem in (10). Therefore, a block-coordinate descent (BCD) approach can be employed to handle problem (7). Readers are referred to (10) for details; herein we only highlight the main steps. Specifically, the WMMSE algorithm cyclically optimizes one of the block variables in $(u, w, V)$ with the other two fixed.1

1For $F_{\text{CP}}$, its convex hull is $F_{\text{CP}} = \{ \theta \in \mathbb{C} \mid |\theta| \leq 1 \}$, and for $F_{\text{DP}}$, its convex hull is a regular polygon with vertices $\{1, \omega, \ldots, \omega^{L-1}\}$.
In this way, it is shown in [10] that each block variable can be updated in a closed-form manner, which is given by

\[ u_k = \frac{(h_{d,k} + H_{r,k}^H v_k)}{\sigma_k^2 + \sum_{j \in K} |(h_{d,k} + H_{r,k}^H v_j)|^2}, \quad \forall k \in K \]  
\[ (9a) \]

\[ w_k = c_k^{-1}(u_k, V), \quad \forall k \in K \]  
\[ (9b) \]

\[ v_k = u_k w_k F(u, w) h_k, \quad \forall k \in K, \]  
\[ (9c) \]

where

\[ F(u, w) = \left( \sum_{j \in K} w_j |u_j|^2 (h_{d,j} + H_{r,j}^H (h_{d,j} + H_{r,j}^H) + \mu I) \right)^{-1} \]

and \( \mu \in \mathbb{R}_+ \) is Lagrangian multiplier associated with the transmit power constraint (4b).

B. Solution to problem \((5)\)

Problem \((5)\) can be simplified as the following problem after dropping the terms irrelevant to \( \theta \)

\[
\min_{\theta} f_\lambda(\theta) \triangleq \theta^H A \theta + 2\Re(b^H \theta) - \lambda \| \theta \|^2 \\
\text{s.t. } \theta_n \in \bar{F}, \quad \forall n = 1, \ldots, N, 
\]  
\[
(10) \]

where

\[
A = \sum_{k \in K} w_k |u_k|^2 H_{r,k}^H (\sum_{j \in K} v_j v_j^H) H_{r,k} \\
B = \sum_{k \in K} b_{k,j} \\
b_{k,j} = \begin{cases} 
(\sum_{k \in K} |v_j|^2 h_{d,k}^H - w_k u_k^H h_{r,k}^H v_k & j = k \\
(\sum_{k \in K} |v_j|^2 h_{r,k}^H v_j v_j^H h_{d,k} & j \neq k
\end{cases}
\]

Problem \((10)\) is generally a nonconvex problem with the objective \(f_\lambda(\theta)\) in the form of difference-of-convex (DC) functions. We tackle it by gradient extrapolated majorization minimization (GEMM) [3]. Consider a convex surrogate function \(G_\lambda(\theta|\theta)\) of \(f_\lambda(\theta)\), which is given by

\[
G_\lambda(\theta|\theta) = \theta^H A \theta + 2\Re(b^H \theta) - \lambda \| \theta \|^2 \\
= 2 \Re(b^H \theta) + 2 \Re(\theta^H(b^H \theta - \bar{\theta}))
\]

for some \(\bar{\theta} \in \bar{F}\); i.e., \(G_\lambda(\theta|\theta)\) approximates \(f_\lambda(\theta)\) by linearizing the concave term \(-\lambda \| \theta \|^2\) at the point \(\theta\). It is easy to check that

\[
f_\lambda(\theta) \leq G_\lambda(\theta|\theta), \quad \forall \theta, \bar{\theta} \in \bar{F} \text{ and } f_\lambda(\theta) = G_\lambda(\bar{\theta}|\theta), \quad \forall \theta, \bar{\theta} \in \bar{F},
\]

i.e., \(G_\lambda(\theta|\theta)\) is a majorant of \(f_\lambda(\theta)\). Given some initial \(\theta^0\), we recursively update \(\theta\) via

\[
\theta^{\ell+1} = \text{arg min}_{\theta \in \bar{F}} G_\lambda(\theta|\theta^\ell), \quad \ell = 0, 1, 2, \ldots
\]  
\[
(11) \]

and outputs the final \(\theta^\ell\) as \(\theta^{\ell+1}\). In \((12)\), \(1/\beta_\ell > 0\) is the step size, which can be determined by backtracking line search; \(z^i\) is an extrapolated point and is given by

\[
z^i = \theta^i + \zeta_i (\theta^i - \theta^{i-1})
\]

with

\[
\zeta_i = \frac{\eta_i - 1}{\eta_i}, \quad \eta_i = 1 + \sqrt{1 + 4\eta_{i-1}^2}, \quad \eta_{i-1} = 0,
\]

and \(\Pi_{\bar{F}}(\theta)\) denotes the projection of \(\tilde{\theta}\) onto the set \(\bar{F}\) in an element-wise manner. In particular, for \(\bar{F}_{CP}\), \(\Pi_{\bar{F}_{CP}}(\theta)\) is given by

\[
\Pi_{\bar{F}_{CP}}(\theta) = \begin{cases} 
\theta & \text{if } |\theta| \leq 1, \\
\theta/|\theta| & \text{otherwise}.
\end{cases}
\]

For \(\bar{F}_{DP}\), \(\Pi_{\bar{F}_{DP}}(\theta)\) is given by [9]

\[
\Pi_{\bar{F}_{DP}}(\theta) = e^{i \frac{2 \pi m}{\pi/L} \left( [\theta(t_0) \cos(\pi/L)] + j [\theta(t_0) \sin(\pi/L)] \right)}
\]

where \(m = \lfloor \frac{\theta^c + \pi/L}{2 \pi/L} \rfloor\), \(\hat{\theta} = \theta e^{-i \frac{2 \pi m}{\pi/L}}\) and \([x]_a^b\) defines the thresholding operator, i.e., \([x]_a^b = \min\{b, \max\{x, a\}\}\).

One may notice that for each MM update in \((11)\), we need to run multiple APG iterations, which could incur high computation burden. A more straightforward and efficient strategy is to perform the MM update inexact by running the APG with only one iteration. By doing so, one can update \(\theta^\ell\) more frequently and speed up the convergence. In addition, under some technical conditions [9], this inexact MM has been shown to converge to the Karush-Kuhn-Tucker (KKT) solution of problem \((10)\). Algorithm \(2\) summarizes the whole procedure for solving problem \((10)\). We should mention that by our numerical experience, it is preferable to start with a relatively small \(\lambda\) and gradually increase it (cf. line 6) in order to avoid an ill-posed problem.

**Algorithm 2. GEMM Method for Problem (10)**

1: given a feasible point \(\theta^0 \in \bar{F} \times 1\), an initial penalty \(\lambda > 0\) and its upper bound \(\lambda_{\text{upp}} > 0\), threshold \(\epsilon > 0\), integers \(J \geq 1\), \(c > 1\).
2: \(\theta^{-1} = \theta^0\)
3: \(\ell = 0\)
4: repeat
5: \(z^\ell = \theta^\ell + \zeta_\ell (\theta^\ell - \theta^{\ell-1})\)
6: \(\theta^{\ell+1} = \Pi_{\bar{F}} \left( z^\ell - \frac{1}{\beta_\ell} \nabla \theta G_{\lambda}(z^\ell|\theta^\ell) \right)\)
7: \(\ell = \ell + 1\)
8: until \(\lambda > \lambda_{\text{upp}}\)

**IV. Simulation Results**

The simulation scenario is as follows: The BS and the IRS are located at (0, 0) and (100, 50), resp., and there are four users (\(K = 4\)) which are randomly and uniformly distributed within a circle centered at (200, 0) with radius 10 m. We
set $M = 4$, $\eta = 0.8$, $\sigma^2_k = -117$dBm, $\forall k$. The large-scale path loss is $PL = \sqrt{C_0 \zeta d^{-\alpha}}$, where $C_0$ is the path loss at the reference distance 1 meter, $\zeta$ denotes the product of the source and the terminal gain, $d$ and $\alpha$ denote the link distance and the path loss exponent, resp. Specifically, for the direct channel between the BS and the user $k$, we set $C_0 \zeta_B,k = -30$dB, $\alpha_B,k = 3.2$, $\forall k \in K$. Regarding the IRS link, the total path loss is $C^2_0 \zeta_B,k \zeta_I,k = -122$dB and $\alpha_B,I = \alpha_I,k = 2$. The small-scale fading follows Rayleigh distribution. All the results were averaged over 1,000 trials.

Based on the above settings, we compare our algorithm with the following schemes: 1) The WMMSE scheme without using the IRS [10]; 2) the algorithm in [8], which is state-of-the-art algorithm for the IRS-aided SRM problem. In Fig. 2 we compare the average sum rate and simulation time between our design and the method in [8] with $P_{\text{max}} = 2$dBm, when the number of IRS elements is increased from 10 to 70. From Fig. 2(a) we see that the sum rate achieved by the proposed design outperforms that in [8], and from Fig. 2(b) we also see that the complexity of our design is not very sensitive to the number of IRS elements. Particularly, for $N \leq 40$ our design exhibits a comparable complexity with that in [8], while for $N > 40$ our design has much lower complexity than [8]. Therefore, the proposed approach could be more preferable for the large-scale IRS case. Moreover, to illustrate the performance of the design under different transmit power, we fix the number of IRS elements $N = 30$, and increase $P_{\text{max}}$ from $-4$dBm to $14$dBm. The result is shown in Fig. 3. From the figure, we can see that the proposed design attains the best performance among the compared schemes under different resolutions of the phase shifters.

V. Conclusion

In this paper, we have investigated the joint transmit beamforming and reflecting phase shift design for the IRS-aided MU-MISO downlink transmission. The sum rate maximization problem under both the continuous-phase and the discrete-phase constraints on the IRS is studied. By leveraging on the recent advances in discrete-phase optimization, a block-coordinated APG algorithm is custom-derived. Numerical results have demonstrated that the inclusion of IRS can effectively improve the spectral efficiency for the conventional wireless systems. Moreover, the proposed design can attain much higher sum rate than state-of-the-art design.

Appendix

Firstly, by employing the well-known rate-MMSE relationship [10], we have

$$
\log(1 + \text{SINR}_k) = \log([e_k^{\text{MMSE}}]^{-1}),
$$

where $e_k^{\text{MMSE}}$ represents the MMSE of user $k$’s received signal $s_k$, and it is given by

$$
e_k^{\text{MMSE}} = \min_{u_k \in \mathbb{C}} e_k(u_k, V),
$$

where $e_k(u_k, V)$ is defined in [6]. It follows from Eqns. 13 and 14 that

$$
\log(1 + \text{SINR}_k) = \log([\min_{u_k \in \mathbb{C}} e_k(u_k, V)]^{-1}) = \max_{u_k \in \mathbb{C}} \log([e_k(u_k, V)]^{-1}).
$$

Notice the following identity

$$
\log(x^{-1}) = \max_{w \geq 0} -wx + \log(w) + 1
$$

with the optimal $w^* = x^{-1}$. The last line of 15 can be further expressed as

$$
\log(1 + \text{SINR}_k) = \max_{u_k \in \mathbb{C}, w_k \geq 0} -w_k e_k(u_k, V) + \log(w_k) + 1.
$$

Therefore, problem 3 is equivalent to

$$
\min_{\mathbf{\theta}, w_k \in \mathbb{C}, w_k \geq 0} \sum_{k \in K} w_k e_k(u_k, V) - \log(w_k) \quad (17a)
$$

s.t. $\sum_{k \in K} ||\mathbf{\eta}_k||^2 \leq P_{\text{max}},$ \hspace{1cm} (17b)

$$\mathbf{\theta}_n \in \mathcal{F}, \forall n = 1, \ldots, N.$$

(17c)
Next, we leverage the following lemma to tackle the discrete-phase constraints.

**Lemma 1 ([9])** Consider the following two problems

\[
\min_{x \in \mathbb{C}^N} f(x) \quad \text{s.t.} \quad x_n \in \mathcal{F}, \quad \forall \ n = 1, \ldots, N, \quad (18)
\]

and

\[
\min_{x \in \mathbb{C}^N} f(x) - \lambda \|x\|^2 \quad \text{s.t.} \quad x_n \in \bar{\mathcal{F}}, \quad \forall \ n = 1, \ldots, N, \quad (19)
\]

where \( \lambda > 0 \), \( f : \mathbb{C}^N \to \mathbb{R} \) is a \( \mu \)-Lipschitz continuous function, and \( \bar{\mathcal{F}} \) denotes the convex hull of \( \mathcal{F} \). Then, there exists a constant \( \bar{\lambda} > 0 \) such that for any \( \lambda > \bar{\lambda} \), any (globally) optimal solution to problem (18) is also a (globally) optimal solution to problem (19): the converse is also true. Specifically, we have \( \lambda = \mu \) for the continuous-phase \( \mathcal{F}_{CP} \) case, and \( \bar{\lambda} = \mu / \sin(\pi/L) \) for the discrete-phase \( \mathcal{F}_{DP} \) case.

By applying Lemma [1] for problem (17), we arrive at the result in Theorem [1]. This completes the proof.

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