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POSTER: A low-complexity model for IRS-aided beyond 5G wireless networks

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***Abstract*—The intelligent reflecting surface is a key reflecting mirror in fifth-generation (5G) and beyond 5G communication which ensures enhanced coverage by generating phase shift at the IRS. Solving IRS's phase shift optimization problem is challenging and non-convex. In this regard, a low complexity model (LCM) is proposed for a multiple-input single-output (MISO) system to optimize passive and active beamforming. Based on initial results, the proposed method estimates IRS phase shifts accurately while being computationally less complex, which will allow it to be studied for multiple users and multiple-input multiple-output (MIMO) systems in the future.**

***Index Terms*—Intelligent reflecting surface, received SNR, beamforming**

I. Introduction

The fifth-generation (5G) and beyond 5G standards ensure high bandwidth, continuous connectivity, and low latency. Millimeter waves (mmWave) in 5G are susceptible to attenuation and obstruction, leading to shrinking coverage and intermittent connection. To overcome the obstacle and coverage issue of mmWave communication systems, intelligent reflecting surfaces (IRS) have emerged as a promising solution [1], [2]. Low-cost, low-power IRS can electronically reconfigure the propagation environment to achieve a better rate, spectral efficiency, and energy efficiency [2]–[5].

IRS phase optimization is a quadratically constrained quadratic problem (QCQP). In [7], [8], [12], the problem is formulated as a standard semidefinite problem (SDP) and solved through the CVX tool. In [9], the non-convex problem is solved using Dinkelbach's procedure. Block Coordinate Descent (BCD) is used in [6] to decompose the optimization problem into multiple suboptimal problems. Iterative alternating optimization (AO) is proposed in [10] to improve sum-rate maximization.

For these conventional methods, channel state information (CSI) is required. The computational complexity of these optimization methods is proportional to $\mathcal{O}\left(Q^{3.5}\right)$ [10], [11]. Thus, with an increasing number of reflecting elements (Q), phase optimization becomes increasingly difficult.

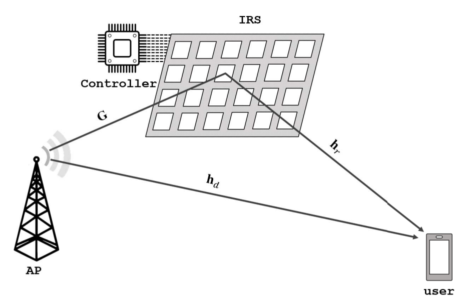

Fig. 1. Illustration of an IRS-aided MISO system.

II. System Model

Consider IRS-aided wireless communication for MISO systems, as illustrated in Fig. 1. There are uniform planar arrays (UPA) of N antennas on the AP, which transmit data. The IRS is equipped with Q uniformly spaced planar programmable reflective elements. AP-IRS, IRS-user, and AP-user links are specified with $\mathbf{G} \in \mathbb{C}^{Q\times N}$, $\mathbf{h}_r^H \in \mathbb{C}^{1\times Q}$, and $\mathbf{h}_d^H \in \mathbb{C}^{1\times N}$ channel denotation and are expressed as

$$\mathbf{h}_d^H = \sqrt{\alpha_d}\ h_d\ \mathbf{D}_{AP}^H, \tag{1a}$$

$$\mathbf{h}_r^H = \sqrt{\alpha_r}\ h_r\ \mathbf{D}_{IRS}^H, \tag{1b}$$

$$\mathbf{G} = \sqrt{\alpha_g}\ \mathbf{A}_{IRS}\ g\ \mathbf{D}_{AP}^H, \tag{1c}$$

where $\mathbf{A}_x$ denotes the angle of arrival at x and $\mathbf{D}_x$ indicates the angle of departure at x. The path losses for the three channels are denoted as α_d, α_r, and α_g.

The precoder $\mathbf{w} \in \mathbb{C}^{N\times 1}$ is the active beamforming vector. Assume that the controller induces IRS's phase shifts as $\Theta = \mathrm{diag}\left(\Theta_d\right) = \mathrm{diag}\left(\mathrm{e}^{j\theta_1},\ \mathrm{e}^{j\theta_2}, \ldots,\ \mathrm{e}^{j\theta_Q}\right)$. The signal received by the user is given by

$$y = (\mathbf{h}_r^H \Theta \mathbf{G} + \mathbf{h}_d^H)\mathbf{w}x + n, \tag{2}$$

where x denotes the independent and identically distributed signal that follows $\mathbf{E}\left[x^2\right] = 1$ and n represents the additive white Gaussian noise (AWGN) of power σ_n^2.

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III. Proposed beamforming model

In the beamforming model, passive and active beamforming are optimized using the proposed low-complexity model (LCM). The beamforming evaluation algorithm assumes complete CSI availability. As a starting point, the passive beamforming matrix is evaluated, followed by an alternate evaluation of the active beamforming matrix.

An optimal value of Θ improves $\gamma = \mathbf{h}_r^H \Theta \mathbf{G} + \mathbf{h}_d^H$ and leads to the maximum value of received SNR, $\Gamma_r = \frac{\|\gamma \mathbf{w}\|^2}{\sigma_n^2}$. In order to obtain optimal phase shift, γ is reformulated as follows.

$$\gamma = \Theta \ \mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G} + \mathbf{h}_d^H, \tag{3a}$$

$$= \begin{bmatrix} 1 & \Theta \end{bmatrix} \begin{bmatrix} \mathbf{h}_d^H \\ \mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G} \end{bmatrix}, \tag{3b}$$

$$= \begin{bmatrix} 1 & \Theta \end{bmatrix} \begin{bmatrix} \mathbf{h}_{d,1}^H & \dots & \mathbf{h}_{d,N}^H \\ \mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G_1} & \dots & \mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G_N} \end{bmatrix} \tag{3c}$$

Let us perform an elementwise division of IRS channel with the direct channel as follows:

$$\gamma_1 = \begin{bmatrix} 1 & \Theta \end{bmatrix} \begin{bmatrix} 1 & \dots & 1 \\ \frac{\mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G_1}}{\mathbf{h}_{d,1}^H} & \dots & \frac{\mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G_N}}{\mathbf{h}_{d,N}^H} \end{bmatrix} \tag{4a}$$

Taking the cumulative channel response for N transmitting antennas and diving by N.

$$\gamma_2 = \frac{1}{N} \begin{bmatrix} 1 & \Theta \end{bmatrix} \begin{bmatrix} N \\ \sum_{i=1}^{N} \frac{\mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G}_i}{\mathbf{h}_{d,i}^H} \end{bmatrix}, \tag{5a}$$

$$= \begin{bmatrix} 1 & \Theta \end{bmatrix} \begin{bmatrix} 1 \\ \frac{1}{N} \sum_{i=1}^{N} \frac{\mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G}_i}{\mathbf{h}_{d,i}^H} \end{bmatrix} \tag{5b}$$

From (5b), γ_2 will be maximum when Θ_d is optimal

$$\Theta_{opt} = \mathrm{e}^{\left(-j\angle\left(\frac{1}{N} \sum_{i=1}^{N} \frac{\mathrm{diag}\left(\mathbf{h}_r^H\right) \mathbf{G}_i}{\mathbf{h}_{d,i}^H}\right)\right)}, \tag{6}$$

where the phase of the complex value is denoted by the notation $\angle(.)$. The optimal precoding vector for transmit power p can be evaluated as maximal ratio transmission (MRT).

$$\mathbf{w_o} = \sqrt{p} \frac{\mathbf{h}_r^H \Theta_{opt} \mathbf{G} + \mathbf{h}_d}{\|\mathbf{h}_r^H \Theta_{opt} \mathbf{G} + \mathbf{h}_d\|} \tag{7}$$

Received signal strength for noise power σ_n^2 at the user can be expressed as follows:

$$\Gamma_r = \frac{\|\left(\mathbf{h}_r^H \Theta_{opt} \mathbf{G} + \mathbf{h}_d\right) \mathbf{w_o}\|^2}{\sigma_n^2}. \tag{8}$$

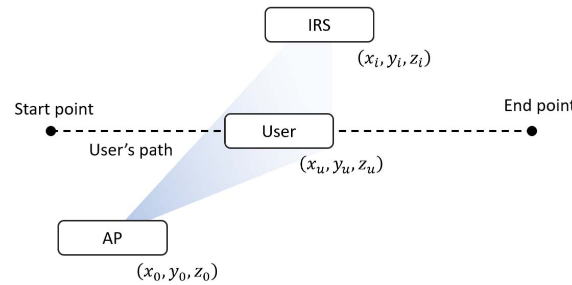

Fig. 2. Simulation setup for a single IRS network

IV. Simulation results and discussion

The user moves from AP to IRS in the simulation, as shown in Fig. 2. An IRS is assumed to be installed in the LOS region of the AP to perform highly correlated passive beamforming. As the IRS is located near the user's path, the link to the user experiences less scattering, so the link is considered LOS. Centrally located APs may be far from users. Thus an NLOS link is assumed between the AP and the user. All LOS links consider a path loss exponent of 2.5 and a reference path loss of -45 dB. Rician fading is used to model LOS links. The NLOS link is modeled as Rayleigh fading with a reference path loss of -42 dB and a path loss exponent of 4.7. The shadow factors of LOS and NLOS links are set as 0.005 and 8.6, respectively [14]. The height of AP, user, and IRS are considered as 9m, 1.3m, and 2m. The absolute distance between AP and the user is denoted as d, where $d^2 = (x_u - x_0)^2 + (y_u - y_0)^2 + (z_u - z_0)^2$.

In Fig. 3, the performance of the proposed LCM model is illustrated. The proposed model and the iterative AO model have the highest accuracy, followed by convex optimization using semidefinite relaxation. The proposed method results in a high received signal strength of approximately 16dB at $d = 85$m compared to a model with IRS's random phase shift and without IRS. In iterative methods, the computational complexity is $\mathcal{O}\left(Q^3 N I_{iter}\right)$, while an SDR method has a computational complexity of $\mathcal{O}\left(\log\left(\frac{1}{\epsilon_s}\right)\left(2\left(Q+1\right)^3 N + 4\left(Q+1\right)^2 N + 8N\right)\right)$ [10], [13] . The notation I_{iter} in iterative AO and ϵ_s in SDR denotes the number of iterations and threshold

TABLE I
Simulation Parameters

Parameters	Value
Number of antennas at the transmitter (N)	64, 256
Number of reflecting element (Q)	64, 256
Power of the transmitted signal (p)	35 dBm
Power of the noise signal (σ_n^2)	-84 dBm
Operating freqency	28 GHz

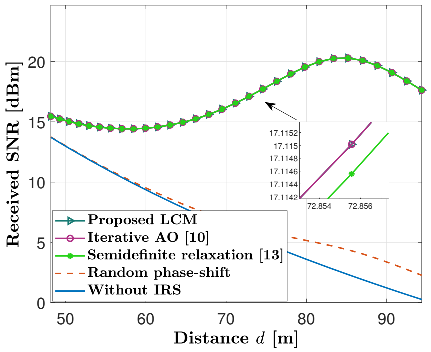

Fig. 3. Received SNR versus AP-user distance

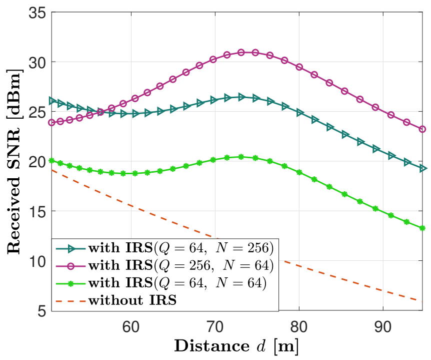

Fig. 4. Received SNR for different combination of N and Q

error, respectively. However, the proposed model exhibits a low computational complexity of $\mathcal{O}\left(QN\right)$.

Fig. 4 illustrates the received SNR for different numbers of transmitting antennas (N) and reflecting elements (Q). Increasing the number of transmitting antennas improves SNR by about 6dB throughout the range. The received SNR increases by 10dB near the IRS as the number of reflecting elements increases. The Monte Carlo method is employed in the simulation to reduce channel randomness. The simulations are carried out using MATLAB 2020b.

V. Conclusion

In this study, LCM performance is evaluated by jointly optimizing active and passive beamforming. According to the results, beamforming with LCM achieves maximum received SNR as the user approaches the IRS. As compared to existing models, the proposed LCM has lower computational complexity. Similarly, performance was assessed for different combinations of transmitting antennas (N) and reflecting elements (Q). The increased value of N produces constant power enhancement in both LOS and NLOS paths, whereas increasing Q shows significant power enhancement in NLOS paths.

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