Performance Analysis of IRS-Aided NOMA Communications in the Presence of Imperfect SIC

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Abstract

The intelligent reflecting surface (IRS) is expected to be a promising technique to achieve a robust spectrum and energy efficiency. This paper investigates the advantages of IRS in enhancing performance of non-orthogonal multiple access (NOMA) communications in the presence of imperfect successive-interference-cancellation (SIC) and phase distortion (PD) caused by a non-ideal IRS. Specifically, average achievable rates (AARs) of the users are the target performance metrics. For performance evaluation, the probabilistic characterizations of signal-to-interference-plus-noise ratios (SINRs) at the users are studied. These results allow for deriving the theoretical formulas for the AAR. Monte Carlo simulations are adopted to verify the accuracy of these theoretical results. The numerical results show the effects of various key system parameters, such as source transmit power, NOMA power allocation (PA) factors, reflecting tile (RTs) allocation, the SIC imperfection factor, and the PD factor, on the AAR that provide useful information for the system's design.

Key Words: Achievable Rate Analysis, Imperfect SIC, NOMA, Non-ideal IRS, 5G and Beyond 5G.

I. INTRODUCTION

The fifth-generation (5G) network has been being rolled out and has exhibited a remarkable performance, such as a higher data rate, larger network capacity, lower latency, and ubiquitous mobile coverage, compared to its predecessors [1]; however, the exponential growth in data usage demand and the number of wireless devices creates new challenges for the 5G network as well as the next-generation networks [2]; especially, in the era of Internet of Things (IoT), this problem becomes more critical. Numerous technical efforts have been done for enlarging the network's capability and enhancing the quality of service toward advanced communication applications. For instance, massive multiple-input-multiple-output (MIMO) technology, which uses a large number of antennas to achieve higher coverage and

capacity, was introduced. A large millimeter-wave (mmWave) spectrum band is exploited to allow high-speed connectives and high network capacity. Moreover, non-orthogonal multiple access (NOMA), which uses a power-domain multiplexing technique to accommodate several users within an orthogonal resource block, showed a significant improvement in spectrum efficiency [3].

Recently, the intelligent reflecting surface (IRS), which is capable of reflecting electromagnetic waves with a controllable phase-shift and direction, has emerged as a promising technique to realize a smart and programmable wireless environment [4]. The key idea of the IRS is based on special electromagnetic properties on repetitive artificial subwavelength structures [5]. For convenience in manufacturing and deployment, the IRS consists of a large number of independent reflecting tiles (RTs)

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that are interconnected and controlled by a programmable controller. To verify the advantages of the IRS, the performance of IRS-aided wireless systems has been studied. The works on IRS's physical channel modeling and channel estimation were done in [6–8], allowing for a performance evaluation of IRS-aided communication systems. In [9–12], the authors proposed potential applications of the IRS in various communication systems, such as wireless powered networks, secure communications, terahertz communications, cognitive radio systems, and MIMO systems, and showed that these systems gain a significant improvement in performance.

In this paper, we study the advantages of the IRS on enhancing performance of a practical indoor NOMA communication with imperfect successive-interference-cancellation (SIC) and in the presence of phase distortion (PD) caused by a non-ideal IRS. Unlike other approaches, where extravagant state variations are required at the IRS, our approach evaluates average performance metrics, i.e., the average achievable rates (AARs) at users, to significantly reduce IRS's state variations, hence achieving robust power efficiency. First, we adopt a novel channel model proposed for indoor IRS-aided wireless communications in [10] to derive the signal-to-interference-plus-noise ratios (SINRs) for each signal at each user. Next, we study the distribution characterization of relevant performance metrics. Finally, the theoretical AAR formulas can be obtained using the above results. Numerical results are presented for analysis verification and performance evaluation. Moreover, these results point out the effects of system parameters, such as source transmit power, NOMA power allocation (PA) factors, RT allocation, the PD factor, and the SIC imperfection factor, on system performance.

The rest of this paper is organized as follows. The system and channel models are described in Section II. The SINR's distribution and theoretical AAR formulas are studied in Section III. Numerical results and discussions are presented in Section IV. Finally, conclusions are presented in Section V.

Notation: Rice(Ω, K) denotes the Rice distribution with scale parameter Ω and shape parameter K ; $CN(0, \delta^2)$ denotes the complex Gaussian distribution with zero mean and variance δ^2 ; $\mathbb{E}\{\cdot\}$ is the expectation of a random variable (RV) X .

II. SYSTEM AND CHANNEL MODELS

We consider an IRS-aided NOMA communication, as illustrated in Fig. 1, in which a source S communicates with two users, a near user U_1 and a far user U_2 , with the help of an IRS attached to a wall. All nodes are single-antenna devices. The NOMA technique is used for data transmission for the users. The IRS includes N RTs, $R_n, n = 1, \dots, N$, and each RT is a reconfigurable metasurface and can independently direct the signal to a user under the control of a programmable controller.

Fig. 1. The system model.

The channel model proposed for indoor IRS-aided wireless communications in [10, 13] is adopted to characterize the wireless channels of our system. For instance, the received signal at $U_m, m \in 1, 2$, comprises of (i) the signals propagating over the line-of-sight (LoS) (or the direct link) and reflecting links, both are the dominant links, and (ii) scattering signals caused by multi-path fading or multiple reflections on the IRS. Let $\mathcal{R}_m$ be the set of RTs allocated to $U_m, m = \{1, 2\}$; hence, the channel between S and U_m is expressed as

$$h_m = h_m^{\text{LoS}} + h_m^{\text{Fad}} + \sum_{R_n \in \mathcal{R}_m} h_{\text{SR}_n U_m}^{\text{Ref}}, \quad (1)$$

where h_m^{LoS} and $h_{\text{SR}_n U_m}^{\text{Ref}}$ are channel coefficients for the LoS link between S and U_m and for the reflecting link between S and U_m via R_n , i.e., the $S - R_n - U_m$ link, respectively, and $h_m^{\text{Fad}} \sim CN(0, \delta_{\text{Fad}}^2)$ is the channel coefficient representing the effect of multi-path fading and/or multi-reflection on the RTs.

Let $G_S(\theta, \phi)$ and $G_{U_m}(\theta, \phi)$, respectively denote directivity patterns of the transmit antenna of S and the receive antenna of U_m where θ and ϕ are respectively the zenith and azimuth angles. For given locations of S and U_m and given antenna's orientations, we can calculate the values for θ and ϕ , hence obtaining the values for transmit/receive directivities for wireless links. Let $G_S^{S \rightarrow X} \triangleq G_S(\theta_X, \phi_X)$ be the transmit directivity for the link from S to a given point X , and let $G_{U_m}^{Y \rightarrow U_m} \triangleq G_{U_m}(\theta_Y, \phi_Y)$ be the receive directivity for the link from a given point Y to U_m . Using these transmit/receive directivities and the free-space path loss (FSPL) model, we can determine h_m^{LoS} and $h_{\text{SR}_n U_m}^{\text{Ref}}$ as

$$h_m^{\text{LoS}} = \sqrt{G_S^{S \rightarrow U_m} G_{U_m}^{S \rightarrow U_m}} \frac{\lambda_0 e^{-j\varphi_{\text{SU}_m}}}{4\pi d_{\text{SU}_m}}, \quad (2)$$

$$h_{\text{SR}_n U_m}^{\text{Ref}} = \sqrt{G_S^{S \rightarrow R_n} G_{U_m}^{R_n \rightarrow U_m}} \frac{\lambda_0 \psi_n e^{-j\varphi_{\text{SR}_n U_m}}}{4\pi d_{\text{SR}_n U_m}}, \quad (3)$$

where $\frac{\lambda_0}{4\pi d_{SU_m}}$ and $\frac{\lambda_0}{4\pi d_{SR_n U_m}}$ are the loss factors for the FSPL model with the wavelength λ_0 ; d_{SU_m} is the distance between S and U_m ; $d_{SR_n U_m} = d_{SR_n} + d_{R_n U_m}$ is the distance of the $S - R_n - U_m$ link; d_{SR_n} and $d_{R_n U_m}$ are the distances between S and R_n and between R_n and U_m , respectively; $\psi_{R_n} = \beta_{R_n} e^{j\varphi_{R_n}}$ is the reflecting coefficient of R_n with $\beta_{R_n} \leq 1$ and $\varphi_{R_n} \in [0, 2\pi)$; and $\varphi_{SU_m} = \frac{2\pi d_{SU_m}}{\lambda_0}$ and $\varphi_{SR_n U_m} = \frac{2\pi d_{SR_n U_m}}{\lambda_0}$ are the phase shifts caused by wave propagation over the distances d_{SU_m} and $d_{SR_n U_m}$.

As shown in (1), h_m is a Rice distributed RV. Let $b \sim \text{Rice}(\Omega, K)$ be a general Rice distributed RV, then the cumulative distribution function (CDF) of $|h|^2$ is expressed as [10]:

$$F_0(x; K, \Omega) = 1 - e^{-\mu x} \sum_{k=0}^{\infty} B_k (\mu x)^k, \quad (4)$$

where $\mu = \sigma_{\text{Fad}}^{-2}$ and $B_k = \left(1 - e^{-K} \sum_{i=0}^{k-1} \frac{K^i}{i!}\right) / k!$.

Next, we study the effect of PD caused by non-ideal RTs on the received signal at the users. Regarding the NOMA principle, S transmits a combined signal $x_s = \sqrt{\alpha_1} x_1 + \sqrt{\alpha_2} x_2$ to the users where $x_1, E\{|x_1|^2\} = P_0$, and $x_2, E\{|x_2|^2\} = P_0$, are respectively the desired signals for U_1 and U_2 ; α_1 and α_2 are the PA factors of NOMA satisfying $\alpha_1 + \alpha_2 = 1$. When x_s is reflected from the RTs, the PD caused by the non-ideal RTs make it distorted. The effect of non-ideal hardware on system performance has been widely studied in [14–16] and is characterized as a complex Gaussian noise $\kappa_{R_n}^{\text{PD}} \sim \text{CN}(0, \tau_{R_n} P_0)$ with the PD factor $\tau_{R_n} \ll 1$. Using (1), the received signal for U_m in the presence of PD is given by

$$y_m = h_m^{\text{LoS}} x_s + h_m^{\text{Fad}} x_s + \left(\sum_{R_n \in \mathcal{R}_m} h_{SR_n U_m}^{\text{Ref}} \hat{x}_s \right) + n_m, \quad (5)$$

where $n_m \sim \text{CN}(0, N_0)$ is the additive white Gaussian noise (AWGN) at the receive antenna, and $\hat{x}_s \triangleq x_s \sqrt{1 - \tau_{R_n}} + \kappa_{R_n}^{\text{PD}}$ is the distorted signal reflected on R_n . The factor $\sqrt{1 - \tau_{R_n}}$ is the power-constraint factor for guaranteeing $E\{|\hat{x}_s|^2\} = E\{|x_s|^2\} = P_0$, or equivalently, the IRS does not add any power to the reflected signal. We can rewrite (5) as

$$y_m = \underbrace{\left(h_m^{\text{LoS}} + h_m^{\text{Fad}} + \sum_{R_n \in \mathcal{R}_m} \sqrt{1 - \tau_{R_n}} h_{SR_n U_m}^{\text{Ref}} \right)}_{\hat{h}_m} x_s + n_{U_m} + \underbrace{\sum_{R_n \in \mathcal{R}_m} h_{SR_n U_m}^{\text{Ref}} \kappa_{R_n}^{\text{PD}}}_{\hat{n}_m}, \quad (6)$$

It is seen from (6) that $\hat{h}_m \sim \text{Rice}(\hat{\Omega}_m, \hat{K}_m)$ is still a Rice distributed RV, and $\hat{n}_m \sim \text{CN}(0, N_0 + \sigma_{\text{PD}}^2)$ is the total noise

including the AWGN and the noise caused by the PD. The values for $\hat{\Omega}_m, \hat{K}_m$, and σ_{PD}^2 are calculated as the follows.

The phase shift of each RT, i.e., φ_{R_n} , plays an important role in system performance. It defines whether the received signals at U_m via the dominant links are constructive or destructive. Let us consider the distance-propagation phase shift of the LoS link, i.e., φ_{SU_m} , as the standard reference in the phase shift at U_m , the signal-combining phase shift denoted by $\varphi_{R_n}^{\text{sc}}$, which measures the difference between the distance-propagation phase shift on the $S-R_n-U_m$ link and φ_{SU_m} . $\varphi_{R_n}^{\text{sc}}$ is calculated as follows [17].

$$\varphi_{R_n}^{\text{sc}} = \text{mod}(\varphi_{R_n} + \varphi_{SR_n U_m} - \varphi_{SU_m}, 2\pi). \quad (7)$$

Using $\varphi_{R_n}^{\text{sc}}, \psi_{R_n}, h_m^{\text{Fad}}$, and the formulas for channel coefficients h_m^{LoS} and $h_{SR_n U_m}^{\text{Ref}}$ (given in (2) and (3)), we can calculate $\hat{\Omega}_m$ and $\hat{K}_m$ using a similar approach in [10]:

$$\hat{\Omega} = \sigma_{\text{Fad}}^2 + \left(\frac{\lambda_0}{4\pi}\right)^2 \left(\left(\frac{\sqrt{G_S^{S \rightarrow U_m} G_{U_m}^{S \rightarrow U_m}}}{d_{SU_m}} + \sum_{R_n \in \mathcal{R}_m} \frac{\sqrt{(1 - \tau_{R_n}) G_S^{S \rightarrow R_n} G_{U_m}^{R_n \rightarrow U_m} \beta_{R_n} \cos(\varphi_{R_n}^{\text{sc}})}}{d_{SR_n U_m}} \right)^2 + \left(\sum_{R_n \in \mathcal{R}_m} \frac{\sqrt{(1 - \tau_{R_n}) G_S^{S \rightarrow R_n} G_{U_m}^{R_n \rightarrow U_m} \beta_{R_n} \sin(\varphi_{R_n}^{\text{sc}})}}{d_{SR_n U_m}} \right)^2 \right), \quad (8)$$

$$\hat{K}_m = \hat{\Omega}_m / \sigma_{\text{Fad}}^2. \quad (9)$$

Then, using the fact that the sum of independent complex Gaussian RVs is a complex Gaussian RV with its variance being the sum of all variances, we can calculate σ_{PD}^2 as

$$\sigma_{\text{PD}, m}^2 = \left(\frac{\lambda_0}{4\pi}\right)^2 \sum_{R_n \in \mathcal{R}_m} \frac{\tau_{R_n} P_0 G_S^{S \rightarrow R_n} G_{U_m}^{R_n \rightarrow U_m} \beta_{R_n}^2}{(d_{SR_n U_m})^2}. \quad (10)$$

III. SINR DISTRIBUTION CHARACTERIZATION AND AVERAGE ACHIEVABLE RATE ANALYSIS

For convenience in analyzing the SINRs, we rewrite (5) as

$$y_m = \hat{h}_m (\alpha_1 x_1 + \alpha_2 x_2) + \hat{n}_m. \quad (11)$$

According to the SIC principle of NOMA, the SINRs for the signal x_2 at both users, i.e., $U_m, m = \{1, 2\}$, are given by

$$Y_m = \frac{\gamma_m |\hat{h}_m|^2 \alpha_2}{1 + \gamma_m |\hat{h}_m|^2 \alpha_1}, \quad (12)$$

where $\gamma_m = P_0 / (N_0 + \sigma_{\text{PD},m}^2)$, and the SINR for the signal x_1 at U_1 is given by

$$X = \frac{\gamma_1 |\hat{h}_1|^2 \alpha_1}{1 + \gamma_1 |g|^2 \alpha_2}. \quad (13)$$

Using Shannon's law, we can calculate the achievable rates (also known as the ergodic capacities) in bits/s/Hz for the signal x_2 at both users as

$$R_2 = \log_2(1 + Y), \quad (14)$$

and for the signal x_1 at U_1 as

$$R_1 = R_{x_1|U_1} = \log_2(1 + X), \quad (15)$$

where $Y = \min(Y_1, Y_2)$ is the condition for guaranteeing successfully decoding x_2 at both users (note that U_1 must decode x_2 before decoding its information x_1), and $g \sim \text{CN}(0, \lambda_{\text{imSIC}})$ in (13) represents the residual signal of x_2 caused by the imperfect SIC receiver at U_1 with variance $\lambda_{\text{SICIm}} = \zeta \mathbb{E}\{|h_{\text{SU}_1}|^2\}$ and the SIC imperfection factor $\zeta \leq 1$.

Proposition 1. The CDF of X denoted by $F_X(t) = \Pr(X < t)$ is approximated as

$$F_X(t) = 1 - \rho_1 e^{\rho_1 \eta} \sum_{k=0}^{\infty} B_k \times \frac{t^k}{(\eta + t)^{k+1}} \Gamma(k+1, \rho_1(\eta + t)), \quad (16)$$

where $\rho_m = \mu / (\alpha_1 \gamma_m)$, and $\eta = \lambda_{\text{imSIC}} \alpha_1 / (\mu \alpha_2)$.

Proof. The approximate CDF of X is calculated as $F_X(t) \approx \Pr(|\hat{h}_1|^2 < (1 + \gamma_1 |g|^2 \alpha_2) t / (\gamma_1 \alpha_1))$. Using the probability density function (PDF) of the exponential distribution given by $f_{|g|^2}(t) = \lambda_{\text{imSIC}} \exp(-\lambda_{\text{imSIC}} t)$ and (4), we obtain $F_X(t)$ via solving the following equation.

$$F_X(t) = \int_0^{+\infty} \lambda_{\text{imSIC}} e^{-\lambda_{\text{imSIC}} x} F_0\left(\frac{1}{\gamma_1 \alpha_1} (1 + \gamma_1 \alpha_2 x) t; \hat{\Omega}_1, \hat{K}_1\right) dx = 1 - \eta \sum_{k=0}^{\infty} B_k (\rho_1)^{k+1} e^{\rho_1 \eta} t^k \int_1^{+\infty} e^{-\rho_1(\eta+t)y} y^k dy, \quad (17)$$

Using [18, Eq. (3.381.3)] and some manipulations, we obtain (16). The proof for Proposition 1 is completed. $\square$

Proposition 2. The CDF of Y denoted by $F_Y(t) = \Pr(Y < t)$ is equal to one if $t > \alpha_2 / \alpha_1$, and otherwise, it is calculated as

$$F_Y(t) = 1 - \left(1 - F_0\left(\frac{t}{\gamma_1(\alpha_2 - t\alpha_1)}; \hat{\Omega}_1, \hat{K}_1\right)\right) \times \left(1 - F_0\left(\frac{t}{\gamma_2(\alpha_2 - t\alpha_1)}; \hat{\Omega}_2, \hat{K}_2\right)\right) \quad (18)$$

Proof. Since $Y = \min(Y_1, Y_2)$, we can rewrite $F_Y(t)$ as $F_Y(t) = 1 - \Pr(Y_1 > t) \Pr(Y_2 > t)$. Then, substituting (4) into it yields (18). The proof for Proposition 2 is completed. $\square$

Using Propositions 1 and 2, we can obtain the theoretical results for AARs as follows.

Proposition 3. The theoretical AAR expressions for the signals x_1 and x_2 are respectively given in the equations (19) and (20) shown in the top of the next page, where $C_k = \sum_{q=0}^k B_{k,q} B_{2,k-q} (\gamma_1)^{-q} (\gamma_2)^{-k+q}$.

$$\bar{C}_1 = \frac{1}{\ln(2)} \eta e^{\eta \rho_1} \sum_{k=0}^{\infty} B_k \int_0^{\infty} \frac{t^k}{(1+t)(\eta+t)^{k+1}} \Gamma(k+1, \rho_1(\eta+t)) dt \quad (19)$$

$$\bar{C}_2 = \frac{1}{\ln(2)} \sum_{k=0}^{\infty} C_k \mu^k \int_0^{\alpha_2/\alpha_1} \frac{t^k}{(1+t)(\alpha_2 - t\alpha_1)^k} e^{-\left(\frac{\mu}{\gamma_1} + \frac{\mu}{\gamma_2}\right) \frac{t}{(\alpha_2 - t\alpha_1)}} dt \quad (20)$$

Proof. According to [19], the AAR for the signal x_m , $m = \{1, 2\}$, is calculated as

$$\bar{C}_m = \mathbb{E}\{R_m\} = \frac{1}{\ln(2)} \int_0^{+\infty} \frac{1 - F_m(t)}{1+t} dt. \quad (21)$$

Substituting (16) and (17) into (21) and using some manipulations, we can obtain (19) and (20). The proof for Proposition 3 is completed. $\square$

IV. NUMERICAL RESULTS AND DISCUSSION

In this section, we present the numerical results for the AAR and discuss the effect of different key parameters on it. The theoretical results are derived for a general system, such as any type of transmit/receive directivity of the device antennas, frequency, wireless indoor environment, and device locations. For the simulation, we chose a specific setup for our system. Without loss of generality, we set the phase shifts of the IRS as $\varphi_{R_n} = 2\pi \times \text{mod}(d_{\text{SR}_n} + d_{\text{R}_n \text{U}_m} - d_{\text{SU}_m}, \lambda_0) / \lambda_0$ to gain the strongest received signal strengths (RSSs) at the users, $\beta_{R_n} = 0.9$ and $\tau_{R_n} = 0.1$ where $n = 1, \dots, 2N_{\text{col}}$. The coordinate setups are shown in Fig. 2, and the setup of other parameters is listed in Table 1 [20]. The transmit antenna is a half-wavelength dipole with its orientation illustrated in Fig. 2, and its directivity pattern is given in Table 1. The receive antennas have the same directivity for all directions. The numerical results include the theoretical AARs obtained using (19) and (20), and the simulation results are obtained by evaluating a simulation system using MATLAB. The good match between the theoretical and the simulation results confirms the accuracy of our study.

In Fig. 3, we show the advantage of using the IRS to assist NOMA communication via examining a simple IRS, including 2 RTs. We consider four possible cases denoted by

Fig. 2. The coordinate setups for the simulation.

Table 1. Parameter setups for the simulation

Parameter	Value
Frequency (GHz)	$f_0 = 6$
Size	
IRS (row $\times$ column)	$2 \times N_{\text{col}}$ RT
RT (m ²)	0.5×0.5
PD factor	$\tau_{R_n} = \tau_{PD} = 0.1$
Transmit power (dBm)	$P_S = 0$
Fading & AWGN (dBm)	$\{\sigma_{Fad}^2, N_0\} = \{-68.7, -70\}$
SIC receiver	$\zeta = 0.04$
Tx-antenna setup	Orientation: $\{\theta_s, \phi_s\} = \{20^\circ, 0^\circ\}$.
$G_s(\hat{\theta}, \hat{\phi}) = 1.67 \sin(\hat{\theta})^3$ [20, Eq. (2-18)].	

Note that $\hat{\theta}$ is the angle between the antenna-orientation vector and the vector from S to the considered point. $\hat{\phi}$ is not important, so we do not consider its value.

 Fig. 3. The results of the system's capacities for the ideal-RT scenario: (a) AAR for signal x_1 , (b) AAR for signal x_2 , and (c) total AAR. For the non-ideal-RT scenario: (d) AAR for signal x_1 , (e) AAR for signal x_2 , and (f) total AAR.

$\text{IRS}[m_1, m_2], m_1, m_2 \in \{1, 2\}$, which indicates that the first and the second RTs are allocated to U_{m_1} and U_{m_2} , respectively. The AARs for x_1 , x_2 and total AAR, $\bar{C}_\Sigma = \bar{C}_1 + \bar{C}_2$, for the

ideal-IRS scenario (without the presence of the PD) are shown in Fig. 3(a)–3(c), respectively, and those of the non-ideal-IRS scenario (with the presence of the PD) are shown in Fig. 3(d)–

3(f), respectively. The AARs are displayed as increasing functions of P_S . Without the IRS, the received signal at each user propagates on the direct link and multi-path fading channel. Therefore, the AARs are very low, as shown as the lowest curves in Fig. 3. When the IRS is utilized, the AARs are significantly enhanced even with the presence of the PD. Although different levels of AAR enhancement are observed, these results still indicate the important role of the IRS in improving the RSSs at the users. As shown in Fig. 3(c) and 3(f), the total AARs for the cases of IRS[1,2] and IRS[2,1] (in these cases, the IRS assists the communication for both users) achieve better values than other cases.

In Fig. 4, we adopt the best IRS setup for the case of $P_S = 0$ dBm with $\tau_{PD} = 0.1$ (found in Fig. 3, i.e., IRS[2,1]) and then investigate the effect of the PA factors on the total AAR. The advantages of the NOMA scheme are confirmed by investigating a reference orthogonal multiple access (OMA) scheme. Due to the advantageous location of the near user, the system achieves a highest or lowest total-AAR when allocating all resources to the near user or the far user, respectively. When the SIC imperfection factor ζ increases, the total AAR becomes worse. For a performance comparison, we use a conventional OMA scheme, which defines the AAR for U_m , $m = \{1, 2\}$, as

$$R_m^{\text{OMA}} = v_m \log_2(1 + \gamma_m |\hat{h}_m|^2), \quad (22)$$

and the total AAR for OMA is given by

$$\bar{C}_\Sigma^{\text{OMA}} = E\{R_1^{\text{OMA}}\} + E\{R_2^{\text{OMA}}\}. \quad (23)$$

where $v_m \in (0,1)$, $v_1 + v_2 = 1$, is the resource allocation (RA) factors for OMA (e.g., time allocation factor for TDMA or bandwidth allocation factor for FDMA). Compared to OMA,

Fig. 4. The effects of PA factors for NOMA and RA factors for OMA on the system's capacities.

NOMA can achieve a higher total AAR in the presence of PD and imperfect-SIC. When ζ is sufficiently large, NOMA yields a lower performance compared to OMA. In Fig. 5, we show the joint effect of the RT-allocation strategy and PA factors, i.e., α_1 and α_2 , on the AAR. Because the size of each RT is relatively small compared to the wave-propagation distances, the difference in distance between reflecting links is trivial. It is reasonable to assume that the signal attenuations of reflecting links toward a user are similar. For this reason, the RT allocation strategy is as follows. Let N_{U_1} be the number of RTs allocated to U_1 . These N_{U_1} RTs are assigned from left to right and from top to bottom. The other $N_{U_2} = N - N_{U_1}$ RTs are assigned to U_2 . For a given value of N_{U_1} , $\bar{C}_1$ increases, while $\bar{C}_2$ decreases when we increase α_1 and vice versa. This is because the power resource allocated for the users is defined by α_1 . For given PA factors, $\bar{C}_1$ is shown as an increasing function of N_{U_1} , whereas $\bar{C}_2$ improves with the first increase in N_{U_1} and then reaches the optimal value. Finally, it degrades with further increases in N_{U_1} . The trend of $\bar{C}_2$ is explained as follows. Since both users

(a)

(b)

Fig. 5. The joint effect of PA factors and RT allocations on (a) the AARs for users and (b) the total AAR.

need to decode x_2 successfully, the IRS needs assist both users to gain the highest value of $\bar{C}_2$. Therefore, using very high or very low values of N_{U_1} does not yield good values for $\bar{C}_2$. Although the highest total AAR is observed at $(N_{U_1} = 10, \alpha_1 = 1)$, the NOMA system does not use this configuration. The reason is that at this point, all system resources, including P_S and RTs, are allocated to U_1 ; hence, U_2 is disconnected from the network. For our proposed NOMA system, we focus on the points lying on the intersection curve of the two user-AAR surfaces denoted by $\ell(N_{U_1}, \alpha_1^*)$, on which two users can communicate with the source at the same AAR. Let the highest value observed on $\ell(N_{U_1}, \alpha_1^*)$ be the optimal NOMA-AAR $\ell^* = \max \ell(N_{U_1}, \alpha_1^*)$ and its respective coordinate $(N_{U_1}^*, \alpha_1^*)$ be the optimal solution for ℓ^* . The trends of ℓ^* and $(N_{U_1}^*, \alpha_1^*)$ are studied in Fig. 6.

In Fig. 6, we consider an IRS with a size of 2-by- N_{col} RTs and then plot the optimal NOMA-AAR as shown in Fig. 6(a) and its respective optimal solution, i.e., optimal PA factors as shown in Fig. 6(b) and optimal RT-allocation as shown in Fig.

6(c) and 6(d), under different parameters of P_S and ζ . Fig. 6(a) shows that the more RTs are used, the higher value for optimal NOMA-AAR is achieved. When the IRS's size is sufficiently large, the increase in size does not gain much improvement in optimal NOMA-AAR. Therefore, the size of the IRS should be chosen carefully to achieve a reasonable trade-off between performance and cost. For instance, at $\zeta = 0.04$ and $P_S = 0$ dBm, the 2-by-1 and 2-by-4 IRSs can support 1 bit/s/Hz and 2 bit/s/Hz optimal NOMA-AAR, respectively. In Fig. 6(b)–6(d), we show the trends of the obtained optimal solutions. Regarding the NOMA principle, first, both users need to decode the signal allocated with stronger power, which is x_2 for our proposed system, and consider the rest signals as interference, as shown in (12). After successfully decoding x_2 , U_1 eliminates the signal x_2 from its received signal and then decodes its desired signal x_1 , which is allocated with lower power. The allocated power for each signal is defined by the PA factors. Compared to x_2 , the decoding process of x_1 deals with lower interference, which is caused by the imperfect-SIC decoder as shown

Fig. 6. The results of (a) optimal NOMA-AAR, (b) the optimal PA allocation factors, and (c, d) the optimal RT-allocation strategy.

in (13). For this reason, the AAR for x_2 , $\bar{C}_2$, significantly depends on the PA factors, and $\bar{C}_2$ increases when α_1^* decreases (or equivalently α_2^* increases). Moreover, $\bar{C}_1$ significantly depends on the RSS at U_1 . Hence, $\bar{C}_1$ is enhanced when more RTs are allocated to U_1 . Using the above assessments, it is easy to explain the trends of α_1^* and $N_{U_1}^*$. When the total number of RTs increases, the signal quality at users becomes greater, and the system can boost both AARs at two users by allocating more RTs to U_1 and using higher values for α_2^* (or equivalently using lower values for α_1^*). Therefore, α_1^* is a decreasing function of N_{col} , as shown in Fig. 6(b), whereas $N_{U_1}^*$ and $N_{U_2}^*$ are increasing functions of N_{col} , as shown in Fig. 6(c) and 6(d). Moreover, $N_{U_1}^*$ increases more significantly than $N_{U_2}^*$. On the other hand, the increase in transmit power P_S or the decrease in the SIC-imperfection factor ζ also give the same effect on the signal quality at the users.

Hence, the smaller value for α_1^* and higher values for $N_{U_1}^*$ are observed when P_S increases or ζ decreases.

In Fig. 7, we show the effects of the PD factor τ_{PD} and the

Fig. 7. (a) The effect of PD and (b) the effect of reflection amplitude on the optimal NOMA-AAR. Two cases, with and without LoS links, are considered. Other parameters: $P_S = 5$ dBm for Fig. 7(b).

reflection amplitude β_{Rn} on the system performance for two cases: with and without the presence of the LoS links. The results for the non-LoS case are obtained by set the source-to-user distances in (8) by very large values, i.e., $d_{SU_m} \rightarrow \infty$. The optimal NOMA-AAR drops around 0.8 bits/s/Hz when the LoS links are not available. This indicates the importance of the LoS links; however, without the LoS links, the system can achieve a moderately optimal NOMA-AAR with the help of the reflecting link and the multi-path fading channel. When τ_{PD} increases, the total noise at each user increases. Hence, the optimal NOMA-AAR is displayed as a decreasing function of τ_{PD} as shown in Fig. 7(a). With the presence of a multi-path fading channel and/or LoS links, the influence of the PD on the optimal NOMA-AAR can be reduced; however, when the user is blocked and its communication only relies on the reflecting links, the value of τ_{PD} becomes a crucial factor to performance. Fig. 7(b) presents the optimal NOMA-AAR for different values of β_{Rn} . When β_{Rn} increases, the stronger received signal strengths are achieved at the users. Therefore, the optimal NOMA-AAR is shown as an increasing function of β_{Rn} . Moreover, the optimal NOMA-AAR for the case using the IRS is higher than that for the non-IRS case. This indicates the effectiveness of the IRS in improving system performance. Since β_{Rn} only affects the quality of the reflecting links, the influence of low β_{Rn} values can be reduced in the case of the presence of the LoS links or the stronger multi-path fading channel.

V. CONCLUSION

This paper considered an IRS-aided NOMA communication in the presence imperfect SIC. The theoretical results are derived and confirmed by Monte Carlo simulations. The numerical results showed that the NOMA communication achieves a significant enhancement in performance with the help of an IRS, and it outperforms the reference OMA system. The residual interference caused by the imperfect SIC leads to a remarkable degradation in performance. There is an optimal setup for the PA factors and the number of RTs allocated to each user that allows the best equal capacity for both users. The effects of key system parameters on the trends of the optimal setup were studied. For instance, more RTs are allocated to the near user when the number of available RTs increases. Moreover, a higher portion of transmit power is allocated to the far user when the source transmit power increases.

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