

Sum-rate Maximization and Outage Analysis of RIS assisted RSMA System under Electromagnetic Interference

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Abstract—Rate-splitting multiple access (RSMA) and reconfigurable intelligent surfaces (RIS) are key enablers for 6G communications. This study examines a downlink multiple-input single-output system using RIS-assisted RSMA (RIS-RSMA) in the presence of electromagnetic interference (EMI). Although RIS improves wireless propagation through passive beamforming, it can also reflect EMI from ambient sources, which may disrupt the decoding order and limit system performance. To overcome this, we jointly optimize the transmit beamformer and phase shifts of the RIS using an alternating optimization approach. We derive and validate the analytical expressions for the outage probability (OP) and sum rate of the system via Monte Carlo simulations. The obtained results validate that RIS-RSMA reduces the OP by 60% and achieves a higher sum rate of 2.20×10^3 bit/s compared with RIS-assisted NOMA, and mitigates the impact of EMI.

Link to graphical and video abstracts, and to code:
<https://latam.tiieee9.org/index.php/transactions/article/view/10605>

Index Terms—Electromagnetic interference (EMI), reconfigurable intelligent surfaces (RIS), rate splitting multiple access (RSMA), multiple access techniques, alternating optimization (AO), outage probability (OP).

I. INTRODUCTION

EXPONENTIAL growth in connected devices across diverse applications has driven a dramatic increase in user demand and network traffic. To keep pace, wireless systems must continuously evolve, leveraging advanced technologies to enhance SE, maximize network capacity, boost user throughput, minimize latency, and enable ultra-dense connectivity [1]. Based on ITU and 3GPP, the 6G network is widely categorized into three fundamental services: massive machine-type communication, which enables extensive connectivity for millions of devices, mission-critical with more reliable low-latency communication, which ensures millisecond-level latency, and enhancing the mobile broadband, which supports high data rates for a huge number of connected devices [2], [3]. The 6G network requires advanced multiple access techniques that are capable of accommodating densely connected users

while minimizing interference and delivering high data rates [4], [5].

Existing reports reveal that space-division multiple access (SDMA) achieves better SE for nearly orthogonal channels but fails when user ordering is based on channel strength. This limitation gives rise to another multiple access technique, namely non-orthogonal multiple access (NOMA), in which users are ordered based on channel strength to achieve spectral efficiency (SE). However, the complexity increases significantly when an enormous number of user equipment is deployed [6], [7]. To address this limitation, rate-splitting multiple access (RSMA) is introduced to balance complexity and improve the SE of the communication system [8], [9]. In RSMA, the common stream is decoded by all users, while the private streams are decoded by their respective users, enabling robust interference management and enhancing the SE and the energy efficiency (EE) of the system [10].

However, reconfigurable intelligent surfaces (RIS) are metasurfaces developed to control radio-wave propagation in the wireless environment, which in turn maximizes the desired signal and suppresses interference while improving both the SE and EE of the system [11], [12]. By utilizing these benefits, extensive research has been conducted and few works are discussed here. In [13], RIS-RSMA is utilized to provide a reliable and secure transmission framework, and the limitations caused by hardware impairments are discussed. Furthermore, reconfigurable intelligent surfaces (RIS) are widely acknowledged as an essential tool for the development of sixth-generation (6G) wireless communication systems due to their ability to accurately control the propagation of electromagnetic waves [14], [15].

The amalgamation of RSMA and RIS improves both the EE of the system and user fairness [16], [17]. In [18], an RIS-assisted RSMA (RIS-RSMA) framework for uplink transmission is considered, with the goal of maximizing the system sum rate. In [19], the key finding is to initially optimize the active beamformer at the base station (BS) and then consecutively optimize the passive beamformer at the RIS in order to build an energy-efficient system.

Furthermore, [20] presents a performance analysis of the outage probability (OP) using first-layer rate splitting (1L-RS) and compares it with the conventional DF-assisted framework, referred to as DF-RS. The performance benefits of RSMA are demonstrated through the OP results. Despite these improvements, electromagnetic interference (EMI) remains a significant challenge, as it arises from uncontrollable sources

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in the RIS far-field region and can substantially degrade communication performance [21]. While prior work has examined EMI effects in RIS-assisted communication, its impact on RIS-RSMA systems remains unexplored.

In turn, only few works have addressed this problem in the backdrop of RIS-assisted systems. In [22], a novel EMI cancellation approach is proposed to mitigate EMI, which leverages the peculiar time-domain form and considers an optimal passive beamforming technique at RIS. Furthermore, a multi-user RIS-assisted uRLLC paradigm that brings weak channel state information at transmitter (CSIT) and spatially correlated RIS channels under EMI into account was reported in [23]. The impact of EMI on physical-layer secure communication was further investigated in [24], where the secrecy efficiency of RIS-assisted MISO systems was analyzed. Building upon these insights, this work investigates RSMA in RIS-enhanced systems by jointly considering EMI effects and RIS channel modeling. To the best of the authors' knowledge, the effects of EMI on RIS-RSMA systems have not yet been examined in the literature. The primary contributions of this paper are listed as follows

- A RIS-assisted RSMA (RIS-RSMA) system operating in a MISO downlink scenario is considered for analysis in the effects of EMI.
- The performance analysis of sum-rate and OP under EMI is analytically derived and examined via Monte Carlo simulations.
- An ON-OFF selection framework is utilized to regulate passive beamforming and to analyze the OP of the system.
- Given the non-convex structure of the problem, the transmit beamforming matrix and RIS phase shifts are iteratively optimized through an alternating optimization (AO) algorithm to effectively mitigate EMI and to boost the sum rate of the system.
- The system performance is evaluated under random and AO-optimized phase to highlight the effectiveness of the proposed technique.
- To summarize, the RIS-RSMA is benchmarked against the RIS-assisted NOMA (RIS-NOMA) to highlight its performance advantage in EMI-affected environments.

The rest of the work are categorized as follows. At first, section II discusses the channel modeling and transmission architecture of the MISO system under EMI system. Section III develops the sum-rate and OP analysis under EMI conditions. Section IV introduces the AO algorithm to maximize the sum rate and Section V evaluates computational findings which illustrate the effectiveness of the MISO system and Section VI concludes the work.

II. SYSTEM MODEL

In the proposed system, a BS featuring N_t transmit antennas serves K single-antenna users, which is indexed by $k \in \{1, 2, \dots, K\}$, as shown in Fig. 1. The RIS consists of N reflecting elements, indexed by $n \in \{1, 2, \dots, N\}$, ordered in a uniform planar array with $N = N_h \times N_v$, here N_h and N_v refer to the number of elements in both the

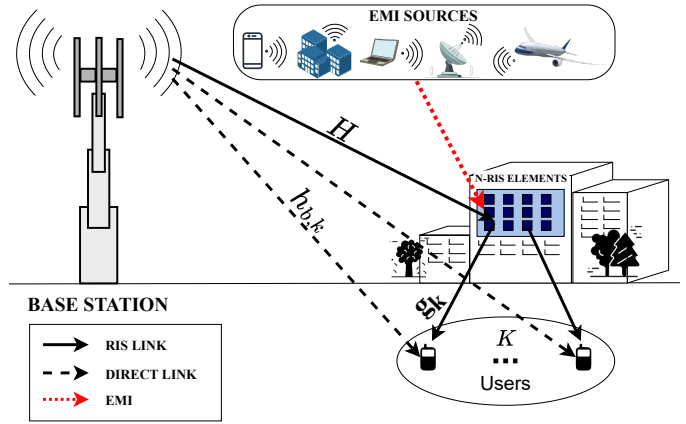


Fig. 1. RIS-RSMA MISO system under EMI.

horizontal and vertical dimensions, respectively. The reflection matrix for user- k is given by $\Theta_k = \text{diag}(\phi_k)$, where $\phi_k = [\zeta_1 e^{j\theta_1}, \zeta_2 e^{j\theta_2}, \dots, \zeta_N e^{j\theta_N}]^T$ is a vector of RIS reflection coefficients. Here, ζ_n and θ_n denote the magnitude and phase shift of the n -th RIS element, respectively, for $n = 1, 2, \dots, N$. In the considered system model, let $\mathbf{H} = [\mathbf{h}_1, \mathbf{h}_2, \dots, \mathbf{h}_N] \in \mathbb{C}^{N_t \times N}$ denote the channel matrix in between the BS and the RIS, where $\mathbf{h}_n$ is the channel vector to the n -th RIS element. The RIS-to-user- k channel is $\mathbf{g}_k = [g_{1,k}, g_{2,k}, \dots, g_{N,k}]^T \in \mathbb{C}^{N \times 1}$, where $g_{n,k}$ represents the channel link between user- k and the n -th RIS element.

A. Transmission Framework

The BS uses 1L-RS in the system model. A shared component $W_{c,k}$ and a private component $W_{p,k}$ make up each user message W_k . The common components of all users are combined to generate a single message W_c . After that, it is encoded using a common codebook into the shared stream s_c , which any user can decode. The streams $s_1, \dots, s_K$ are mapped to the private components separately, with s_k containing data intended only for user- k . The overall transmit symbol vector is $\mathbf{s} = [s_c, s_1, \dots, s_K]^T \in \mathbb{C}^{(K+1) \times 1}$, and its unit covariance is $\mathbb{E}\{\mathbf{s}\mathbf{s}^H\} = \mathbf{I}$. The precoders for the common and private streams are given by $\mathbf{p}_c$ and $\mathbf{p}_k$, respectively. Where $\mathbf{P} = [\mathbf{p}_c, \mathbf{p}_1, \dots, \mathbf{p}_K] \in \mathbb{C}^{N_t \times (K+1)}$ employs linear precoding. Consequently, the BS transmits $\mathbf{x} = \mathbf{p}_c s_c + \sum_{k=1}^K \mathbf{p}_k s_k$. Therefore, the received signal y_k at user- k is given as follows

$$y_k = (\mathbf{H}\Theta_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \mathbf{x} + \mathbf{g}_k^H \Theta_k \boldsymbol{\Omega} + \omega_k, \quad (1)$$

Let $\omega_k \sim \mathcal{CN}(0, \sigma_k^2)$ denotes the thermal noise with variance σ_k^2 , while $\boldsymbol{\Omega} \in \mathbb{C}^{N \times 1}$ denotes the EMI vector. Following the isotropic EMI assumption in [21, Corollary 1], the EMI is modelled as $\boldsymbol{\Omega} \sim \mathcal{CN}(0, A\sigma_E^2 \mathbf{R})$, with σ_E^2 representing the EMI power. Practically, EMI is generated by uncontrolled electromagnetic waves arriving at the RIS from the external environment, most commonly originating from coexisting wireless transmitters and other ambient radiation sources. Illustrative examples of these EMI sources are provided in Fig. 1 for improved comprehension.

B. Channel Modeling

The channels $\{\mathbf{H}, \mathbf{g}_k, \mathbf{h}_{b,k}\}$ are deterministic or stochastic depending on the propagation conditions. The effective channel $\mathbf{h}_k$ observed at the receiver nodes is taken, ignoring the effects of EMI, can be written as [25]

$$\mathbf{h}_k \sim (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H, \quad \forall k \in \{1, 2, \dots, K\}. \quad (2)$$

In the considered model, the channel matrices are distributed as $\mathbf{H} \sim \mathcal{CN}(0, A\beta_{br}\mathbf{R})$, $\mathbf{g}_k \sim \mathcal{CN}(0, A\beta_{rk}\mathbf{R})$, and $\mathbf{h}_{b,k} \sim \mathcal{CN}(0, A\beta_{b,k})$. The corresponding large-scale fading factors for the BS–RIS channels, RIS–user- k and BS–user- k are represented by the parameters β_{br} , β_{rk} , and $\beta_{b,k}$, respectively. With $\mathbf{R}$ representing the RIS correlation matrix [21], each reflecting element of the RIS occupies the area $A = d_H d_V$, where the horizontal dimension is denoted by d_H and the vertical dimension is denoted by d_V , respectively. Furthermore, the magnitude of each channel h_k follows a Rayleigh distribution, as each channel is defined as non-dependent and identically distributed (i.i.d.), i.e., $\mathcal{CN}(0, 1)$.

III. PERFORMANCE ANALYSIS OF RIS-RSMA SYSTEM UNDER ELECTROMAGNETIC INTERFERENCE

This section evaluates the sum rate and derives a closed-form expression for the OP of RIS-RSMA in the presence of EMI.

A. Sum-Rate Behaviour

For every link, it is assumed that the BS acknowledges the CSI at the transmitter is perfect. As derived from (1), the instantaneous rate R_k^c is given as

$$R_k^c = B \log_2(1 + \gamma_k^c), \quad (3)$$

with the received SINR γ_k^c at user- k for decoding s_c is obtained as

$$\gamma_k^c = \frac{|(\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \mathbf{p}_c|^2}{\sum_{j=1}^K |(\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \mathbf{p}_j|^2 + A\sigma_E^2 \mathbf{g}_k^H \mathbf{R} \mathbf{g}_k + \sigma_k^2}. \quad (4)$$

To ensure that every individual user can reliably decode the common stream s_c , the required rate is limited to $R^c = \min\{R_1^c, \dots, R_K^c\}$. Since all users share this rate, it is divided as follows: For decoding $W_{c,k}$, $\sum_{k=1}^K C_k = R^c$, here C_k is the proportion of the common rate given to user- k . After canceling the estimated portion of s_c , user- k decodes its private stream s_k , achieving a target rate of R_k^p is formulated as

$$R_k^p = B \log_2(1 + \gamma_k^p), \quad (5)$$

with the received SINR γ_k^p at user- k for decoding s_k is expressed as

$$\begin{aligned} \gamma_k^p &= \frac{|(\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \mathbf{p}_k|^2}{\sum_{j=1, j \neq k}^K |(\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \mathbf{p}_j|^2 + A\sigma_E^2 \mathbf{g}_k^H \mathbf{R} \mathbf{g}_k + \sigma_k^2}, \\ &= \frac{|(\mathbf{H}\mathbf{G}_k \boldsymbol{\phi}_k + \mathbf{h}_{b,k})^H \mathbf{p}_k|^2}{\sum_{j=1, j \neq k}^K |(\mathbf{H}\mathbf{G}_k \boldsymbol{\phi}_k + \mathbf{h}_{b,k})^H \mathbf{p}_j|^2 + A\sigma_E^2 \mathbf{g}_k^H \mathbf{R} \mathbf{g}_k + \sigma_k^2}. \end{aligned} \quad (6)$$

Here, the vector $\boldsymbol{\phi}_k$ is obtained by extracting the diagonal entries of $\boldsymbol{\Theta}_k$ and has dimension $N \times 1$, while its Hermitian transpose $\boldsymbol{\phi}_k^H$ forms a $1 \times N$ row vector containing the same diagonal elements. Furthermore, $\mathbf{G}_k$ represents an $N \times N$ diagonal matrix composed of the components of the vector $\mathbf{g}_k$. Moreover, the second term in the denominator captures the EMI effect, wherein an increase in the RIS surface area A magnifies the interference and affects the common and the private rates. Furthermore, the total attainable sum-rate for user- k is given by $R_{k,\text{tot}} = C_k + R_k^p$.

B. Outage Analysis in the Presence of EMI

This section investigates the impact of EMI on the RIS-RSMA system under the assumption that the direct link is not available, i.e. $\beta_{b,k} = 0$, which corresponds to $-\infty$ dB. However, continuous phase shift [26] and discrete phase shift [27] in the literature have their own advantages. However, continuous phase shifting does not facilitate the derivation of the OP due to the random nature of the wireless channel, while discrete phase shifting requires dedicated quantized phase shifters, which increase the complexity of the resulting non-convex optimization problem and the overall system cost. Furthermore, to derive a tractable closed-form for the OP, an ON–OFF selection mechanism is employed, where each RIS element is restricted to operate in an active or inactive mode. Let $\mathbf{F} \in \mathbb{C}^{M \times N}$ represent the beamforming matrix, with each row corresponding to a distinct ON–OFF configuration of the RIS. The matrix $\mathbf{F}$ is constructed as $\mathbf{F} = \mathbf{I}_M \otimes \mathbf{z}$, where $\mathbf{z} \in \mathbb{C}^{1 \times D}$ is a unit-norm vector with identical entries, D denotes the number of concurrently active reflecting elements, and $M = N/D$ specifies the total number of available ON–OFF beamforming configuration, such that $N = MD$. Consequently, each row of $\mathbf{F}$ activates a unique subset of RIS elements while maintaining unit norm. For user- k , the reflection vector $\boldsymbol{\phi}_k^H$ is selected as the m^{th} row vector $\mathbf{f}_m$ of the matrix $\mathbf{F}$, where $m = 1, 2, \dots, M$. So as to maximize the instantaneous received SINR. The resulting SINR expression $\tilde{\gamma}_k^b$ for user- k is therefore given by

$$\tilde{\gamma}_k^b \triangleq \max_{m=1, \dots, M} \{\gamma_k^b(\mathbf{f}_m)\}, \quad b \in \{c, p\}, \quad (7)$$

where

$$\gamma_k^c(\mathbf{f}_m) = \frac{|(\mathbf{H}\mathbf{G}_k \mathbf{f}_m + \mathbf{h}_{b,k})^H \mathbf{p}_c|^2}{\sum_{j=1}^K |(\mathbf{H}\mathbf{G}_k \mathbf{f}_m + \mathbf{h}_{b,k})^H \mathbf{p}_j|^2 + A\sigma_E^2 \mathbf{g}_k^H \mathbf{R} \mathbf{g}_k + \sigma_k^2}, \quad (8)$$

$$\gamma_k^p(\mathbf{f}_m) = \frac{|(\mathbf{H}\mathbf{G}_k\mathbf{f}_m + \mathbf{h}_{b,k})^H \mathbf{p}_k|^2}{\left| \sum_{j=1, j \neq k}^K (\mathbf{H}\mathbf{G}_j\mathbf{f}_m + \mathbf{h}_{b,j})^H \mathbf{p}_j \right|^2 + A\sigma_E^2 \mathbf{g}_k^H \mathbf{R} \mathbf{g}_k + \sigma_k^2} \quad (9)$$

Furthermore, an outage is avoided only when the SINR values of the common $\gamma_k^c(\mathbf{f}_m)$ and private messages $\gamma_k^p(\mathbf{f}_m)$ exceed the corresponding threshold levels defined by the required QoS constraints. Consequently, the OP serves as a key performance metric for assessing the reliability of the system under various channel fading conditions. Accordingly, the OP event P_k^{out} is given by

$$P_k^{\text{out}} = 1 - \Pr(\tilde{\gamma}_k^c > \hat{\gamma}_{\text{th}}^c, \tilde{\gamma}_k^p > \hat{\gamma}_{\text{th}}^p), \quad (10)$$

where, $\hat{\gamma}_{\text{th}}^c = 2^{\tilde{R}_c} - 1$ and $\hat{\gamma}_{\text{th}}^p = 2^{\tilde{R}_p} - 1$ denote the SINR thresholds required to decode the common and private messages, respectively, where $\tilde{R}_c$ and $\tilde{R}_p$ represent the corresponding target data rates. The joint probability is denoted by $\Pr\{\cdot, \cdot\}$. By applying the principles of joint probability theory [28], the OP in (10) can be reformulated as follows

$$P_k^{\text{out}} = Q_{\tilde{\gamma}_k^c}(\hat{\gamma}_{\text{th}}^c) + Q_{\tilde{\gamma}_k^p}(\hat{\gamma}_{\text{th}}^p) - Q_{\tilde{\gamma}_k^c, \tilde{\gamma}_k^p}(\hat{\gamma}_{\text{th}}^c, \hat{\gamma}_{\text{th}}^p), \quad (11)$$

Here, $Q_X(\cdot)$ denote marginal cumulative distribution function (CDF) corresponding to random variable (RV) X , while $Q_{X,Y}(\cdot, \cdot)$ stands for combined CDF of the X and Y RVs. As established in Observation 1 of [20], the RVs $\gamma_k^c(\mathbf{f}_1), \gamma_k^c(\mathbf{f}_2), \dots, \gamma_k^c(\mathbf{f}_M)$ are mutually independent. Similarly, the RVs $\gamma_k^p(\mathbf{f}_1), \gamma_k^p(\mathbf{f}_2), \dots, \gamma_k^p(\mathbf{f}_M)$ are also mutually independent. Moreover, for any given index m , the RVs $\gamma_k^c(\mathbf{f}_m)$ and $\gamma_k^p(\mathbf{f}_m)$ are statistically dependent. Nevertheless, the SINR pair $\{\gamma_k^c(\mathbf{f}_i), \gamma_k^p(\mathbf{f}_i)\}$ is statistically independent of $\{\gamma_k^c(\mathbf{f}_j), \gamma_k^p(\mathbf{f}_j)\}$ for all $i \neq j$, where $i, j = 1, 2, \dots, M$. Therefore, the OP in (11) can be reformulated as

$$P_k^{\text{out}} = \prod_{m=1}^M Q_{\gamma_k^c(\mathbf{f}_m)}(\hat{\gamma}_{\text{th}}^c) + \prod_{m=1}^M Q_{\gamma_k^p(\mathbf{f}_m)}(\hat{\gamma}_{\text{th}}^p) - \prod_{m=1}^M Q_{\gamma_k^c(\mathbf{f}_m), \gamma_k^p(\mathbf{f}_m)}(\hat{\gamma}_{\text{th}}^c, \hat{\gamma}_{\text{th}}^p). \quad (12)$$

In addition, the marginal and joint distributions of $\gamma_k^c(\mathbf{f}_m)$ and $\gamma_k^p(\mathbf{f}_m)$ are identical for all m , as a result, the OP in (12) can be rewritten as

$$P_k^{\text{out}} = \left[Q_{\gamma_k^c(\mathbf{f}_m)}(\hat{\gamma}_{\text{th}}^c) \right]^M + \left[Q_{\gamma_k^p(\mathbf{f}_m)}(\hat{\gamma}_{\text{th}}^p) \right]^M - \left[Q_{\gamma_k^c(\mathbf{f}_m), \gamma_k^p(\mathbf{f}_m)}(\hat{\gamma}_{\text{th}}^c, \hat{\gamma}_{\text{th}}^p) \right]^M. \quad (13)$$

Consequently, the following is the CDF of the common and private SINRs that correspond to the beamforming vectors $\mathbf{p}_c$ and $\mathbf{p}_k$, equation (13) can be expressed as [20]

$$Q_{\gamma_k^c(\mathbf{f}_m), \gamma_k^p(\mathbf{f}_m)}(\hat{\gamma}_{\text{th}}^c, \hat{\gamma}_{\text{th}}^p) = 1 - \frac{2}{\Gamma(D)} (D\hat{\gamma})^{\frac{D}{2}} K_D \left(2\sqrt{D\hat{\gamma}} \right), \quad (14)$$

where $\hat{\gamma} > 0$ is defined as $\hat{\gamma} = \min(\delta_1(\hat{\gamma}_{\text{th}}^c), \delta_2(\hat{\gamma}_{\text{th}}^p))$, with

$$\delta_1(\hat{\gamma}_{\text{th}}^c) = \frac{\hat{\gamma}_{\text{th}}^c (A\sigma_E^2 R + \sigma_k^2)}{\mathbf{p}_c^H \mathbf{p}_c - \hat{\gamma}_{\text{th}}^c \sum_{j=1}^K \mathbf{p}_j^H \mathbf{p}_j}, \quad (15a)$$

$$\delta_2(\hat{\gamma}_{\text{th}}^p) = \frac{\hat{\gamma}_{\text{th}}^p (A\sigma_E^2 R + \sigma_k^2)}{\mathbf{p}_k^H \mathbf{p}_k - \hat{\gamma}_{\text{th}}^p \sum_{j=1}^K \mathbf{p}_j^H \mathbf{p}_j}, \quad (15b)$$

where $K_v(\cdot)$ denotes v^{th} order Bessel function and $\Gamma(\cdot)$ pointing the Gamma function. Therefore, the marginal CDFs $Q_{\gamma_k^c(\mathbf{f}_m)}(\gamma_k^c)$ and $Q_{\gamma_k^p(\mathbf{f}_m)}(\gamma_k^p)$ are obtained by taking the limits $\hat{\gamma}_{\text{th}}^c \rightarrow \infty$ and $\hat{\gamma}_{\text{th}}^p \rightarrow \infty$ provides the joint CDF given in (14),

$$Q_{\gamma_k^c(\mathbf{f}_m)}(\gamma_k^c) = 1 - \frac{2}{\Gamma(D)} (D\delta_1(\hat{\gamma}_{\text{th}}^c))^{\frac{D}{2}} K_D \left(2\sqrt{D\delta_1(\hat{\gamma}_{\text{th}}^c)} \right), \quad (16a)$$

$$Q_{\gamma_k^p(\mathbf{f}_m)}(\gamma_k^p) = 1 - \frac{2}{\Gamma(D)} (D\delta_2(\hat{\gamma}_{\text{th}}^p))^{\frac{D}{2}} K_D \left(2\sqrt{D\delta_2(\hat{\gamma}_{\text{th}}^p)} \right). \quad (16b)$$

Therefore, the OP of user- k is obtained by substituting (14)–(16) into (13). Decoding common and private messages jointly makes the decoding reliable and enhances the OP performance.

IV. SUM-RATE MAXIMIZATION OF RIS-RSMA IN PRESENCE OF EMI

Owing to its passive structure, the RIS reflects both the desired BS signal and the EMI present in the environment [21], [22]. These uncontrollable EMI signals may combine non-coherently with the desired signal at the receiver, leading to destructive interference, which deteriorates the overall signal quality and reduces the achievable sum-rate.

1) *Problem formulation:* Our target focuses on sum-rate maximization. In particular, the transmit beamformers $\mathbf{p}_c$ and $\mathbf{p}_k$ are optimized jointly with the RIS phase shift matrix θ_n . Moreover, EMI is a random variable and its interference power is convex in nature. By ignoring the constant terms and assuming continuous reflection amplitude with adjustable phase shifts, the problem $\mathcal{P}_1$ is formulated as

$$\mathcal{P}_1 : \max_{\mathbf{p}_c, \{\mathbf{p}_k\}, \{\theta_n\}} \left| (\mathbf{H}\mathbf{G}_k\mathbf{g}_k + \mathbf{h}_{b,k})^H \mathbf{p}_c \right|^2 + \left| (\mathbf{H}\mathbf{G}_k\mathbf{g}_k + \mathbf{h}_{b,k})^H \mathbf{p}_k \right|^2 \quad (17a)$$

$$\text{s.t. C1: } \|\mathbf{p}_c\|^2 + \sum_{k=1}^K \|\mathbf{p}_k\|^2 \leq P_t, \quad (17b)$$

$$\text{C2: } \text{tr}(\mathbf{P}\mathbf{P}^H) \leq P_t, \quad (17c)$$

$$\text{C3: } \sum_{k \in \mathcal{K}} C_k \leq R^c, \quad \forall k, \quad (17d)$$

$$\text{C4: } |\zeta_n| = 1, \quad \forall n, \quad (17e)$$

$$\text{C5: } \mathbb{E}[\|\mathbf{\Omega}\|^2] = A\sigma_E^2 \text{tr}(\mathbf{R}) \leq \mathbf{\Omega}_{\text{th}}. \quad (17f)$$

where, C3 ensures that overall common rate R^c is distributed across all users, C4 applies the unit-modulus constraint of RIS coefficients, and to persist, C1 and C2 represent the power allocation requirements at the BS. C5 enforces an upper bound on the average EMI power to ensure it remains within an acceptable limit. The optimization problem $\mathcal{P}_1$ is difficult to solve because of its intrinsic non-convex nature with respect to the variables $\{\mathbf{p}_c, \mathbf{p}_k, \theta_n\}$. The beamformer structure at the BS and the RIS element adjustment are the two interrelated sub-problems that $\mathcal{P}_1$ is restructured in order to address this challenge. As described in the next section, the AO framework is then used to solve these sub-problems one after the other.

Algorithm 1 Alternating Optimization

- 1: **Input:** P_t , $\alpha = 0.5$, σ_E^2 , Ω_{th}
 - 2: **Initialize:** $(\theta_n^q, \bar{\mathbf{p}}_c^q, \bar{\mathbf{p}}_k^q)$
 - 3: Set iteration index $q = 1$
 - 4: **repeat**
 - 5: $\angle((\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k)^H \bar{\mathbf{p}}_c) = \angle(\mathbf{h}_{b,k}^H \bar{\mathbf{p}}_c) \triangleq \varphi_0$
 - 6: Given $(\bar{\mathbf{p}}_c^q, \bar{\mathbf{p}}_k^q)$, update θ_n^q by solving (18a) and set $\theta_n^{q+1} = \theta_n^*$
 - 7: Given θ_n^{q+1} , update $(\bar{\mathbf{p}}_c^q, \bar{\mathbf{p}}_k^q)$ and set $\bar{\mathbf{p}}_c^{q+1} = \bar{\mathbf{p}}_c^*$ and $\bar{\mathbf{p}}_k^{q+1} = \bar{\mathbf{p}}_k^*$
 - 8: Update $q \leftarrow q + 1$
 - 9: **until** the relative change in the objective value is smaller than $\varepsilon > 0$
 - 10: **return** $(\theta_n^*, \bar{\mathbf{p}}_c^*, \bar{\mathbf{p}}_k^*)$
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2) *Problem Solving via Alternating Optimization:* The both transmit precoders and the RIS configuration matrix is handled using the AO technique. One variable in the set $\{\mathbf{p}_c, \mathbf{p}_k, \theta_n\}$ is updated at each iteration, while the others remain unchanged. In this way, until the process converges, the RIS coefficients and the BS beamforming vectors are refined alternately. As discussed earlier, the problem $\mathcal{P}_1$ is partitioned into two simpler optimization tasks, which makes the overall design more tractable. The first subproblem is given in $\mathcal{P}_{1.1}$. Accordingly, the precoders are written as $\mathbf{p}_c = \sqrt{P_t} \bar{\mathbf{p}}_c$ and $\mathbf{p}_k = \sqrt{P_t} \bar{\mathbf{p}}_k$. For fixed transmit beamforming vectors $\{\bar{\mathbf{p}}_c, \bar{\mathbf{p}}_k\}$, the problem $\mathcal{P}_1$ is transformed into RIS phase shift optimization $\mathcal{P}_{1.1}$, and expressed as

$$\mathcal{P}_{1.1} : \max_{\theta_n} \left| (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \bar{\mathbf{p}}_c \right|^2 + \left| (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \bar{\mathbf{p}}_k \right|^2 \quad (18a)$$

$$\text{s.t. } 0 \leq \theta_n \leq 2\pi, \quad \forall n \in \{1, 2, \dots, N\}. \quad (18b)$$

To handle the non-convex nature of the problem, the triangle inequality [11] is applied to derive an upper bound for the numerator of the common stream $\left| (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \bar{\mathbf{p}}_c \right|$, which is written as

$$\begin{aligned} \left| (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H \bar{\mathbf{p}}_c \right| &= \left| (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k)^H \bar{\mathbf{p}}_c + \mathbf{h}_{b,k}^H \bar{\mathbf{p}}_c \right| \\ &\stackrel{(a)}{\leq} \left| (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k)^H \bar{\mathbf{p}}_c \right| + \left| \mathbf{h}_{b,k}^H \bar{\mathbf{p}}_c \right|. \end{aligned} \quad (19)$$

where (a) follows from triangle inequality, and equality holds precisely when $\arg((\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k)^H \bar{\mathbf{p}}_c) = \arg(\mathbf{h}_{b,k}^H \bar{\mathbf{p}}_c) \triangleq \varphi_0$. In the following, it is demonstrated that θ_n can be appropriately chosen such that condition (a) holds with equality, while also satisfying the phase shift constraint in (18b). Let $\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k \bar{\mathbf{p}}_c = \mathbf{u}^H \mathbf{a}$, where $\mathbf{u} = [e^{j\theta_1}, e^{j\theta_2}, \dots, e^{j\theta_N}]^H$ and $\mathbf{a} = \text{diag}(\mathbf{g}_k)^H \mathbf{H} \bar{\mathbf{p}}_c$. With (19), problem (18a) is equivalent to

$$\max_{\mathbf{u}} |\mathbf{u}^H \mathbf{a}|^2 \quad (20a)$$

$$\text{s.t. } |\mathbf{u}_n| = 1, \quad \forall n \in \{1, \dots, N\}, \quad (20b)$$

$$\arg(\mathbf{u}^H \mathbf{a}) = \varphi_0. \quad (20c)$$

The problem above can be solved directly with the formula $\mathbf{u}^* = e^{j(\varphi_0 - \arg(\mathbf{a}))} = e^{j(\varphi_0 - \arg(\text{diag}(\mathbf{g}_k)^H \mathbf{H} \bar{\mathbf{p}}_c))}$. Consequently, the n -th phase shift of the RIS θ_n^* can be determined as

$$\theta_n^* = \varphi_0 - \arg(g_{k,n}^H) - \arg(\mathbf{h}_{b,k}^H \bar{\mathbf{p}}_c). \quad (21)$$

The ideal phase shift is θ_n^* , whereas the n -th element of the vector $\mathbf{g}_k$ and the n -th row of the matrix $\mathbf{H}$ are represented by $g_{k,n}^H$ and $\mathbf{h}_n^H$, respectively. Notably, the product $\mathbf{h}_n^H \bar{\mathbf{p}}_c$ characterizes the effective channel encountered by the n -th passive element at the RIS by encapsulating the combined effect of the transmit beamforming and the RIS-assisted channel. Similarly, the same optimization framework is used to compute the private precoder $\bar{\mathbf{p}}_k$ for user- k , which is assessed according to the user's effective channel conditions.

Consequently, after updating the RIS phases, the expressions in (4) and (6) can be written as γ_k^c and γ_k^p , respectively, which are given as

$$\gamma_k^c = \frac{\sum_{n=1}^N |g_{k,n}^H \mathbf{h}_n^H \bar{\mathbf{p}}_c| e^{j\angle(\mathbf{h}_{b,k}^H \bar{\mathbf{p}}_c)} + |\mathbf{h}_{b,k}^H \bar{\mathbf{p}}_c|}{\sum_{j=1}^K \left| \sum_{n=1}^N g_{k,n}^H e^{j\theta_n^*} \mathbf{h}_n^H \bar{\mathbf{p}}_j + \mathbf{h}_{b,k}^H \bar{\mathbf{p}}_j \right|^2 + A\sigma_E^2 g_k^H R g_k + \sigma_k^2}, \quad (22a)$$

$$\gamma_k^p = \frac{\sum_{n=1}^N |g_{k,n}^H \mathbf{h}_n^H \bar{\mathbf{p}}_k| e^{j\angle(\mathbf{h}_{b,k}^H \bar{\mathbf{p}}_k)} + |\mathbf{h}_{b,k}^H \bar{\mathbf{p}}_k|}{\sum_{\substack{j=1 \\ j \neq k}}^K \left| \sum_{n=1}^N g_{k,n}^H e^{j\theta_n^*} \mathbf{h}_n^H \bar{\mathbf{p}}_j + \mathbf{h}_{b,k}^H \bar{\mathbf{p}}_j \right|^2 + A\sigma_E^2 g_k^H R g_k + \sigma_k^2}. \quad (22b)$$

Next, the second subproblem is to optimize the transmit beamforming for the given θ_n^* in (21). It is well established that singular value decomposition (SVD) provides the optimal solution for designing the transmit precoder for the common stream. The combined effective channel matrix is given by $\boldsymbol{\Psi}^H = [\Psi_1^H, \Psi_2^H, \dots, \Psi_K^H]^T$, where $\Psi_k^H = (\mathbf{H}\boldsymbol{\Theta}_k \mathbf{g}_k + \mathbf{h}_{b,k})^H$ with $k \in \{1, 2, \dots, K\}$. Applying SVD to the effective channel gives $\boldsymbol{\Psi}^H = \mathbf{U} \boldsymbol{\Sigma} \mathbf{V}^H$, where $\mathbf{V}$ contains the right singular vectors. The common precoder is defined as $\bar{\mathbf{p}}_c^* = \mathbf{u}_1 \sqrt{(1 - \alpha)P_t}$, with $\mathbf{u}_1$ denoting the dominant right singular vector of $\boldsymbol{\Psi}^H$, and $0 \leq \alpha \leq 1$ represents power

allocated for the common and private streams. Furthermore, for the private streams, maximum ratio transmission (MRT), is used, wherein each private precoder $\bar{\mathbf{p}}_k^*$ is aligned with the corresponding user's effective channel and is given by $\bar{\mathbf{p}}_k^* = \frac{\mathbf{H}\Theta_k \mathbf{g}_k + \mathbf{h}_{b,k}}{\|\mathbf{H}\Theta_k \mathbf{g}_k + \mathbf{h}_{b,k}\|} \sqrt{\frac{\alpha P_t}{2}}$, where $\frac{\alpha P_t}{2}$ denotes the individual power allocated to each private stream [5].

In summary, Algorithm 1 is employed to maximize the sum-rate of user- k . With the transmit beamformer fixed, the RIS elements are assigned to user- k . By referring to (21), the RIS phase shift θ_n is updated. Consequently, (22a) and (22b) are obtained with the optimal θ_n^* . Once θ_n^* is fixed, the common and private precoders $\mathbf{p}_c$ and $\mathbf{p}_k$ are updated, where the common precoder is designed using SVD and the private precoder using MRT, yielding the optimal vector solutions $\bar{\mathbf{p}}_c^*$ and $\bar{\mathbf{p}}_k^*$ with the complexity of $O(qN^2 + 3qN + q)$. Here q denotes the number of iterations required for the transmit beamformer to maximize the sum rate when N is large. This process is repeated iteratively until the improvement becomes smaller than a predefined threshold $\epsilon > 0$. The method is said to converge when the objective function drops below the specified threshold. If the objective function is convex, convergence is guaranteed. However, in the non-convex case, the algorithm may converge to a local optimum. Nevertheless, the AO algorithm proves highly effective for enhancing the sum-rate, owing to its simplicity and its ability to efficiently tackle non-convex optimization problems.

V. RESULTS AND DISCUSSIONS

To evaluate the impact of EMI on the RIS-RSMA system, simulations are carried out for two scenarios: without RIS and with RIS. In both scenarios, downlink transmit beamforming is applied at the BS. However, in the RIS-assisted case, the RIS additionally provides passive beamforming through its phase shifts, which are considered in two configurations: random and AO-optimized. NOMA is also included as a baseline scheme for benchmarking the RSMA performance. The simulation setup consists of a BS, an RIS, and K users located at $(0, 0, 10\text{ m})$, $(98\text{ m}, 2\text{ m}, 10\text{ m})$, and $(100\text{ m}, 0, 0)$, respectively, in a three-dimensional coordinate system. The communication channels account for both large-scale and small-scale fading, and the corresponding parameter values are provided in Table I. Furthermore, to characterize the impact of EMI, we introduce the metric $\rho = \frac{P\beta_1}{\sigma_{\text{EMI}}^2}$, which denotes the ratio between the received signal power and the EMI power at each reflecting element of the RIS [21].

Fig. 2, illustrates the convergence behavior of the proposed RIS-RSMA scheme. The achievable sum rate is presented as a function of the number of iterations for different numbers of reflecting elements N , with $\rho = 20$. The results indicate that the AO-optimized phase design converges after approximately 35–40 iterations for all considered values of N . In addition, the AO-optimized phase design is compared with a random phase shift scheme, which exhibits inferior performance with an $O(N)$ scaling behavior. Furthermore, the achievable sum rate increases with larger values of N , as a greater number of reflecting elements provides higher beamforming gain.

In Fig. 3 and Fig. 4, the numerical results are compared with the Monte Carlo simulations to validate the analytical

TABLE I
SIMULATION PARAMETERS

Key Parameter	Symbol	Value
Number of transmit antennas at the BS	N_t	8
Number of receiving antennas at user- k	K	4
Path loss exponent for the BS-RIS	β_{br}	-48 dB
Path loss exponent for the RIS-user- k	β_{rk}	-38 dB
Path loss exponent for the direct path	$\beta_{b,k}$	-80 dB, -100 dB
Path loss exponent for the reference distance of 1 m	β_0	-30 dB
Total number of RIS elements	N	100
Bandwidth	BW	1 MHz
BS power budget	P_t	23 dBm
Thermal noise variance for the cascaded channel	σ_k^2	-114 dBm
EMI noise variance at the RIS	σ_E^2	-35 dBm

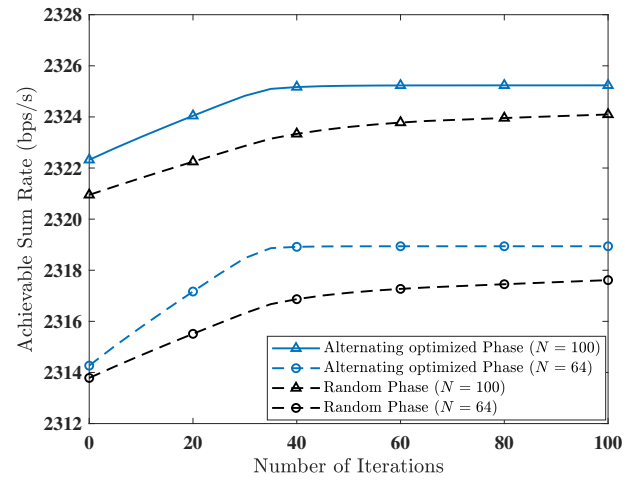


Fig. 2. Convergence pattern of the alternating optimization algorithm.

model. The effect of imperfect CSIT in RSMA is considered with $\alpha = 0.5$, where α denotes the CSIT scaling factor [3].

In Fig. 3, the numerical evaluation of the achievable sum rate aligns closely with the simulation results throughout N values, showing that the analytical model accurately reflects the behavior of the RIS-RSMA system. In EMI-dominated scenarios, RSMA mitigates severe performance degradation by adaptively managing interference through partial decoding, which conventional NOMA does not offer. RSMA also remains robust under imperfect CSIT. Therefore, imperfect CSIT is considered for RSMA and compared with the corresponding perfect-CSIT RSMA and NOMA cases under the adopted AO framework. The results show that RSMA with imperfect CSIT still outperforms perfect-CSIT NOMA due to its flexible power-splitting capability, highlighting its suitability for practical IoT systems with a large number of connected devices.

Similarly, Fig. 4 indicates a strong consistency between the numerical and simulation curves for the OP. The impact of imperfect CSIT on RSMA is studied and compared to the perfect-CSIT RSMA and NOMA benchmarks. The OP of RSMA with imperfect CSIT remains lower than that of conventional NOMA, highlighting its flexible interference management. This indicates that RSMA can overcome path

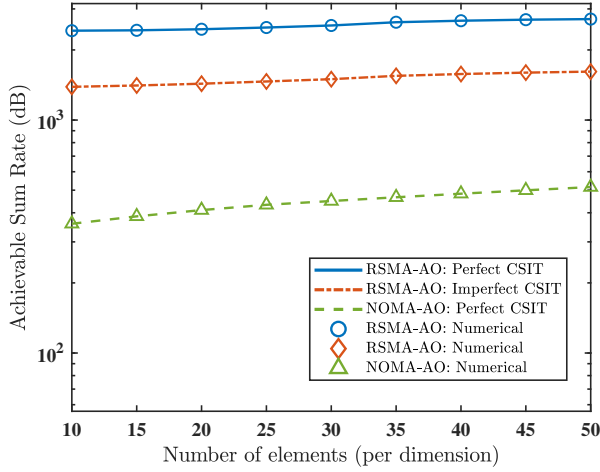


Fig. 3. Numerical validation of the achievable sum-rate performance for RSMA and NOMA with $\rho = 20$.

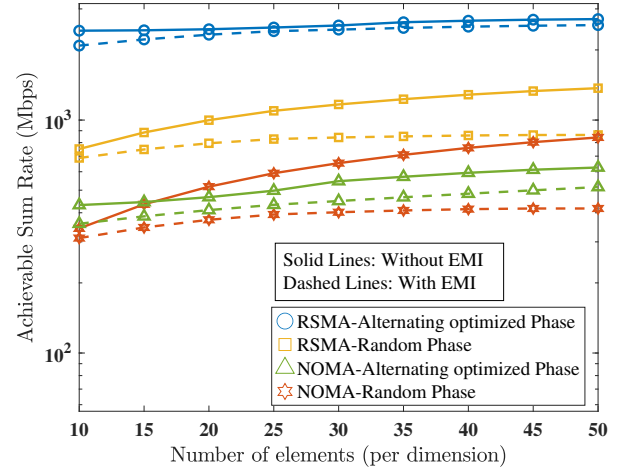


Fig. 5. Comparison of RSMA and NOMA achievable sum-rate for varying RIS elements with $p_c = 0.8$, $p_k = 0.2$, and $\rho = 20$.

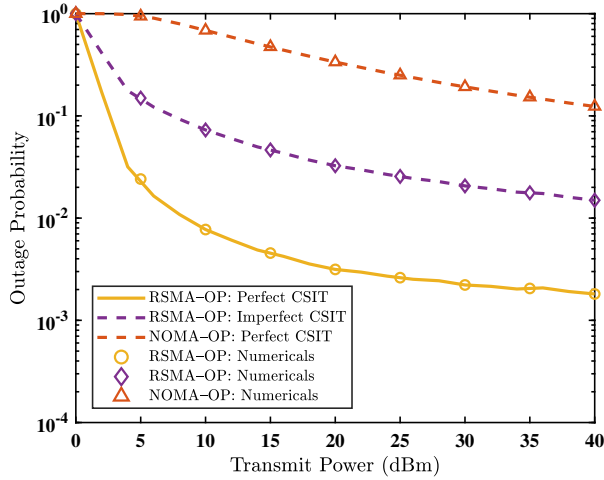


Fig. 4. Numerically validated outage analysis of RSMA and NOMA for $N = 64$.

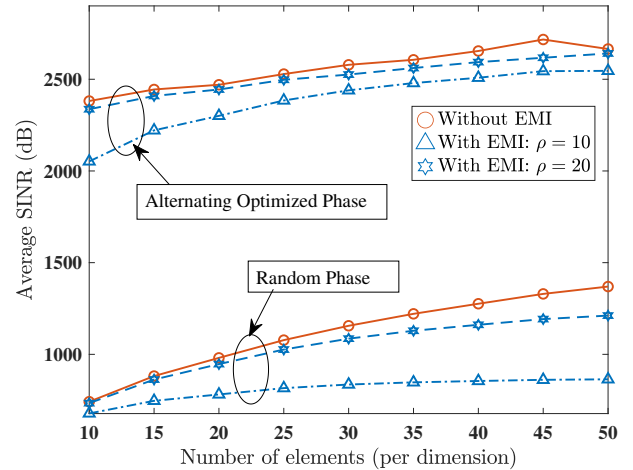


Fig. 6. Average SINR of the RIS-RSMA system, without considering the direct link, is evaluated for different values of N under various EMI scenarios.

loss and mitigate the severe effects of EMI even when 80% of EMI is present in the environment. Thus, RSMA is a promising candidate for cell-edge users and uRLLC applications in 6G communications. Thus the analytical and simulated results for both achievable sum rate and OP highlight the robustness of the model and confirm its effectiveness for evaluating the performance of the system considered.

Fig. 5 presents a performance comparison between random and AO-optimized phase shifts for both RIS-RSMA and RIS-NOMA schemes. To facilitate a fair comparison, RIS-NOMA is adopted as a reference case, noting that RIS-RSMA reduces to the conventional RIS-NOMA model when the common stream is turned off (i.e. $p_c = 0$) and the entire transmit power is distributed to the private stream [2]. The results clearly demonstrate that the AO-optimized phase significantly outperforms the random phase configuration under both EMI and no-EMI conditions due to its effective handling of coupled optimization variables. Furthermore, EMI introduces a noticeable degradation in system performance across all considered

schemes, as the unwanted EMI impinging on the RIS give rise to destructive interference at the receiver. Furthermore, the result ensure that RIS-RSMA provides effective inter-user interference management and also maintains strong resilience against EMI which eventually stabilizes the achievable sum rate of the system.

Fig. 6 illustrates the performance of average SINR as a function of N RIS elements. The initial analysis is made without the direct channel by setting $\beta_{b,k} = 0$ dB. In the simulations, ρ is considered to take values of $\{10, 20\}$ dB, with a system bandwidth of 1 MHz. Notably, the SINR gap between different ρ values remains nearly constant, indicating robustness against EMI. Besides this, it can be observed that the AO-optimized phase consistently achieves superior SINR performance across all values of N , while the random phase is included only for benchmarking. In contrast, the performance gap for the random phase widens with increasing ρ , indicating its limited robustness to EMI. By suppressing the interference

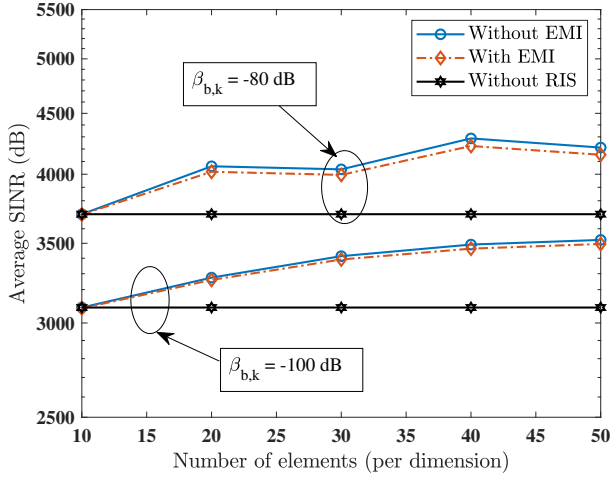


Fig. 7. Average SINR for varying N with AO-optimized phase in the presence of the direct path.

over the random phase shift, the AO-optimized phase delivers up to a fivefold improvement in SINR, which greatly improves overall system performance even though both configurations show N^2 scaling behavior.

Fig. 7 illustrates the impact of average SINR with an AO-optimized RIS phase for N values by setting $\rho = 20$ for $\beta_{b,k} = -100$ dB and $\beta_{b,k} = -80$ dB [21]. In order to analyze the importance of the RIS, the direct path is considered and analyzed in the system. Although the direct path is included to match the phase with the cascaded link, a key finding reveals that the presence of a stronger direct path affects the average SINR, as evidenced from Fig. 7. Another important parameter, which tells about OP is Fig. 8, which illustrates the OP as a function of the transmit power for $N = 64$ and $N = 100$ reflecting elements under both EMI and no-EMI conditions. The results show that as the number of RIS elements rises, the probability of an outage decreases exponentially. This is because RIS effectively improves channel quality by adjusting the phase shift, which strengthens the SINR and attains the desired targets. According to the findings for RIS-NOMA and RIS-RSMA, the OP increases with smaller RIS elements and decreases exponentially with larger RIS elements.

Fig. 9 illustrate the OP varies with transmit values, assuming a target rate of $\hat{\gamma}_{th}^c = \hat{\gamma}_{th}^p = 1$ bit/s. With the common stream assigned $\mathbf{p}_c = [0.6, 0.8]$ and the private streams assigned $\mathbf{p}_k = [0.4, 0.2]$, the results further examine the OP behavior of RIS-RSMA transmission under various power allocation variables. It is observed that increasing $\mathbf{p}_k$ reduces the OP by improving the channel conditions experienced by individual cell-edge users. However, allocating less power to the common message $\mathbf{p}_c$ can adversely affect its decodability if a proper balance between the common and private messages is not maintained, which may consequently impact cell-edge users. Moreover, the minimum OP is achieved at $\mathbf{p}_c = 0.8$ and $\mathbf{p}_k = 0.2$, where the OP decreases to below 10^{-4} .

Fig. 10 highlights the impact of inter-element spacing on the performance of RIS-RSMA under EMI conditions. This emphasizes the need for an RIS configuration to minimizes spatial

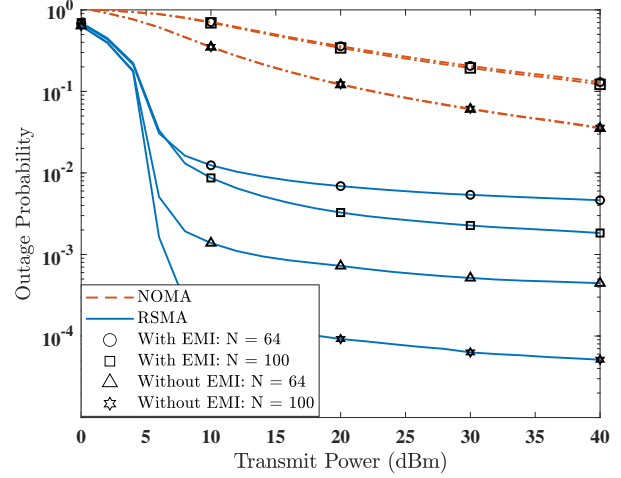


Fig. 8. Outage probability results for RSMA versus NOMA when $N = 64$ and $N = 100$.

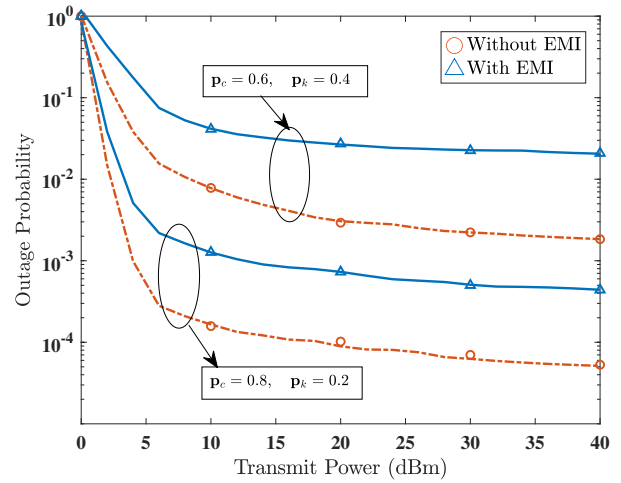


Fig. 9. Evaluation of outage performance for different transmit power levels in the presence and absence of EMI, with $\hat{\gamma}_{th}^c = \hat{\gamma}_{th}^p = 1$ bit/s.

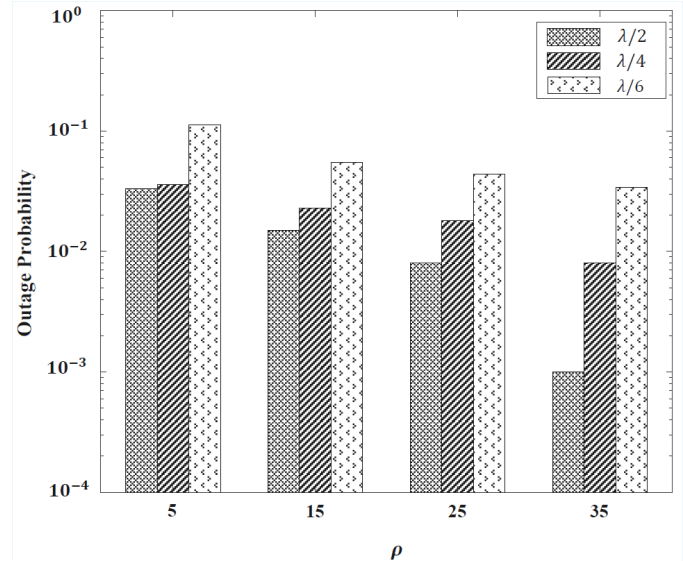


Fig. 10. Impact of RIS inter-element spacing, set at $(\lambda/2, \lambda/4, \lambda/6)$ on outage probability for varying ρ

correlation by enhancing interference mitigation capabilities. By changing the inter-element spacing λ while keeping a fixed number of elements N , the analysis examines the effect of changing the RIS area on the OP. Specifically, the considered inter-element spacings are $d_H = d_V \in \{\frac{\lambda}{2}, \frac{\lambda}{4}, \frac{\lambda}{6}\}$. A notable observation is that setting the inter-element spacing to $\lambda/2$ results in a significant decrease in the OP, whereas further reducing the spacing causes degradation as the number of RIS elements increases. Furthermore, two distinct behaviors are observed based on the values of ρ . Under strong EMI conditions (i.e., lower ρ), a compact RIS configuration with reduced inter-element spacing enhances interference suppression. Conversely, under weaker EMI (i.e., higher ρ), wider spacing improves performance by maximizing passive beamforming gain.

VI. CONCLUSION

In the presence of EMI, this study examines a MISO downlink system using RIS-RSMA. To improve the attainable sum rate, the RIS phase shifts and the transmit beamforming matrix are designed. An AO framework is used to solve the resulting non-convex optimization problem. Both the achievable sum rate and the OP have closed-form formulations that take EMI effects into consideration. According to simulation studies, the suggested RIS-RSMA technique routinely performs better than the RIS-NOMA with feasible sum rate and OP. This approach will be extended to STAR-RIS-assisted RSMA systems in future research.

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