Propagation Modeling and Analysis of Reconfigurable Intelligent Surfaces for Indoor and Outdoor Applications in 6G Wireless Systems

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Abstract—In 5G networks, the enhanced mobile broadband, massive Internet-of-things, and ultra-reliable and low latency communications are defined as three use-cases with diverse and strict quality-of-service (QoS) requirements. In order to cope with these objectives, more radical physical layer concepts are needed to comprehend the potential requirements in future wireless systems. This has pushed researchers to look into new paradigms beyond 5G and 6G wireless systems are conceptualized in recent years. One of the promising technologies is the so-called reconfigurable intelligent surfaces (RISs) which convert the wireless channel into an intelligent transmit entity by manipulating the impinging waves using artificial elements in real time. In this paper, the potential benefits of using RISs are investigated for indoor/outdoor setups and various frequency bands (from 2 to 100 GHz). Many RIS-assisted communication scenarios are studied to ensure reliable transmission in deteriorated channels. First, a general propagation model with a single RIS is considered and the effect of the total number of reflecting elements on the probabilistic distribution of the received signal-to-noise ratio and error performance is investigated. Also for this case, the path loss exponent is derived for below and above 6 GHz by considering empirical path loss models. Furthermore, propagation models with multiple RISs are developed and analyzed for indoor and outdoor non line-of-sight (NLOS) scenarios. The bit error performance is evaluated and the relation between error performance and the number of reflecting elements of RISs is determined. RIS selection strategies are also investigated for systems equipped with multiple RISs for the first time. Through extensive simulations, it is demonstrated that the RIS-assisted systems provide promising solutions for indoor/outdoor scenarios at various operating frequencies and exhibit significant results in error performance and achievable data rates even in the presence of system imperfections such as limited range phase adjustment and imperfect channel phase estimation at RISs.

Index Terms—Reconfigurable intelligent surfaces, metasurfaces, path loss analysis, error performance analysis.

I. INTRODUCTION

One of the areas that has made the most progress as a result of the digitalization experienced by human life in recent decades is wireless communications [1]. Due to the enormous increase in data traffic, existing communication systems have faced challenges in providing high quality of service. Both academia and industry have carried out intensive research activities in order to overcome the deficiencies of legacy transmission concepts. All these efforts and targets have attracted the attention of the wireless community in recent times, most notably within the context of fifth-generation (5G) wireless networks. 3rd Generation Partnership Project (3GPP) Release 15, which is the first full set of 5G standard, has been completed in 2018. This initial 5G standard enables high data rate, ultra-reliability and low latency using novel technologies, such as scalable orthogonal frequency-division multiplexing (OFDM) numerologies, millimeter-wave (mmWave) wireless communications and massive multiple-input multiple-output (MIMO) systems [2]. Massive machine type communications, enhanced mobile broadband, and ultra-reliable and low latency communications are defined as three use-cases with diverse requirements and applications in 5G wireless networks. Beyond all these approaches, more radical physical layer concepts are needed to comprehend the potential requirements of future networks due to sophisticated applications with extremely high quality-of-service (QoS) requirements, such as augmented reality, massive Internet-of-everything, and autonomous vehicles. Despite the high expectations, 5G has not brought these sophisticated applications into reality to date [3]. This has pushed researchers to look new paradigms beyond 5G and start to conceptualize sixth-generation (6G) wireless networks. Within the frame of 6G studies, it is aimed to design systems that are expected to have substantial differences from previous generations by achieving a radical transformation in wireless networks.

In light of the above, it is also essential to develop flexible, versatile, and inclusive physical layer solutions in order to support potential requirements of 6G wireless networks. Although researchers have put forward promising solutions for beyond 5G, including index modulation schemes [4], non-orthogonal multiple access [5], and alternative waveforms [6], the overall progress has been still relatively slow in terms of QoS, reliability and security.

In radio communication environments, one of the most deteriorating characteristics of the channel is its random and dynamic nature. However, when the spectrum shortage below 6 GHz frequency bands has taken into account, it is inevitable to migrate to the mmWave (30-100 GHz) and sub-mmWave (above 100 GHz) bands. In higher frequencies above 6 GHz, the signal will be more susceptible to blockage and interference, since collecting high energy in the receiver will be
difficult because of small antenna sizes. Therefore, the severe signal attenuation and blockage prevent mmWaves to reach long distances. In order to ensure reliable wireless transmission and to alleviate these disruptive effects at high frequencies, we need to overcome the negative and uncontrollable effects of the wireless channel.

Most of the promising methods for next-generation wireless networks are mainly based on the intelligence or reconfigurability of the building blocks of communication systems. Recently, reconfigurable intelligent surfaces (RISs) have been put forward by researchers to enable the control of wireless environments via their unique and effective functionalities, such as wave absorption, anomalous reflection, polarized reflection, wave splitting, wave focusing, and phase modification \((7-9)\). Recent results have revealed that these attractive electromagnetic functionalities are possible without complex operations such as decoding, encoding, and radio frequency (RF) processing and the communication system performance can be enhanced by exploiting the implicit randomness of wireless propagation \((10-12)\).

The use of RISs is particularly useful when the line-of-sight (LOS) link is blocked or not strong enough since it is possible to provide additional transmission paths by utilizing the reflecting elements of an RIS. Although it is possible to alleviate the negative effects of the channel using relays, the hardware cost, power consumption and latency increase considerably as the signal has been actively processed at each relay \((13)\). In addition, real-time controlled phase shifts of each RIS element allow the optimization of certain system performance metrics, such as transmit power, achievable rate, energy efficiency, and received signal-to-noise ratio (SNR) \((14-16)\). On the other hand, various secrecy enhancing schemes have been proposed by optimizing the transmit beamformer and RIS phase shifts jointly in \((17-19)\). In \((20)\), the communication through RISs has been investigated in terms of error performance and the symbol error probability (SEP) is derived for a scenario without a direct path between the transmitter and the receiver. However, to the best of our knowledge, a general analysis that elaborates the performance of single and multiple RIS-assisted systems in indoor and outdoor scenarios as well as under different conditions is missing in the literature.

In this paper, we investigate the performance of single and multiple RIS-assisted systems with or without a direct path between the transmitter and the receiver in indoor and outdoor propagation environments. First, in Section II, we show that the employment of an RIS is a dominant factor in the received signal power even in the presence of a direct communication link. By deriving mathematical expressions for the error performance under both large-scale and small-scale fading effects, we show that the RIS-assisted link acts as a superior LOS path by suppressing the other paths. Furthermore, we investigate the effects of RISs on the path loss exponent (PLE) for the log-distance path loss model \((21)\), and determine the achievable data rate of an RIS-assisted system under empirical path loss models for below and above 6 GHz operating frequencies. In Section III, we investigate the wireless communications in the presence of multiple RISs under two envisioned scenarios: simultaneous transmission over two RISs in indoor environment and the double-RIS reflected transmission in outdoor environment. We introduce a mathematical framework on the error performance for these two scenarios and demonstrate that the use of multiple RISs can be a remedy for the blockage problem in 6G and beyond systems. We also introduce novel RIS selection strategies for systems equipped with multiple RISs. Finally, we study the RIS imperfections, such as practical phase shifts and phase estimation errors, and demonstrate the robustness of RIS-assisted systems. Our comprehensive theoretical and numerical results in Section IV show that the transmission through RISs appear as an attractive candidate for indoor/outdoor communication systems at various operating frequencies by suppressing the destructive effects of wireless propagation.

II. COMMUNICATIONS THROUGH A SINGLE RIS

In this section, we consider the wireless communication system as shown in Fig. 1 with a single transmitter/receiver pair and a single RIS. Here, we provide first time a unified error performance analysis that is applicable to various path loss models in indoor and outdoor communication scenarios. Furthermore, the effect of an RIS on the overall path loss is examined in frequency bands below and above 6 GHz.

A. The General Model and Performance Analysis

As illustrated in Fig. 1, a generic RIS-assisted communication scenario is considered, where the transmission is carried out via an RIS in the presence of a direct path between the source (S) and the destination (D). We assume that the real-time controlled RIS is equipped with \(N\) reconfigurable reflect elements that can be adjusted according to channel phases. In Fig. 1 and the remainder of the paper we use the notations \(d_{SD}, d_{SR}\) and \(d_{RD}\) to represent the distances from source-to-destination (S-D), source-to-RIS (S-R), and RIS-to-destination (R-D), respectively. In this setup, \(d_v\) and \(d_H\) respectively denote the vertical and horizontal distance between S and RIS. The small-scale fading coefficients of S-D, S-R and R-D links are respectively shown by \(h_{SD}, h_{SR}^i\) and \(h_{RD}^i\) (for \(i = 1, \ldots , N\)), and follow \(CN(0,1)\) distribution under Rayleigh fading assumption.
Under the condition of slow and flat fading channels, the baseband received signal at D can be expressed as

$$r = \sqrt{p_t} \left[ P_L(d_{SD}) h^{SD} \right. $$

$$+ \left. \sqrt{P_R} \left( \sum_{i=1}^{N} \alpha_i e^{j \phi_i} h^{RD}_i \right) \right] x + w \quad (1)$$

where $p_t$ is the total transmitted power, $\phi_i$ stands for the controllable phase shift prompted by the $i$th RIS element under the assumption of unit-gain reflection coefficients, $x$ is the data symbol selected from $M$-ary phase shift keying/quadrature amplitude modulation (PSK/QAM) constellations with unit power, and $w \sim \mathcal{CN}(0, N_0)$ represents the additive white Gaussian noise (AWGN) sample, where $N_0$ is the noise variance. Additionally, $P_L(d_{SD})$ represents the path loss of the direct path and $P_R$ stands for the path loss of the RIS-assisted path. Under the specular reflection [22], $P_R$ will be proportional to total length of the RIS-assisted path ($P_R \propto d_{SR} + d_{RD}$).

Expressing the channel fading coefficients as $h^{SD} = \rho e^{-j \mu}$, $h^{RD}_i = \alpha_i e^{-j \theta_i}$ and $h^{RD}_i = \beta_i e^{-j \varphi_i}$, where $\rho, \alpha_i$ and $\beta_i$ stand for channel amplitudes while $\mu, \theta_i$ and $\varphi_i$ denote channel phases, the received signal can be rewritten as

$$r = \sqrt{p_t} \left[ P_L(d_{SD}) \rho e^{-j \mu} \right. $$

$$+ \left. \sqrt{P_R} \left( \sum_{i=1}^{N} \alpha_i \beta_i e^{j \Delta \Phi_i} \right) \right] x + w \quad (2)$$

where $\Delta \Phi_i = \phi_i - \theta_i - \varphi_i$ is the phase difference term for $i = 1, ..., N$. Therefore, the instantaneous SNR at D is given by

$$\gamma = \frac{\left| \sqrt{P_L(d_{SD}) \rho e^{-j \mu} + \sqrt{P_R} \left( \sum_{i=1}^{N} \alpha_i \beta_i e^{j \Delta \Phi_i} \right) \right|^2}{N_0} \frac{p_t}{}. \quad (3)$$

Here, the SNR can be maximized by aligning the phases of the reflected signals from the RIS to the phase of the direct path, that is, $\Delta \Phi_i = \mu$ for $i = 1, 2, ..., N$ considering trigonometric identities [20]. Under this intelligent reflection model with manipulated reflection phases, the maximized instantaneous SNR at D is obtained as

$$\gamma_{\text{max}} = \frac{\left( \sqrt{P_L(d_{SD}) \rho} + \sqrt{P_R} \left( \sum_{i=1}^{N} \alpha_i \beta_i \right) \right)^2}{N_0} \frac{p_t}{N_0} = \frac{B^2 p_t}{N_0} \quad (4)$$

where $\alpha_i$, $\beta_i$ and $\rho$ follow Rayleigh distribution. Using Lyapunov variant of the central limit theorem (CLT) [23], it can be shown that $B$, which is the sum of $N+1$ independent but non-identical random variables, follows the Gaussian distribution with the following mean and variance: $E[B] = \sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}}$ and $\text{VAR}[B] = P_R N \left( 1 - \frac{\pi^2}{16} \right) + P_L(d_{SD}) \left( 1 - \frac{\pi^2}{4} \right)$ for $N \gg 1$. Since $\gamma$ follows non-central chi-square distribution with one degree of freedom, its moment generating function (MGF) [24] is obtained as

$$M_{\gamma_{\text{max}}}(s) = \exp \left( \frac{e^{s(\sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}) p_t}}}{1 - \left( e^{s(\sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}) p_t}} \right)^2} \right) \quad (5)$$

Considering the generic SEP expression of [25] for $M$-PSK signaling, we obtain

$$P_e = \frac{1}{\pi} \int_0^{\frac{(M-1)\pi}{M}} M_{\gamma_{\text{max}}} \left( \frac{- \sin \left( \frac{\pi}{M} \right)}{\sin^2 \eta} \right) d\eta. \quad (6)$$

When binary PSK (BPSK) signaling is used, the average SEP can be simplified as

$$P_e = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \frac{1}{1 + \left( \sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}) p_t} \right)} \times \exp \left( \frac{- \frac{e^{s(\sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}) p_t}}}{1 - \left( e^{s(\sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}) p_t}} \right)^2} \right) d\eta. \quad (7)$$

Here, the average SEP in (7) can be further upper bounded by letting $\eta = \pi/2$:

$$P_e \leq \frac{1}{2} \exp \left( \frac{- \left( \sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}) p_t} \right)^2}{1 + \left( \sqrt{P_R N \frac{\pi}{4}} + \sqrt{P_L(d_{SD}) \frac{\pi}{4}) p_t} \right)} \right) \quad (8)$$

As shown in Section IV, the average SEP decreases with $N^2$ for the low $p_t/N_0$ region. Furthermore, we observe from (8) that for sufficiently large $N$, the effect of the direct path becomes less significant since the signals reflected from the RIS dominates the received SNR. Finally, we note that (8) can be used under different path loss models and generalizes the SEP derivations of [11].

### B. Performance under Empirical Path Loss Models for Outdoor Applications: Below and Above 6 GHz

In this subsection, we examine the effect of an RIS on the overall path loss between S and D in frequency bands below and above 6 GHz for the scenario of Fig.1. Empirical path loss models for outdoor communication scenarios have been taken into account to assess the potential of RIS-assisted systems. Then, the impact of an RIS on the PLE and achievable data rate is observed under the empirical transmission scenarios.
Fig. 2. The achievable data rate comparison for direct and RIS-assisted transmission under (a) 3GPP UMi path loss model with \( f_c = 2.4 \) GHz and (b) 5G UMi-Street Canyon path loss model with \( f_c = 60 \) GHz.

1) Urban Micro (UMi) Path Loss Models with a Single RIS: In order to assess the potential of an RIS over an RIS:

\[
\begin{align*}
L_{P_L}(d) &\equiv 22 \log_{10}(d) + 28 + 20 \log_{10}(f_c), \\
P_{NL}(d) &\equiv 36.7 \log_{10}(d) + 22.7 + 26 \log_{10}(f_c).
\end{align*}
\]

At frequency bands from 6 GHz to 100 GHz, the UMi-Street Canyon path loss models both LOS and NLOS cases are respectively given by \([27]\)

\[
\begin{align*}
P_{SL}(d) &\equiv 21 \log_{10}(d) + 32.4 + 20 \log_{10}(f_c), \\
P_{NL}(d) &\equiv 31.7 \log_{10}(d) + 32.4 + 20 \log_{10}(f_c).
\end{align*}
\]

The received signals at D for the RIS-assisted scenarios with \( N \) reflecting elements is obtained as in \([11]\) under specular reflection with \( P^R_l \propto P_L(d_{SR} + d_{RD}) \), that is, assuming LOS links for RISs. By considering the same analysis in (3) and assuming NLOS path-loss between S and D, the received SNR (\( \gamma \)) expressions for direct transmission and RIS-assisted models are respectively given by

\[
\begin{align*}
\gamma_{Direct} &= \frac{\sqrt{P_{NL}(d_{SD})h_{SD}^{h^{\text{SD}}}}^2}{N_0}, \\
\gamma_{RIS} &= \frac{\sqrt{P_{NL}(d_{SD})h_{SD}^{h^{\text{SD}}} + \sum_{i=1}^{N} h_{i}^{SR}e^{j\Phi_i}h_{i}^{RD}}^2}{N_0}.
\end{align*}
\]

The achievable data rate expressions for the direct transmission and a single RIS-assisted system are given by

\[
R_{Direct}(RIS) = \log_2 \left( 1 + \gamma_{Direct}(RIS) \right)
\]
corresponding distance length \((d)\) is calculated by

\[
p_r(d) = \left( \sum_i \sqrt{ \frac{p_t}{P_L(d_i)} } \right)^2
\]

where \(d_i\) is the total propagation distance of the \(i\)th path, \(p_t\) is transmitted power, \(P_L(d_i)\) is UMi path loss calculated for \(i\)th path. The average path loss is obtained as the ratio of \(p_t\) and \(p_r(d)\) using (13):

\[
P_L(d) = \left( \sum_i \sqrt{ \frac{1}{P_L(d_i)} } \right)^{-2}
\]

(14)

\(\text{PLE} (n)\) is then calculated by employing log-distance path loss model [21]:

\[
P_L(d) \text{ [dB]} = P_L(d_0) + 10n \log_{10} \left( \frac{d}{d_0} \right)
\]

(15)

where \(d_0\) is the reference distance.

The effect of an RIS on path loss is investigated for the scenario depicted in Fig. 1 for \(d_V = 10\) m and \(d_{SD}\) from 10 m to 200 m with \(N = 1.4\) and 16. The calculated total path loss is shown in Fig. 3 with LOS and NLOS UMi path loss models at 2.4 and 60 GHz. We observe that path loss of the RIS-assisted scenario with \(N = 1\) converges to the system modeled with UMi LOS when \(d_{SD}\) gets larger compared to \(d_V\), whereas it performs better when the distances are relatively small. Even in an RIS with a single reflector, a superior path loss is obtained over the LOS system for a closer \((d < 40\) m) distance because of the constructive effect from the NLOS path. Due to the fact that NLOS path loss increases with \(n = 3.67\) for 3GPP UMi model and \(n = 3.1\) for 5G UMi model by distance, the constructive effect of the NLOS path becomes negligible for larger \((d > 100\) m) distances. NLOS constructive effect becomes even smaller for an RIS with multiple reflectors, where the path loss reduces inversely by the square of the number of RIS elements, which is in agreement with [5] and [20].

PLEs are calculated for different reference \((d_0)\) and maximum \((d_{\text{max}})\) distances for different number of reflectors and listed in Table I. Our results indicate lower PLEs for smaller distances, where the constructive effect of the NLOS path is significant; whereas PLE approximates to the PLE of the LOS UMi channel model \((n_{\text{LOS}} = 2.2\) for 3GPP UMi model and \(n_{\text{LOS}} = 2.1\) for 5G UMi) for larger distances, where the effect of the NLOS path becomes negligible \((d > 50\) m). Our analysis shows that the LOS path loss and PLE are achievable even in the presence of obstacles blocking the LOS path between the transmitter and the receiver. In other words, an RIS transforms the wireless propagation environment into a controllable entity by providing LOS-level path loss even in NLOS propagation environments.

### III. WIRELESS COMMUNICATION THROUGH MULTIPLE RISs

In future wireless systems, the ubiquitous communications of many devices with various sizes are envisaged. However, it may not be reasonable to use multiple antennas or high-power consuming components in those devices to enable reliable and high-speed wireless transmission. Therefore, it becomes necessary to transfer this cost from wireless devices to the available RISs in propagation environments, where we can increase the number of RISs in a more flexible manner. It is worth noting that although many RIS-assisted systems have been investigated in recent times, the communication scenarios with multiple RISs have not been explored in a comprehensive manner so far. Within this perspective, the use of multiple RISs may have the potential to bring more promising advantages in terms of QoS, reliability and flexibility in the system design.

In this section, we conduct performance analysis of multiple RISs under two envisaged scenarios, namely, i) simultaneous transmission over two independent RISs in indoor environments and ii) double-RIS reflected transmission over a single link in outdoor environments. Under these multiple RIS-assisted transmission scenarios, we derive a generalized mathematical framework on the error performance which is valid for various path loss models.

#### A. Case Study I: Indoor Communications with Multiple RISs

In our first case study, we consider the indoor communications model of Fig. 3(a), where the direct path between S and D is blocked owing to the obstacles and the signal transmission is accomplished over two independent paths supported by two different RISs. This indoor scenario includes typical open and closed large office environments with a maximum S-D separation of 100 m.

In this setup, \(d_{SR_i}\) and \(d_{RD}\) respectively represent the distances of S-to-kth RIS and kth RIS-to-D for \(k = 1, 2\). Furthermore, small-scale fading coefficients of the S-to-kth RIS and kth RIS-to-D channels are denoted by \(h_i^{SR_k}\) and \(h_i^{RD}\) respectively.

Under the condition of slow and flat Rayleigh fading channels, the baseband signal at D is given by

\[
r = \sqrt{P_t} \sqrt{P_L^{R_i} \sum_{i=1}^N h_i^{SR_1} e^{j \phi_i(1)} h_i^{R_1 D}} + \sqrt{P_L^{R_2} \sum_{j=1}^N h_j^{SR_2} e^{j \phi_j(2)} h_j^{R_2 D}} x + w
\]

(16)

### Table I

| Channel Model | \(d_0\) | \(d_{\text{max}}\) | \(N\) | PLE \((n)\) |
|---------------|--------|----------------|------|-------------|
| 10 m 50 m     | 1      | 1.65           | 4    | 1.25        |
| 10 m 200 m    | 1      | 2.15           | 4    | 2.09        |
| 100 m 2000 m  | 1      | 2.21           | 4    | 2.20        |
| 50 m 200 m    | 1      | 1.58           | 4    | 1.20        |
| 50 m 200 m    | 150 m  | 2.08           | 4    | 1.98        |
| 5G UMi       | 10 m 50 m | 1.20         | 4    | 1.07        |
| 5G UMi       | 50 m 200 m | 2.00         | 4    | 1.98        |
Therefore, (16) can be expressed as

\[ h_{RIS} = \alpha_k e^{-j\theta_k}, \quad h_{RIS}^D = \beta_k e^{-j\phi_k}, \quad h_{RIS}^{S_2} = \alpha_j e^{-j\theta_j}, \quad \text{and} \quad h_{RIS}^{D_2} = \beta_j e^{-j\phi_j} \]

in terms of their amplitudes and phases. Therefore, (16) can be expressed as

\[
r = \sqrt{P_t} \left[ \sum_{i=1}^{N} \alpha_i^{(1)} \beta_i^{(1)} e^{j\Delta\phi_i} + \sum_{j=1}^{N} \alpha_j^{(2)} \beta_j^{(2)} e^{j\Delta\Phi_j} \right] x + w
\]

where \( \Delta\phi_i = \phi_i^{(1)} - \phi_i^{(2)} \) and \( \Delta\Phi_j = \phi_j^{(2)} - \phi_j^{(2)} \) are phase difference terms. We assume that each reflecting element behave as an specular reflector, thus received signal power is proportional to the total distances of the links: \( P_{R}^{k} \propto d_{S_{R}} + d_{R_{D}} \). For simplicity, distances of the first and second paths are accepted as equal: \( P_{R}^{k} = P_{R_1} = P_{R_2} \).

The instantaneous SNR at D can be expressed as

\[
\gamma = \frac{P_{R}^{k} \left[ \sum_{i=1}^{N} \alpha_i^{(1)} \beta_i^{(1)} e^{j\Delta\phi_i} + \sum_{j=1}^{N} \alpha_j^{(2)} \beta_j^{(2)} e^{j\Delta\Phi_j} \right]^2}{N_0} \text{ (18)}.
\]

In (18), the SNR is maximized with the phase alignment satisfying \( \Delta\Phi_1 = \Delta\Phi_2 \), for all \( i = 1, 2, \ldots, N \) and \( j = 1, 2, \ldots, N \); therefore, the maximized SNR can be obtained as in (4):

\[
\gamma_{\text{max}} = \frac{P_{R}^{k} \left[ \sum_{i=1}^{N} \alpha_i^{(1)} \beta_i^{(1)} + \sum_{j=1}^{N} \alpha_j^{(2)} \beta_j^{(2)} \right]^2}{N_0} = \frac{A^2 p_t}{N_0} \text{ (19)}.
\]

It should be noted that \( \alpha_i^{(k)} \) and \( \beta_i^{(k)} \) \((k = 1, 2)\) are independent and follow Rayleigh distribution. Using the CLT, it can be shown that \( A \), which is the sum of \( 2N \) independently identical distributed (iid) random variables, follows Gaussian distribution with \( 2N \sqrt{P_{R}^{k}} \pi \) mean and \( 2N P_{R}^{k} \left(1 - \frac{\pi^2}{16}\right) \) variance for \( N \gg 1 \). It is worth noting that \( \gamma \) will have the same distribution regardless of the type of small-scale fading due to the CLT. Using the MGF approach and following the same steps as in Section II with (8), upper bounded SEP for BPSK can be obtained as

\[
P_e \leq \frac{1}{2} \exp \left( -\frac{2N^{2} p_t^{2} \pi^2}{8N_0} \left(1 + \frac{N P_{R}^{k} \left(16 - \pi^2\right)p_t}{8N_0} \right)^{1/2} \right) \text{ (20)}.
\]

We can generalize this simultaneous transmission scenario with two RISs to the general case of \( K \) independent RISs. If each independent RIS consist of \( N \) reflector elements, the generalized SEP is expressed as follows for BPSK:

\[
P_e \leq \frac{1}{2} \exp \left( -\frac{K^2 N^2 p_t^{2} \pi^2}{16N_0} \left(1 + \frac{K N P_{R}^{k} \left(16 - \pi^2\right) p_t}{8N_0} \right)^{1/2} \right) \text{ (21)}.
\]

While the error performance is inversely proportional to \( (2N)^2 \) in the two RIS-assisted transmission, the simultaneous transmission in the presence of \( K \) different RISs provides

\[
P_e \propto \exp \left( -\frac{K^2 N^2 p_t^{2} \pi^2}{16N_0} \right) \text{ (22)}
\]

where \( K \) is the total number of reflectors in the system. In (22), it is observed that the error performance is inversely proportional to \( (K N)^2 \), which provides more flexibility in design when the RIS sizes are limited in indoor environments. In other terms, instead of employing a single large RIS, the same performance can be provided by multiple smaller RISs, which is better suited to indoor applications.

**B. Case Study II: Outdoor Communications with Multiple RISs**

The system model of the double-RIS reflected transmission in outdoor propagation environment is illustrated in Fig. 4(b). The outdoor scenario includes typical dense urban environments with a minimum S-D separation above 100 m. In this
scenario, since the direct path between S and D is blocked, the signal is transmitted via two RISs located near to S and D. This system can be used to alleviate the shortcomings of the use-cases that may arise particularly in outdoor environments in 6G and beyond communications systems, especially in mmWave and THz bands. For instance, if it is desired to convey data in an outdoor environment over a long-distance, at high frequencies, the signal can be attenuated or lost due to many obstacles in between. In this case, the nearest RISs, which are equipped with reflecting elements, are chosen by S and D for transmission, and communication is carried out by these RISs. We later show that using two RISs, the wireless environment is transformed into a virtual MIMO system.

The double-RIS reflected transmission scenario includes a single path from S to D. The channels between S-to-RIS 1 and RIS 2-to-D can be modeled as deterministic LOS channels while the channel between ith element of RIS 1-to-jth element of RIS 2 is modeled by Rayleigh fading, where $h_{ij}R_1, R_2 \sim CN(0, 1)$ for $i = 1, 2, \ldots, N$ and $j = 1, 2, \ldots, N$. Distance of the S-to-RIS 1, RIS 1-to-RIS 2 and RIS 2-to-D is respectively denoted by $d_{SR_1}$, $d_{R_1R_2}$ and $d_{R_2D}$. $d_{SR_1}$ and $d_{R_2D}$ are sufficiently small such that small-scale fading can be ignored. From the standpoint of channel amplitudes and phases, $h_{ij}R_1, R_2$ can be expressed as $h_{ij}R_1, R_2 = \beta_{ij}e^{-j\varphi_{ij}}$.

The baseband signal at D is obtained as follows:

$$r = \sqrt{P_t}p_R \left( \sum_{i=1}^{N} \sum_{j=1}^{N} e^{j\phi_i^{(1)}} h_{ij}R_1 e^{j\phi_j^{(2)}} \right) x + w$$

(23)

where $\phi_i^{(k)}$ is as defined in (16). Then, the instantaneous SNR at D can be easily calculated as

$$\gamma = \left( \frac{\sqrt{P_t}}{N_0} \sum_{i=1}^{N} \sum_{j=1}^{N} \beta_{ij} e^{j(\phi_i^{(1)} + \phi_j^{(2)} - \varphi_{ij})} \right)^2 \frac{p_t}{P_t R^{N^2}}$$

(24)

SNR at D is maximized with the phase elimination ($\phi_i^{(1)} + \phi_j^{(2)} = \varphi_{ij}$ for $i = 1, 2, \ldots, N$ and $j = 1, 2, \ldots, N$). Consequently, the maximized SNR is obtained as

$$\gamma_{\text{max}} = \left( \frac{\sqrt{P_t}}{N_0} \sum_{i=1}^{N} \sum_{j=1}^{N} \beta_{ij} \right)^2 \frac{p_t}{N_0}$$

(25)

where $A$ follows Gaussian distribution with $N^2\sqrt{P_t}(\sqrt{\pi/4})$ mean and $N^2P_t^2(1 - \pi/4)$ variance due to the CLT for $N \gg 1$. Then, following the same steps as in previous sections, the upper bounded SEP for BPSK can be obtained using (8) as

$$P_e \leq \frac{1}{2} \left( \frac{\exp \left( - \frac{N^2 P_t^2 R^2 \pi p_t}{4 N_0} \right)}{\left( 1 + \frac{N^2 P_t^2 (4 - \pi) p_t}{2 N_0} \right)^{1/2}} \right).$$

(26)

As can be seen from (26), in the low SNR region, the error performance of the double-RIS reflected transmission systems ($P_e$) can be approximated by

$$P_e \propto \exp \left( - \frac{N^4 P_t^2 R^2 \pi p_t}{4 N_0} \right).$$

(27)

The term $N^4$ in (27) brings a significant improvement in error performance due to the virtual $N \times N$ MIMO channel created between RIS 1 and RIS 2. It should be also noted that (26) can be used for various path loss models directly, while for other small-scale fading models, only slight modifications are required using the MGF approach due to the CLT.

C. RIS Selection Strategies

Although it is possible to improve the QoS using multiple RISs in future wireless networks, this will increase the overall complexity and render simultaneous phase adjustment a challenging task. By utilizing RIS selection over multiple RISs, a low-complexity and cost-effective transmission can be provided while many advantages of multiple RIS-assisted systems are preserved. For this purpose, we propose the novel concept of RIS selection for the communication scenarios of Fig. 5.
As shown in Fig. 5(a), $K$ RISs are placed in an indoor propagation environment, where each RIS has $N$ reflecting elements. Here, $d_{SR_k}$ and $d_{RD}$ respectively represent the distances of $S$-to-$k$th RIS and $k$th RIS-to-$D$ for $k = 1, 2, \ldots, K$. Furthermore, small-scale fading coefficients of the $S$-to-$k$th RIS and $k$th RIS-to-$D$ channels are denoted as in (17) by $h_{i}^{SR_k} = \alpha_i^{(k)} e^{-j\theta_i^{(k)}}$ and $h_{i}^{RD} = \beta_i^{(k)} e^{-j\varphi_i^{(k)}}$ for $i = 1, 2, \ldots, N$ and $k = 1, \ldots, K$, respectively. The received SNR of the signal transmitted through only the $k$th RIS ($\gamma_k$) is calculated as

$$\gamma_k = \sqrt{P_L \left( \sum_{i=1}^{N} \alpha_i^{(k)} \beta_{i}^{(k)} e^{j\Delta\Phi_i} \right)^2} p_t,$$  \hspace{2cm} (28)

where $\Delta\Phi_i = \phi_i^{(k)} - \phi_i^{(k)} - \varphi_i^{(k)}$ is the phase difference for the $k$th path. Considering the maximized SNR values as in (18), an RIS selection is conducted over $K$ RISs to choose the RIS, which has highest SNR, for transmission as

$$\gamma'_L = \max(\gamma_{\text{max}, 1}, \gamma_{\text{max}, 2}, \ldots, \gamma_{\text{max}, K})$$  \hspace{2cm} (29)

where $\gamma'_L$ is the maximized received SNR for RIS selection based system in indoor environment and $\gamma_{\text{max}, k}$ is the maximized SNR of the $k$th path with proper phase adjustment. Here, we focus on the selection of a single RIS while a generalization might be straightforward. The generalized system model of the RIS selection based outdoor communication system is illustrated in Fig. 5(b). In this model, we assume $K_S$ and $K_D$ RISs, respectively, within the proximity of the transmitter and receiver. Here, the distances of $S$-to-$k$ RIS, $k$-to-$RIS$ $l$ and RIS $l$-to-$D$ are respectively denoted by $d_{SR_k}$, $d_{RL_l}$ and $d_{RD}$ for $k = 1, \ldots, K_S$ and $l = 1, \ldots, K_D$ as in (25). In this setup, $K_S \times K_D$ possible transmission paths are available for communications. By selecting the path that has the highest SNR, the overall cost can be reduced using a single RIS for each terminal. As in (25), the maximized SNR is obtained for each path as $\gamma'_{k,l}$.

Therefore, RISs selection for outdoor environment is conducted over $K_S \times K_D$ paths as

$$\gamma'_O = \max_{k,l} (\gamma'_{k,l})$$  \hspace{2cm} (30)

where $\gamma'_O$ is the maximized received SNR for RIS selection based systems in outdoor environment.

**IV. NUMERICAL RESULTS**

In this section, we used simulation models and studied the effects of various RIS applications taking into account path losses and channel fading.

In Fig. 6, we demonstrate the transformative effect of an RIS on the propagation medium. In this comparison, BER analysis of two different RIS-assisted systems with and without direct path is performed, where distances of the paths are assumed as the same as in Fig. 5. It is clearly seen that in the presence of a single RIS, the maximized SNR of (4) leads to a significant improvement in the error performance. In addition, when the total number of reflectors ($N$) increases, the difference in BER for these two systems diminishes. This indicates that the increase in $N$ suppresses the effect of the direct path. In other words, even a single RIS with sufficiently large number of reflecting elements RIS can provide an effective solution even if the LOS path is blocked.

In Fig. 7, we further study the analytical error performance of systems with single and two RISs is examined by making comparisons with a reference system utilizing only direct transmission under AWGN and Rayleigh fading channels. The distances are taken as $d_{SD} = 100$ m and $d_{SR} = d_{RD} = 50$ m in the single RIS-assisted system, $d_{SR_1} = d_{SR_2} = d_{R_1D} = d_{R_2D} = 50$ m in two simultaneous RIS-assisted setup, and...
When two RISs are available a flexible transmission in outdoor propagation environment = 4 single RIS-assisted systems when the direct path is blocked for based transmission provides a better error performance than

selecting the best RIS from (29). Instead of a single RIS-

RIS-assisted scenarios under BPSK.

Fig. 8. Comparison of exact and upper bounded average SEP for different RIS-assisted scenarios under BPSK.

d_{SR_1} = d_{RD} = 5 \text{ m} \text{ and } d_{R_1R_2} = 90 \text{ m} \text{ in double-RIS reflected transmission scenario. The double-RIS system provides a better error performance than the other setups, since } P_e \text{ is inversely proportional to the fourth power of the number of reflectors } (P_e \propto N^{-4}), \text{ while we have } P_e \propto (2N)^{-2} \text{ in simultaneous transmission scenario over two RISs and } P_e \propto N^{-2} \text{ in a single RIS-assisted system for low } p_t/N_0 \text{ values. It should be noted that the double-RIS reflected scheme is the most challenging scheme in its design while offering a far better error performance than the others. In a nutshell, significant performance gains are obtained with respect to the AWGN channel using single or multiple RISs. In other words, it is possible to convert a Rayleigh fading channel to a super communication channel that acts as a non-fading one by intelligent reflections. It can be also stated that the direct path is transformed into an } N \times N \text{ virtual MIMO channel with the help of RISs, located close to the transmitter and receiver and each of which contains } N \text{ reflectors in the double-RIS reflected transmission setup.}

In Fig. [9] we implement RIS selection strategies for the indoor and outdoor scenarios depicted in Fig. [5]. First, we consider the indoor transmission with three available RISs by selecting the best RIS from (29). Instead of a single RIS-assisted transmission for } d_{SR} = 25 \text{ m} \text{ and } d_{RD} = 25 \text{ m}, \text{ the best path can be selected among the three possible paths with the following distances: } d_{SR_1} = 25 \text{ m}, \text{ } d_{R_1D} = 25 \text{ m}, \text{ } d_{SR_2} = 10 \text{ m}, \text{ } d_{R_2D} = 45 \text{ m}, \text{ } d_{SR_3} = 20 \text{ m} \text{ and } d_{R_3D} = 30 \text{ m}. \text{ As shown in Fig. [9]a, the RIS selection based transmission provides a better error performance than single RIS-assisted systems when the direct path is blocked for } N = 4 \text{ and } 16. \text{ Furthermore, our RIS selection strategy enables a flexible transmission in outdoor propagation environment when the LOS path is blocked. When two RISs are available for both the transmitter and the receiver } (K_S = 2 \text{ and } K_D = 2), \text{ transmission can be carried out over four possible links } (K_S \times K_D = 4). \text{ In this case, the RIS pairs that provide the highest SNR value are selected by using (30) and transmission is provided via a single path. Using the system models in (23), the distances for the four possible paths are assumed as follows: } d_{SR_1} = 10 \text{ m}, \text{ } d_{SR_2} = 5 \text{ m}, \text{ } d_{R_1K_1} = 400 \text{ m}, \text{ } d_{R_1K_2} = 410 \text{ m}, \text{ } d_{R_2K_1} = 380 \text{ m}, \text{ } d_{R_2K_2} = 395 \text{ m}, \text{ } d_{R_1D} = 20 \text{ m}, \text{ and } d_{R_2D} = 10 \text{ m}. \text{ As a reference system, a single-RIS assisted system with } d_{SR} = d_{RD} = 205 \text{ m}, \text{ and double-RIS reflected system as illustrated in [6]b) are considered. As shown in Fig. [6]b), an improved error performance is obtained under RISs selection with respect to single and multiple-RIS assisted systems while the overall system complexity and phase adjustment costs are decreased. Here, the most dominant factor in the RIS selection is the path loss rather than small-scale fading. Under same fading conditions, the best way to ensure a reliable transmission is choosing the shortest path for transmission.}

In order to give useful insights into the practical implementation of RISs, we consider the practical reconfigurable metasurface designed in [28]. This metasurface is capable of adjusting reflection phases from } -150^\circ \text{ to } 140^\circ \text{ for } |\Gamma| = -1 \text{ dB. In Fig. [10]a), under a single RIS-assisted scenario, an error performance comparison is given for this RIS and the ideal one, which is capable of reflecting the signal within the } [-180^\circ, 180^\circ] \text{ phase range for } |\Gamma| = 0 \text{ dB. We observe that...}
even in the case of non-ideal RIS-assisted transmission, there will be no significant changes in error performance. Another problem that may be encountered in real-world applications is the imperfect estimation of the channel phase. This imperfect phase estimation results in the worsening of the received SNR by preventing the constructive combination of received signals. It is assumed that the phase estimation errors follow a zero-mean von Mises distribution with the concentration parameter $\kappa$, where $\kappa$ is a measure of the estimation accuracy. In Fig. 10(b), a single RIS-assisted system under imperfect and ideal phase estimation is investigated in terms of the error performance for varying $\kappa$ and $N$ values. Although small $\kappa$ values lead to a degradation in error performance, this effect is less noticeable for large $N$ and $\kappa$.

V. Conclusions

In this paper, we provide unique RIS-oriented solutions for potential communication scenarios that may emerge in 6G and beyond wireless networks. In this context, we investigate a number of RIS-assisted systems with single or multiple RISs, in terms of error performance, achievable data rate and path loss characteristics in indoor and outdoor environments. To our knowledge, for the first time, our paper explores the effect of RISs on the overall path loss under empirical path loss models and the cases of multiple RISs, as well as RIS selection strategies, using a unified error performance framework. Our simulation results indicate that RIS-assisted link acts as an LOS path by suppressing the effect of NLOS paths when the number of reflecting elements increases and double-RIS reflected transmission further improves the performance. In other words, one can eliminate the need for LOS transmission utilizing emerging RISs and multiple RIS-assisted systems may be a potential remedy for future dense wireless networks. We also note that the analysis under MIMO setups and THz bands appear as interesting future research directions. To sum up, the communication with real-time controlled RISs can be considered as an exciting technology to meet the demands of 6G and beyond wireless networks.

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