

Reflection of Wideband Signals With Reconfigurable Intelligent Surfaces: A Tutorial

Pedro H. C. de Souza and Luciano Mendes

Abstract—Recently, the reconfigurable intelligent surface (RIS) technology has ushered in the prospect of control over the wireless propagation environment. By establishing alternative propagation paths for the transmitted signals, and by reflecting them in a controllable manner, the RIS is able to improve the signal reception.

In this tutorial, we present the fundamental modeling aspects of RIS-aided communication systems, accompanied by practical examples and intuitive explanations. The main concepts underlying wideband signal reflection with RISs systems are also explored. We, moreover, evaluate the impact of bandwidth restrictions on the wideband signal reflected by the RIS. If not carefully considered, this can become an impediment for the adoption of the RIS in the next generation of communications systems. Therefore, in this tutorial we show a RIS configuration method that provides frequency selective signal reflection for wideband signals.

Index Terms—RIS, frequency selective, wideband, OFDM, interference management.

I. INTRODUCTION

WHILE the fifth generation of mobile networks (5G) is being deployed, and well underway globally [1], [2], the shaping of the next generation of mobile communications also begins to acquire more definite traces. Study cases and requirements for the sixth generation of mobile networks (6G) are now being discussed by the academia and industry alike and several new technologies are under scrutiny. More specifically, with the emergence of metamaterials (e.g. graphene) surfaces in the context of signal transmission in communication systems, new technologies for controlling the propagation of electromagnetic waves were introduced. One such technology is typically known as the reconfigurable intelligent surface (RIS), which can be described as a planar array of tunable resonant elements designed to dynamically and precisely manipulate the reflection of incident electromagnetic waves. The RIS technology is currently on track to be considered as a study-item by important standardization bodies, such as the 3rd generation partnership project (3GPP) [3].

In principle, an RIS, or multiple RISs, can be deployed in wireless communication systems with the aim of improving

the signal transmission through the communications channel. This concept disrupts the status quo present in wireless communication systems for decades: that the signal transmitted by a base station, should propagate through the communications channel unaided until its reception by a terminal. In other words, the usual approach involves processing the signal before transmission and during reception in order to improve the signal quality delivered by communication systems. This motivated the development of robust signal processing algorithms and also several antennas technologies over the years [4].

However, the concept of improving the signal propagation through the communication channels is not new. Relays, for instance, are devices equipped with high-cost and high-power radio frequency (RF) components that can aid the signal transmission between the transmitter and receiver [5]. Thanks to RF amplifiers and baseband signal processing, such relays are capable of actively generating the forwarded signals and even detecting and decoding the received signals. Despite these advantages, there are limitations associated with relays that need to be taken into account as, for example, introduction of noise, increased latency and low energy efficiency. On the other hand, RISs are constituted by passive components, such as varactors or varistors with adjustable impedance values, and practically no signal processing is required for signal reflection. Consequently, RISs are envisioned to be a very low cost technology and also very efficient in both energy consumption and in terms of computational burden. Moreover, they can be also physically placed in building facades, ceilings, and advertising panels, allowing for a convenient deployment in both indoor or outdoor wireless systems' environments.

In summary, the RIS can be described as a planar surface composed of multiple elements or reflectors, each of them built with tunable resonant elements. These passive reflectors provide adjustable impedance values [6], which can then be used to rotate (phase-shift) the electromagnetic waves impinging on its surface. As long as the RIS is properly positioned between a transmitter and receiver, it is able to reflect signals through additional propagation paths [7]. Consequently, the RIS can create favorable propagation conditions for the signal transmission, specially if the direct channel between the transmitter and receiver is in non-line-of-sight (NLOS).

Typically, the RIS reflectors are configured in such a way as to promote the most constructive combination of signals at the receiver [8]. This ensures that the channel capacity at the receiver is increased, since the signal reflected by the RIS is now being combined coherently with signal received via the direct channel. However, the interference management of the signals reflected by the RIS is one aspect that is overlooked

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Submission: 2025-12-02, First decision: 2026-02-18, Acceptance: 2026-07-20, Publication: 2026-07-28.

by the aforementioned approach [3]. On the other hand, works such as [9], for example, propose an RIS hardware design that provides a greater degree of control over the reflected signals, allowing a frequency selective RIS configuration to be attained. Furthermore, in [10], an evolution of the RIS technology, known as beyond diagonal reconfigurable intelligent surface (BD-RIS), is used for multi-band operation in a multiple-input multiple-output (MIMO) communications system. In [11], an algorithm is proposed for configuring the precoding vectors and the RISs' phase configurations in a distributed manner, taking into account the frequency selectivity behavior of the RIS.

In this tutorial we provide the foundational definitions and modeling for RIS-aided communication systems, while also delving into more advanced concepts such as wideband and frequency selective signal reflection with RIS systems. From an elementary point-to-point communications system, this work builds up from each new concept introduced, using practical examples and a variety of diagrams and figures to facilitate the understanding of communication systems assisted by RISs. With this knowledge, we also demonstrate that a time-variant RIS configuration is capable of selecting sub-bands for wideband signal reflection. More specifically, we show that a frequency selective signal reflection can be achieved within a bandwidth of interest, and without special requirements for the RIS hardware. To the best of authors' knowledge, such a frequency selective RIS configuration method for wideband signal reflection is, as of yet, unexplored for interference management in RIS-aided communication systems.

This work is organized as follows: Section II introduces the signal propagation model for the RIS-aided communications system; Section III details the RIS configuration method used for wideband and frequency selective signal reflection; in Section IV the main findings are discussed with the assistance of numerical results generated via computational simulations and, finally, Section V concludes the paper.

A. Notation

Throughout this paper, italicized letters (e.g. x or X) represent scalars, boldfaced lowercase letters (e.g. $\mathbf{x}$) represent vectors, and boldfaced uppercase letters (e.g. $\mathbf{X}$) denote matrices. The n th entry of the vector $\mathbf{x}$ is represented by $x[n]$. The entry on the i th row and j th column of the matrix $\mathbf{X}$ is denoted by $X_{i,j}$. The number of elements contained in a set $\mathcal{X}$ is denoted by $|\mathcal{X}|$. The modulo operation is represented by $\text{mod}(x)$, $\lfloor x \rfloor$ denotes the rounding of scalars to the nearest integer, and $\lceil x \rceil$ rounds to the closest integer smaller than $x \in \mathbb{R}$. The sets of vectors (matrices) of dimension X ($X \times Y$) with real and complex entries are respectively represented by $\mathbb{R}^X$ ($\mathbb{R}^{X \times Y}$) and $\mathbb{C}^X$ ($\mathbb{C}^{X \times Y}$). The transposition and conjugate transposition operations of a vector or matrix are represented as $(\cdot)^\top$ and $(\cdot)^\dagger$, respectively. The X -dimensional vector of all ones is denoted by $\mathbf{1}_X$. The ℓ_p -norm, $p \geq 1$, of the vector $\mathbf{x} \in \mathbb{C}^N$ is given by $\|\mathbf{x}\|_p$ and also $\langle \mathbf{x} \rangle = N^{-1} \sum_n x[n]$ for $\mathbf{x} \in \mathbb{R}^N$. Finally, $x \sim \mathcal{U}[a, b]$ is a random value drawn from the uniform distribution over the interval $[a, b]$ and $x \sim \mathcal{N}(0, 1)$ is a random value drawn from the normal distribution.

II. RIS-AIDED COMMUNICATION SYSTEMS FUNDAMENTALS

A fundamental wireless communication link is established when a transmitter sends electromagnetic waves to several receivers spread over a region. For ease of presentation, in this work we take as reference a point-to-point communications system [12] between a transmitter and receiver, henceforth denoted as access point (AP) and user equipment (UE), respectively. Consequently, we focus on a single communication link or channel established by the pair AP/UE. The channel between the AP and UE is typically assumed to be in line-of-sight (LOS), meaning essentially that there are no obstacles for the signal propagation through the channel. However, since wireless communications are expected to work under diverse regions and scenarios, usually the point-to-point communications channel is very weak or completely unavailable, also known as a NLOS channel. A classical scenario where this might occur is when the signal propagates through urban environments, where buildings, houses and other structures hinder the propagation of electromagnetic waves [13]. Therefore, recently a sizable research effort was directed at devising solutions for improving the signal propagation. One solution denoted as RIS has shown promising results and is going to be discussed in more details in what follows.

A. RIS Systems

Let a point-to-point communications system be assisted by an RIS with N reflectors or elements. The precise RIS location is conveniently determined so that the AP and UE have LOS propagation paths with the RIS, as illustrated in Fig. 1. Therefore, the RIS can then enhance the signal reception in a scenario where the direct signal propagation paths between the AP and UE are in NLOS (see buildings in Fig. 1). In other words, by reflecting signals transmitted from the AP back to the UE, the RIS creates new propagation paths with more favorable conditions for signal transmission. For this, the n -th RIS reflector is configured with a continuous phase-shift, $\theta_n = [-\pi, \pi] \quad \forall n \in \{1, 2, \dots, N\}$. We assume for simplicity that the RIS can provide lossless signal reflection and that mutual coupling is negligible across all reflectors.

B. RIS-aided Communications

To evaluate the improvement brought by the RIS for point-to-point communications, first it is convenient to define a channel model. For this, consider the diagram depicted in Fig. 2. Observe in Fig. 2 that the direct channel is represented by L_d (NLOS) propagation paths between the AP and UE, whereas the composite channel, composed of the cascade of the channels between the AP and RIS and RIS to the UE, have L_a and L_b propagation paths, respectively. The relative azimuth angle of arrival, ϕ_a , and the elevation angle, φ_a , are also being illustrated in Fig. 2¹. These angles are used to compute the array response of the RIS [14], which accounts for different phase rotations depending upon the location of a given reflector across the RIS plane. Moreover, note that each reflector has sides of size d_H and d_V meters.

¹The redundant L_b , ϕ_b and φ_b symbols are omitted to improve visibility.

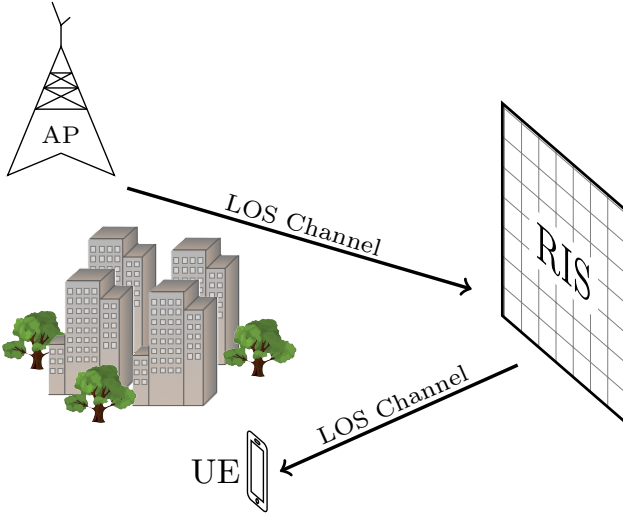


Fig. 1. RIS-aided communications system. The channels between AP and RIS (RIS and UE) are in LOS, while the signal propagation between AP and UE is impaired by physical structures, thus becoming a NLOS channel. Note that the represented dimensions are not to scale.

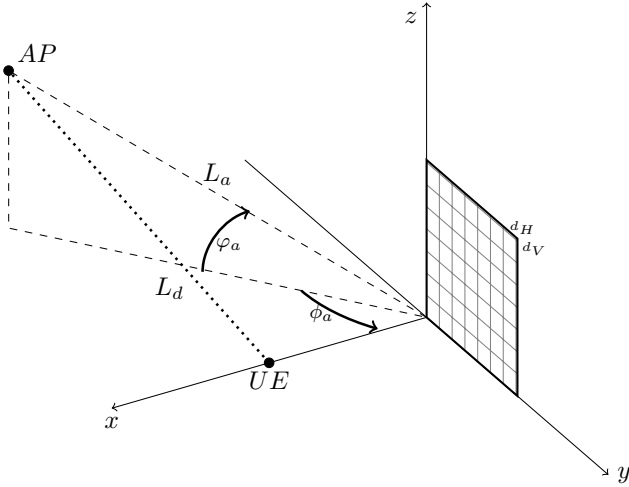


Fig. 2. Diagram of the propagation model on a (x, y, z) coordinate system.

More specifically, the RIS array response, considering the L_a channel paths, can be given by

$$\mathcal{S}(\Phi_a) = e^{j\bar{\Phi}_a^{(1)\top} \psi_n}, \quad (1)$$

where $l \in \{1, 2, \dots, L_a\}$ and $\Phi_a = [\bar{\Phi}_a^{(1)}, \bar{\Phi}_a^{(2)}, \dots, \bar{\Phi}_a^{(L_a)}]$ is the set of all azimuth and elevation angles of arrival, while ψ_n represents the n -th column of $\Psi = [\mathbf{0}, \psi_H, \psi_V]^\top$, and furthermore we have

$$\bar{\Phi}_a^{(l)} = \frac{-2\pi}{\lambda} \left[\cos(\phi_a^{(l)}) \cos(\varphi_a^{(l)}), \sin(\phi_a^{(l)}) \cos(\varphi_a^{(l)}), \sin(\varphi_a^{(l)}) \right]^\top, \quad (2)$$

in which $\lambda = 3 \times 10^8 f_c^{-1}$, f_c being the central frequency of the signal carrier. Additionally, consider the following

$$\psi_H = d_H [\text{mod}(0/N_{\text{row}}), \text{mod}(1/N_{\text{row}}), \dots, \text{mod}(N-1/N_{\text{row}})]^\top; \quad (3a)$$

$$\psi_V = d_V [[0/N_{\text{col}}], [1/N_{\text{col}}], \dots, [N-1/N_{\text{col}}]]^\top, \quad (3b)$$

for which $N = N_{\text{row}} N_{\text{col}}$, where N_{row} and N_{col} are the number of rows and columns of the RIS, respectively. Observe that $\Psi \in \mathbb{C}^{3 \times N}$ can be seen as the area spanned by the RIS planar surface across the $y-z$ plane, as illustrated in Fig. 2. In other words, the RIS reflector farthest from the plane origin imposes more rotation to the signal, with the other reflectors presenting gradually less rotation as they draw closer to the plane origin. These rotation values are then weighted by the impinging signal angles (angles of arrival), with all these operations being computed accordingly in (1). The same formulation of (1) is employed for the L_b channel paths, the main difference being that the angles of departure are now considered instead of the angles of arrival. Note also that the RIS orientation could be modified without loss of generality.

Consequently, the direct channel can be given as

$$h_d[m] = \sum_{l=1}^{L_d} \sqrt{\delta^{(l)}} \text{sinc} \left(m + B \left(\eta - \tau_d^{(l)} \right) \right) \times e^{-j2\pi(f_c \tau_d^{(l)})}, \quad (4)$$

wherein $\delta^{(l)} \forall l \in L_d$ are the propagation losses for the L_d direct channel paths, $\tau_d^{(l)} \geq 0$ represent the propagation delays, B is the channel bandwidth, and $\eta = \min(\tau_d^{(l)})$, $\forall l \in L_d$, in order to assure causality. Likewise, the discrete-time impulse response of the composite channel can be written as

$$h_c[m] = \sum_{n=0}^{N-1} \sum_{l=1}^{L_a} \sum_{\ell=1}^{L_b} \sqrt{\alpha^{(l)} \beta^{(\ell)}} \mathcal{S}(\Phi_a) \mathcal{S}(\Phi_b) \times s(m) e^{-j2\pi f_c (\tau_a^{(l)} + \tau_b^{(\ell)} + \tau_{\theta_n^t})}; \quad (5)$$

in which $m \in \{1, 2, \dots, M\}$ and also,

$$s(m) = \text{sinc} \left(m + B \left(\eta - \tau_a^{(l)} - \tau_b^{(\ell)} + \tau_{\theta_n^t} \right) \right), \quad (6)$$

where $\alpha^{(l)}$ and $\beta^{(\ell)}$ are, respectively, the propagation losses for the L_a and L_b channels, $\mathcal{S}(\Phi_a)$ and $\mathcal{S}(\Phi_b)$ are given by (1), $\tau_a^{(l)} \geq 0$ and $\tau_b^{(\ell)} \geq 0$ are the propagation delays, whereas $\tau_{\theta_n^t} = \theta_n^t / 2\pi f_c$ is the time-varying phase-shift imposed by the n -th RIS element at the discrete-time index $t \in \{1, 2, \dots, T\}$. In the next section, we show in detail how configuring the RIS can improve the signal reception for point-to-point communication systems. In what follows, the frequency selective RIS configuration is also introduced, demonstrating that the RIS can be configured differently depending on the use case.

III. RIS CONFIGURATION

It is well known that for narrowband channels the optimum RIS configuration must satisfy $e^{-j2\pi f_c \tau_d} = e^{-j2\pi f_c (\tau_a + \tau_b + \tau_{\theta_n^t})}$ [7]. In other words, the signal propagated through the composite channel must be in-phase with the signal transmitted via the direct channel. Recall that the point-to-point communications channel is represented by the direct channel for the RIS-aided communications scenario, which means that the signal transmitted through this channel can not be controlled by the RIS. Hence this channel is often referred to as the uncontrollable channel. However, the phase of N reflected signals created by the RIS (composite channel) can be controlled in such a manner

as to combine signals coherently at the UE. Consequently, this establishes a constructive combination of signals at the receiver or UE, thereby increasing the signal receiving power and reception quality in general.

A. RIS Configuration for Wideband Signals

Nowadays, often wideband signals are transmitted by the AP using the orthogonal frequency division multiplex (OFDM) system, for example. OFDM systems have become the main multiple access technique for communication systems in recent years, since they provide multiple access communications while also being robust to small scale channel fading. The transmitted OFDM signal is multiplexed into K sub channels or subcarriers within the total bandwidth B . Therefore, if we assume a constant RIS configuration for the entire bandwidth, then the RIS can not simultaneously maximize the channel capacity at the UE for all K subcarriers [7]. This happens because each subcarrier presents different channel properties and the RIS is unable to configure a different phase-shift for each subcarrier. Nevertheless, the RIS can still be configured with the aim of maximizing the aggregate channel capacity at the UE, also known as the achievable sum-rate [7], [14]:

$$R = \frac{B}{\xi} \sum_{i=0}^{K-1} \log_2 \left(1 + \frac{p_i \|\mathbf{f}_i^\dagger \mathbf{h}_d + \mathbf{f}_i^\dagger \mathbf{h}_c\|_2^2}{BN_0} \right) \text{ bit/s}, \quad (7)$$

for which $\mathbf{f}_i$ represents the i -th row of the discrete Fourier transform (DFT) matrix $F_{i,j} = e^{-j2\pi ij/K}$, while $\xi = K + M - 1$ to take into account the cyclic prefix loss inherent to OFDM signals, with $\mathbf{p} \in \mathbb{R}^K$ being the power vector, in which p_i is the power allocated to the i -th subcarrier, such that $P = \langle \mathbf{p} \rangle$; P being the total transmission power, and, finally, N_0 is the additive white Gaussian noise (AWGN) power density. Moreover, see in (7) that $\mathbf{h}_d = [h_d[1], h_d[2], \dots, h_d[M], \mathbf{0}_{K-M}]^\top$ and $\mathbf{h}_c \in \mathbb{C}^K$ denote K samples (with $K - M$ padding samples) of the direct channel (4) and composite channel (5) discrete-time impulse responses, respectively.

Observe in (7) that maximizing the achievable sum-rate entails in determining the RIS configuration that maximizes the signal magnitude at the UE. Fig. 3 shows the signal magnitude (color coded) for the half-space area covered by the RIS, where the AP is located approximately 200 m from the RIS, as indicated by the coordinates in Fig. 3 (a), and the UE is 20 m from the RIS. In Fig. 3 (a), instead of an RIS, we begin by evaluating the signal magnitude from a reflection created by a metal sheet. Although this surface also reflects the impinging electromagnetic waves, it does so without configuring the properties of the signal. Therefore, notice in Fig. 3 (a) how the signal magnitude of the reflected signal spreads alongside the surface length. On the other hand, for the RIS in Fig. 3 (b), a concentrated beam of signal magnitude is formed in the direction of the UE, showing that the RIS configuration is able to efficiently increase the signal magnitude at the receiver in comparison to a static metal sheet reflector. However, Fig. 3 (c) illustrates the scenario for which the channel between AP and RIS is in NLOS. Notice how the concentrated signal beam vanishes, showing how important is

the correct placement of the RIS in order to increase LOS availability for signal reflection. The scenario of Fig. 3 (c) also shows that the signal magnitude at the UE is low if only the direct channel or a point-to-point communications system is considered.

B. Frequency Selective RIS Configuration

Typically, it is assumed that the signal transmitted on all K OFDM subcarriers are to be reflected by the RIS to an intended UE. However, this may impose restrictions on the RIS implementation if only a subset of subcarriers are allocated to this UE. In other words, the unrestricted reflection of wideband signals by the RIS may become undesirable, since the reflected signal to unintended UEs can occur if the remaining subcarriers are allocated to them. With that in mind and by assuming a constant channel during the RIS configuration, then we can write the achievable rate as follows

$$R = \frac{B}{\xi} \sum_{i \in \mathcal{I}} \log_2 \left(1 + \frac{p_i \|\mathbf{f}_i^\dagger \mathbf{h}_d\|_2^2 + \mathbf{f}_i^\dagger \boldsymbol{\vartheta} \|\mathbf{f}_i\|_2^2}{BN_0} \right) \text{ bit/s}. \quad (8)$$

By letting $\mathcal{I}$ denote the set of subcarriers indexes (bins) selected for the selective signal reflection, then we also have $\xi = |\mathcal{I}| + M - 1$, with the remaining variables being defined in the same fashion as in (7). The composite channel samples can be otherwise expressed as

$$\boldsymbol{\vartheta} = (\mathbf{V} \boldsymbol{\Omega}_\theta)^\top \mathbf{1}_N, \quad (9)$$

wherein $\mathbf{V} \in \mathbb{C}^{N \times K}$ contains the discrete-time impulse response for all N composite channels and $\boldsymbol{\Omega}_\theta \in \mathbb{C}^{K \times T}$ is the RIS configuration. Observe that $\boldsymbol{\Omega}_\theta$ encodes a time-varying RIS configuration, for which its columns can be seen as the phase-shift applied to all samples of the composite channel's impulse response at the discrete-time index t . In other words, the columns of $\boldsymbol{\Omega}_\theta$ represent a given RIS configuration at time t .

1) *Defining the Configuration Matrix $\boldsymbol{\Omega}_\theta$* : It is well known that in order to properly estimate channel coefficients in RIS-aided communication systems, one may resort to orthogonal matrices [15]. Similarly, in order to achieve a frequency selective signal reflection, in this work we define the RIS configuration matrix according to

$$\boldsymbol{\Omega}_\theta = \frac{\mathbf{1}_K \mathbf{w}^\top}{|\mathcal{I}|}; \text{ where } \mathbf{w} = \mathbf{F} \mathbf{b} \in \mathbb{C}^K \quad (10)$$

for which $\mathbf{F} \in \mathbb{C}^{K \times K}$ is the DFT matrix and $b[i] = 1$ for $i \in \mathcal{I}$ and zero otherwise. It is easy to show that this RIS configuration leads to a frequency selective reflection. Firstly, note that the right hand side (composite channel term) of the euclidean norm of (8) can be rewritten as

$$\mathbf{f}_i^\dagger \boldsymbol{\vartheta} = |\mathcal{I}|^{-1} \mathbf{f}_i^\dagger \mathbf{w} \sum_{\forall n} \mathbf{1}_K^\top \mathbf{v}_n, \quad (11a)$$

$$= |\mathcal{I}|^{-1} \mathbf{f}_i^\dagger \mathbf{F} \mathbf{b} \sum_{\forall n} \sum_{\forall k} v_n[k], \quad (11b)$$

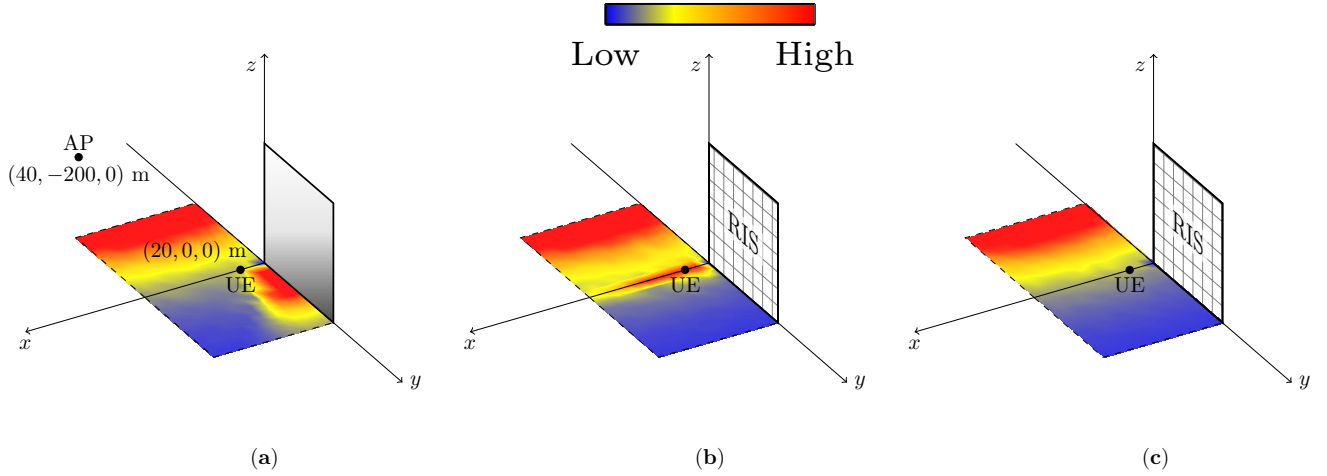


Fig. 3. The signal magnitude is illustrated for the half-space area covered by the RIS: considering a metal sheet surface (a); an RIS (b) and an RIS without the LOS channel with the AP. High signal magnitudes are represented by the red color, whereas low signal magnitudes are represented by the blue color (this figure is better visualized in colors).

where $\mathbf{v}_n$ represents the n -th row of $\mathbf{V}$. Consequently, by knowing that $\mathbf{F}^\dagger \mathbf{F} \mathbf{b} = \mathbf{b}$, then it follows that

$$\mathbf{f}_i^\dagger \boldsymbol{\vartheta} = \begin{cases} |\mathcal{I}|^{-1} \sum_{\forall n} \sum_{\forall k} v_n[k], & i \in \mathcal{I}, \\ 0, & i \notin \mathcal{I}. \end{cases} \quad (12)$$

See that the term $\mathbf{f}_i^\dagger \mathbf{w}$ in (11a) essentially depends on $\mathbf{b}$, which acts as a frequency bin selector based on the subcarriers indexes stored in $\mathcal{I}$. Moreover, note that the columns of the RIS configuration matrix (see (10)) are, in turn, determined by the normalized linear combinations of the DFT matrix columns. Therefore, the RIS reflectors are only required to be configured with a time-varying continuous phase-shift, through which each reflector remains passive, that is, $\|\boldsymbol{\Omega}_{\theta_{i,j}}\|_2^2 \leq 1 \forall i, j$.

As stated before, the row entries of $\boldsymbol{\Omega}_\theta$ are fixed, changing only when a new configuration is needed depending on the time index t . Note, however, that $T \geq K$, meaning that the minimum number of required configurations scales with the total number of subcarriers. This is equivalent to say that more time resources are needed as more bandwidth is used for the signal transmission. Therefore, more stringent requirements on the channel coherence time could pose a challenge for the frequency selectivity of the reflected signal, since multiple RIS configurations must be performed while the channel remains constant. This will be explored further in Section IV.

C. Motivation: A General Communications Scenario

Fig. 4 demonstrates a general communications scenario where the frequency selective reflection could be employed. It shows the intended UE receiving the AP reflected signal only on the selected subcarriers, whereas others UEs in the vicinity (denoted as UE'), are not hampered by unrestricted signal reflections. In principle, this means that signals transmitted by other APs, that is, AP', are not going to be inadvertently interfered by signals reflected from the RIS.

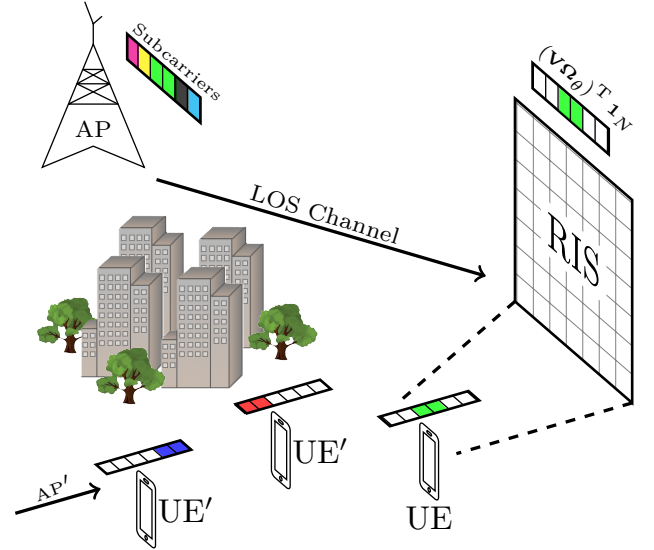


Fig. 4. A communications scenario where the RIS is employed without interfering with the signal received by other UEs (i.e. UE'). Note that the represented dimensions are not to scale and that this figure is better visualized in colors.

In what follows, computational simulations are leveraged to generate numerical results that corroborates the frequency selective properties of the proposed RIS configuration method.

IV. NUMERICAL RESULTS

In this work, numerical results are obtained under the following considerations for the communications system presented in Section II: (i) we ascribe $L_a = 101$ propagation paths for the AP-RIS channel, with each path delay randomly varying according to $\tau_a^l \sim \mathcal{U}[\tau_a^1, 2\tau_a^1]$, $\forall l > 1$ (the first path, $l = 1$, is the strongest LOS path); (ii) the same goes for the RIS-UE channel with $L_b = 51$ path delays; (iii) all $L_d = 100$ path delays of the direct channel (AP-UE) are

randomly distributed such that $\tau_d^l \sim \mathcal{U}[\tau_d, 2\tau_d]$ (τ_d is the time, in seconds, that the signal would take to propagate through the direct channel under LOS conditions); (iv) the azimuth and elevation angles of arrival/departure shown in Fig. 2 can also vary randomly around the LOS path initial angle, with $\phi_{a,b} \sim \mathcal{U}[-40^\circ, 40^\circ]$ and $\varphi_{a,b} \sim \mathcal{U}[-10^\circ, 10^\circ]$; (v) moreover, let $f_c = 3$ GHz be the central frequency of the transmitted signal, $B = 10.5$ MHz and $N_0 = -164$ dBm; and, finally, (vi) the dimensions of the RIS reflector, as indicated in Fig. 2, are defined to be $d_H = d_V = 0.25\lambda$ meters. The remaining parameters for the communications system [14] under analysis are described in Tab. I.

TABLE I
FIXED PARAMETERS OF THE CHANNEL MODEL. OBSERVE THAT
 $u \sim \mathcal{U}[0, 1]$ AND $n \sim \mathcal{N}(0, 1)$.

Parameter	Value
<i>Coordinates</i>	
(x_a, y_a, z_a)	(40, -200, 0) meters
(x_b, y_b, z_b)	(20, 0, 0) meters
<i>Number of Samples</i>	
M	$\left\lceil B \left(2\tau_a^{(1)} + 2\tau_b^{(1)} - \tau_d^{(1)} \right) \right\rceil + 11$ samples
<i>Propagation Losses</i>	
$\alpha^{(l)}$ (as in $\beta^{(l)}$)	$\gamma^{(l)} \frac{d_H d_V}{\lambda^2 / 4\pi} 10^{\frac{-30.18 - 26 \log(d_a)}{10}}$
$\gamma^{(l)}$	$\frac{\omega_P^{(l)}}{\sum_{\forall l} \omega_P^{(l)} (1 + \omega_R)}, \omega_P^{(l)} = 10^{-\tau_a^{(l)} + 0.2n}, \forall l > 1$
$\tau_a^{(1)}$	$\frac{\overbrace{\ (x_a, y_a, z_a) \ ^2}^{d_a}}{3 \times 10^8}$
$\gamma^{(1)}$	$\frac{\omega_R}{1 + \omega_R}, \omega_R = 10^{\frac{13 - 0.03d_a}{10}}$

A. Frequency Selectivity Performance

To determine how much of the reflected signal is received by unintended UEs, that is, the portion of the RIS reflection interfering on unselected subcarriers, we propose the following calculation

$$\frac{S}{I} = \frac{\sum_{\forall i} \|\mathbf{f}_i^\dagger \mathbf{h}_d\| |\mathcal{I}|^{-1} + \|\mathbf{f}_i^\dagger \boldsymbol{\vartheta}\|_2^2}{\sum_{\forall j} \|\mathbf{f}_j^\dagger \boldsymbol{\vartheta}\|_2^2}, \quad (13)$$

in which $i \in \mathcal{I}$ and, consequently, $j \notin \mathcal{I}$. In other words, the ratio between the combined signal magnitude ('S') of the selected subcarriers, over the combined signal magnitude ('I') of the remaining or interfered subcarriers, give us a metric to evaluate the frequency selectivity of the RIS reflection. Note that $\mathbf{f}_j^\dagger \mathbf{h}_d = 0, \forall j$, for simplicity, since we only wish to account for the interfering reflected signal on unselected subcarriers. From (12) it follows that (13) results in $S/I \rightarrow \infty$, granting a perfect selective reflection of signals. However, this result holds if at least $T = K$ is guaranteed. Conversely, if the number of discrete-time RIS configurations is insufficient²,

that is, if $T < K$, it can lead to an incomplete set of DFT coefficients being employed by the RIS configuration. As a consequence, the interfering reflected signals may become more prominent and less frequency selectivity is achieved.

In Fig. 5, the results of (13) are traced for different RIS sizes of N reflectors, where for the sake of brevity $T = N$. It was observed that the frequency selectivity performance depends largely on the number of discrete-time configurations, T . Consequently, the results for $N \neq T$ are omitted. Moreover, the total number of subcarriers³ is fixed to $K = 400$ and the number of selected subcarriers can vary from a single subcarrier $|\mathcal{I}| = 1$ to $|\mathcal{I}| = 39$ and $|\mathcal{I}| = 199$ subcarriers. As can be observed in Figures 5 (a) and (b), a pair of different selection methods are also adopted. The adjacent (Fig. 5 (a)) selection of subcarriers uses a reference subcarrier index, being drawn at random with equal probability, which is then surrounded by an even number of subcarriers for selective signal reflection. Alternatively, the random (Fig. 5 (b)) selection of subcarriers uses a random set of subcarriers, where subcarriers indexes are also chosen with equal probability, but which are not necessarily adjacent. The set of subcarriers indexes according to these selection methods can be given by

$$\mathcal{I} = \{j - \lfloor |\mathcal{I}|/2 \rfloor, \dots, j - 1, j, j + 1, \dots, j + \lfloor |\mathcal{I}|/2 \rfloor\}, \quad (14a)$$

$$\mathcal{I} \sim [\mathcal{U}[0, K - |\mathcal{I}|]], \quad (14b)$$

wherein $j = \{0, 1, \dots, K - 1\}$ and for which we have the adjacent and random selection methods represented in (14a) and (14b), respectively. Notice that for the random selection, $|\mathcal{I}|$ indexes are drawn with no reposition and that consecutive indexes are not allowed. These methods were chosen only as a baseline for evaluating the proposed frequency selective RIS configuration. To the best of authors knowledge, the problem of defining the best selection method for this context is still an open problem.

Therefore, see in Fig. 5 (a) that more interference is observed as the number of discrete-time configurations diminishes, specially when a single subcarrier ($|\mathcal{I}| = 1$) is selected for reflection. Likewise, in Fig. 5 (b), a similar behavior is observed, except that an inflection point is present for $T \sim 200$ when $|\mathcal{I}| = 199$ selected subcarriers are considered. Because the random method does not allow reposition and consecutive subcarriers selection, it becomes practically deterministic in this case. Although the indexes are still chosen at random, it is granted that every other subcarrier is eventually selected when $K/2$ subcarriers are used. In other words, this results in a fixed alternate subcarrier selection (e.g. $\mathcal{I} = \{0, 2, 4, \dots, K - 2\}$). Consequently, the subcarriers used for selective reflection are evenly spaced across all the bandwidth B . As verified in Fig. 5 (b), this has the effect of making the selective reflection considerable better for $T > 200$.

B. Achievable Rate Performance

By assuming that the condition for perfect selective reflection is met ($T = K$), then it remains to be investigated what

²Due to a slow or high number of total RIS configurations required.

³The optimum power allocation is computed by the water filling algorithm.

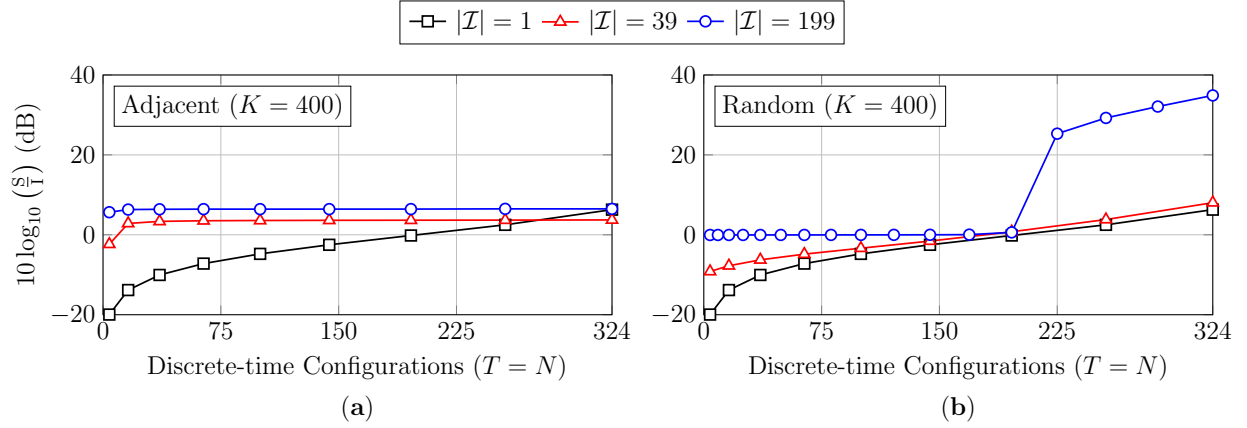


Fig. 5. The ratio between the combined signal magnitude of the selected subcarriers, over the signal magnitude of the remaining subcarriers is traced for different numbers of discrete-time configurations (or RIS sizes) and selection methods. Scenarios with $|I| = 1, 39$ and 199 selected subcarriers are evaluated. Each simulation point consists of averaging (up to) 5×10^3 channel realizations.

is the achievable rate performance. For this, we leverage the relative rate [14], which is defined as follows

$$R_{\text{Rel.}} = \frac{R}{R_C} \times 100 \text{ (\%)}, \quad (15)$$

wherein R is the achievable rate of (8) and the coherent rate, R_C , is defined by the maximum achievable rate. The coherent rate can be seen as the upper bound on the achievable rate, since it is based on the ideal conditions for which the direct and composite channels combine (in-phase) coherently at the UE [7], [14]. In summary, the relative rate allows us to quantify how reflected signals combine (constructively or destructively) with the signals transmitted via the direct channel.

Fig. 6 shows the relative rate results for a range of different quantities of selected subcarriers, where we also have $K = N = 400$. Additionally, the selection methods of (14a) (Adjacent) and (14b) (Random) are also present, with the addition of the fixed adjacent method. This method includes a restriction on the indexes available for the reference subcarrier, such that $\lfloor |I|/2 \rfloor \leq \min(j, K - j - 1)$ is always true. Consequently, the total number of selected subcarriers is always $|I|$ for the fixed adjacent method.

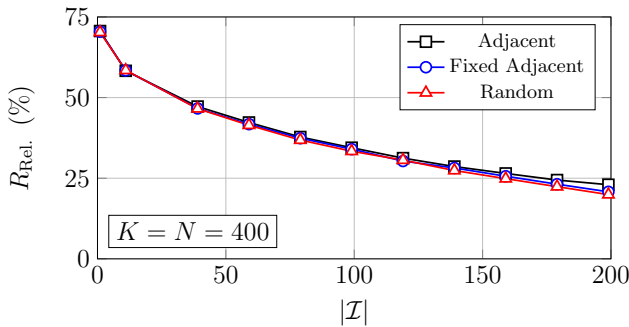


Fig. 6. Relative rate results are traced for a range of different quantities of selected subcarriers, also considering that $K = N = 400$. Besides the selection methods of (14a) and (14b), the fixed adjacent method is also illustrated. Each simulation point consists of averaging (up to) 5×10^3 channel realizations.

Recall that the RIS configuration is being used for a frequency selective signal reflection and, as such, its main objective is not necessarily the channel capacity or rate maximization. Nonetheless, note in Fig. 6 that $R_{\text{Rel.}} \simeq 75\%$ for a low number of selected subcarriers, representing a satisfactory performance. The relative rate performance, however, decreases steadily as more subcarriers are used for selective reflection. This contrasts with the results observed in Fig. 5, where generally more selected subcarriers mean better selectivity performance. Therefore, a trade-off exists between achievable rate and selectivity performance when determining the number of selected subcarriers. Observe also in Fig. 6 that the relative rate of the adjacent method is marginally better in relation to the other methods. This is caused by its lower average number of selected subcarriers, which is reassured by the fact (see also (14a)) that the fixed adjacent has a slightly lower relative rate.

Furthermore, it was verified that the general results observed in Fig. 5 (a) are also true for the fixed adjacent method, with slightly higher values of S/I for the fixed adjacent selection (considering $|I| = 39, 199$). The results obtained with higher values for K and, consequently, N and T , have also shown similar behavior to that observed in Fig. 5 and Fig. 6, except that the S/I and relative rate values are generally higher. These results are omitted since their findings do not bring any new relevant information.

V. CONCLUSIONS

The fundamental concepts that underpin RIS-aided communications systems was laid down in this tutorial. By leveraging a comprehensive wideband signal propagation model, it was demonstrated via practical examples that the RIS can be used to improve signal reception in point-to-point communications systems. Moreover, a frequency selective RIS configuration was put forward. It was also shown that this configuration allows the selection of an arbitrary subset of OFDM subcarriers for the signal reflection performed by the RIS. This configuration method avoids signal reflections in sub-bands that are not necessarily allocated for the intended

receiver. It can be concluded from the numerical results that the proposed RIS configuration can provide beneficial signal reflection properties and interference management, even under non-ideal conditions.

New developments and improvements for the RIS are still being reported in the literature. The BD-RIS [16], for instance, was proposed as an upgrade to the classical RIS, where a more complex hardware is able to provide a significantly improved performance in general. Consequently, in future works, these new relevant developments regarding the future of the RIS technology will need to be investigated thoroughly and presented for a larger audience.

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