Hybrid precoding applied to multi-beam Transmitting Reconfigurable Intelligent Surfaces (T-RIS)

David Demmer, Francesco Foglia Manzillo, Samara Gharbieh, Maciej Smierzchalski, Raffaele d’Errico, Jean-Baptiste Dore, Antonio Clemente

To cite this version:


HAL Id: cea-04416862
https://cea.hal.science/cea-04416862
Submitted on 6 Jun 2024

HAL is a multi-disciplinary open access archive for the deposit and dissemination of scientific research documents, whether they are published or not. The documents may come from teaching and research institutions in France or abroad, or from public or private research centers.

L’archive ouverte pluridisciplinaire HAL, est destinée au dépôt et à la diffusion de documents scientifiques de niveau recherche, publiés ou non, émanant des établissements d’enseignement et de recherche français ou étrangers, des laboratoires publics ou privés.

Distributed under a Creative Commons Attribution 4.0 International License
Hybrid Precoding Applied to Multi-Beam Transmitting Reconfigurable Intelligent Surfaces (T-RIS)

David Demmer *, Francesco Foglia Manzillo †, Samara Gharbieh ‡, Maciej Śmierzchalski, Raffaele D’Errico ‡, Jean-Baptiste Doré ‡ and Antonio Clemente *

CEA-Leti, Université Grenoble Alpes, 38000 Grenoble, France
* Correspondence: david.demmer@cea.fr (D.D.); antonio.clemente@cea.fr (A.C.)

Abstract: In this work, we study hybrid precoding techniques applied to multi-user Transmitting Reconfigurable Intelligent Surface (T-RIS) systems. The T-RIS considered here is a large array of electronically reconfigurable antenna elements illuminated by a small set of active sources. When it comes to digital signal-processing techniques applied to T-RIS systems, it is necessary to consider realistic models to bridge the gap with theoretical results. For this reason, we propose a multi-beam T-RIS propagation model with strong phase quantization constraints and limited beam codebooks. First, the proposed model is validated by characterizing a Ka-band T-RIS. Then, we optimize the quad-beam T-RIS structure by tuning the focal distance between the lens and the focal sources according to two metrics: (i) the per-user antenna gain (analog-only approach), and (ii) the per-user average rate (hybrid digital/analog approach). For both indicators, the system performance is evaluated in a multi-user scenario by assuming imperfect channel state information. We show that considering only the analog precoder is sufficient to optimize the T-RIS. However, the fully hybrid precoding scheme is required to deal with inter-user interference. We propose a codebook-aware optimization that improves the aperture efficiency of the T-RIS system.

Keywords: radio communications; multi-user MISO; spectral efficiency; beamforming; electronically steerable antenna; transmit array; millimeter wave

1. Introduction and Motivations

Ever-increasing data traffic continues to push wireless technologies beyond their limits. Each new generation is expected to increase the achievable capabilities by some orders of magnitude. To this end, three technological enablers have been considered in recent decades: (i) densifying the network, (ii) exploiting the spatial dimension, and (iii) moving to large non-congested spectra [1]. Millimeter-wave (mmWave) communications have received tremendous interest and are being standardized in 5G (up to 52.6 GHz) [2], and a further rise in frequency up to the sub-THz spectrum is even envisioned for beyond-5G (B5G) networks [3]. For cellular technologies, transmitting at mmWave frequencies leads to a tenfold increase in carrier frequency with respect to conventional 4G LTE upper bands. Consequently, over-the-air propagation suffers from far more severe propagation attenuations. Fortunately, the decrease in wavelength allows one to pack large antenna arrays to cope with the high path loss. Large antenna arrays can thus provide sufficient beamforming gain to ensure high-quality links but also enable spatial multiplexing by beamforming simultaneous multiple data streams [4]. Therefore, combining massive Multiple-Input Multiple-Output (mMIMO) with mmWave is promising. Now, the challenges to address are (i) how to efficiently control and reconfigure hundreds of antenna elements, and (ii) how to precode, i.e., beamform multiple data streams and ensure proper multi-user spatial-divison duplexing.

Conventional MIMO systems perform the entire precoding in the digital domain. Each data stream is controlled in amplitude and phase but it requires a dedicated radio
frequency (RF) chain for each antenna element. The required hardware thus scales with the size of the antenna array, which leads to high costs and enormous power consumption when large arrays are involved. MmWave MIMO is not a simple incremental evolution of conventional MIMO: the transceiver architectures must be rethought and redesigned and, consequently, the signal-processing techniques have to be re-adapted. A hybrid architecture has been proposed as an enabling technology of 5G massive MIMO [1,5,6]. In hybrid architectures, the number of RF chains scales with the number of users using the same time/frequency resource, and the precoding is split into both the analog and digital domains. Analog precoders rely on phase shifters and thus do not perform amplitude control. Analog precoders can be grouped into two structural classes: (i) a fully-connected network, where each antenna element is connected to an RF chain with RF lines and a phase shifter, and (ii) partially connected structures where each RF chain is connected to a subset of antenna elements. Thanks to the higher number of phase shifters, the fully connected approach provides more degrees of freedom and outperforms partially connected structures. However, such networks become unfeasible with large antenna arrays. Digital precoders can adjust both the phase and amplitude of transmitted signals as in fully digital systems but for an equivalent MIMO system scaled on the number of users instead of the number of antennas.

In recent years, a brand-new technology has attracted the attention of the scientific community: Reconfigurable Intelligent Surfaces (RISs) [7]. They are made of passive electromagnetic elements with some integrated electronics, e.g., switches, to control their scattering properties. For instance, Reflecting RISs (R-RISs) have been widely studied. They are usually placed far from the base station, typically on a wall, and act as electromagnetic mirrors by re-radiating incident waves in controlled directions to ensure enhanced channel metrics between the base station and the UEs and avoid RF blockages [8,9]. More recently, Transmitting RISs (T-RISs) have been proposed as a practical implementation of MIMO hybrid architectures [10–12]. With T-RISs, the electromagnetic surface is part of the transceiver structure and is placed at the focal distance of a few feeding sources. Unlike R-RISs, the incident wave propagates through the surface, which acts as a lens and can perform beam steering and control the direction of the re-radiated wave. The surface is composed of unit elements named unit cells. Unit cells based on Antenna–Filter Antennas (AFAs) are widely employed up to 140 GHz [13] because they can perform beamforming. They consist of a receiving antenna, a phase shifter, and a transmitting antenna. The unit cell antennas are typically microstrip patches. Additionally, in contrast to phased-array structures, there are no power dividers and networks of phase shifters. Therefore, with T-RISs, the power loss between the RF chains and the transmitting antennas is due to the propagation channel between the focal sources and the unit cells (including the antenna gains) and the losses induced by the propagation through the unit cells. In addition, the number of phase shifters is reduced in contrast to a fully connected structure but is equivalent to a partially connected structure (i.e., one phase shifter per unit cell so one per transmitting antenna). Consequently, they are less compact than phased arrays because of the focal distance between the focal sources and the lens and can achieve high array gain and good beam-steering resolution at more affordable costs. The design of hybrid precoding techniques applied to T-RISs has only recently been studied [12] and only addresses point-to-point MIMO. This work aims to study and design hybrid precoders for multi-user MIMO (MU-MIMO) using T-RISs. We consider the case where the user equipment (UE) is equipped with single antennas and thus we focus on MU multiple-input single-output (MU-MISO) systems.

The designed MU-MISO hybrid precoder differs from the point-to-point MIMO case [4,6,12,14–16]. In point-to-point MIMO, the receiving antennas are all collocated. Ideal combining can thus be assumed at the receiver side and the optimization can focus on the precoder design by maximizing the mutual information. An optimal precoder can be obtained by singular value decomposition (SVD) [4]. Consequently, a conventional approach to the design of hybrid precoders in point-to-point MIMO is to approximate the optimal unconstrained precoder using an orthogonal matching pursuit (OMP) algorithm [6,12]. It
uses the sparsity of the mmWave propagation channels and leads to a solution sufficiently close to the unconstrained optimal [4,14]. It is also possible to decouple the problem into two sub-problems by individually optimizing the RF precoder and then adapting the digital precoder. The process can be performed iteratively [12] or sequentially [15,16]. It is shown that the second approach slightly outperforms the first one [12]. However, the provided gain is limited, especially when large antenna arrays are considered.

Except in MU-MISO, UE combiners are not taken into account, and therefore inter-user interference (IUI) has to be entirely managed by the precoder, leading to completely different design constraints. To simplify the optimization problem, the designs of the RF and digital precoder are usually decoupled [17–19]. The MU-MISO optimization problem is thus simplified by first designing the RF precoder in order to maximize the power received by the users (neglecting the IUI) and then by adapting the digital precoder to compensate for the propagation-induced distortion and mitigate the IUI. The design of the RF precoder commonly assumes single-path channels between the antenna array and each UE [17,18] and the digital precoder often relies on zero forcing (ZF), which is known to be sub-optimal [18,19]. The aforementioned works consider phased-array structures, however, their results cannot be directly extended to T-RISs because the equivalent propagation channel differs.

The design of RIS elements has been intensively studied. Although varactor-based designs provide a quasi-continuous phase resolution, they achieve a very limited bandwidth [20–23]. Therefore, strong phase quantization constraints have been considered in the design of the unit cells, e.g., the use of a few PIN diodes, to enable wideband performance and practical feasibility [24–34]. More recently, innovative solutions based on Phase Change Material (PCM) switches have been proposed to generate energy-efficient RISs [35]. However, the use of advanced integration technologies such as CMOS could enable the development of fine-resolution cells and RISs up to sub-THz frequencies [36]. Some examples of RIS designs are summarized in Table 1. Phase quantization directly affects the design of RF precoders. The determination of the optimal RF precoder is not straightforward when only finite-resolution phase shifters are available. A possible approach is to minimize the distance between the optimal phase solution [37] and the phase profile by using a discrete set of available values.

Table 1. State-of-the-art T-RIS unit-cell designs in various frequency bands.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Reconfigurable Device</th>
<th>Phase Quantization</th>
<th>References</th>
</tr>
</thead>
<tbody>
<tr>
<td>C-band</td>
<td>Varactors</td>
<td>continuous</td>
<td>[20,21]</td>
</tr>
<tr>
<td>C-band</td>
<td>Varactors and PIN diodes</td>
<td>16-bit</td>
<td>[22]</td>
</tr>
<tr>
<td>X-band</td>
<td>PIN diodes</td>
<td>1-bit</td>
<td>[24,25]</td>
</tr>
<tr>
<td>Ku-band</td>
<td>PIN diodes</td>
<td>1-bit</td>
<td>[26–28]</td>
</tr>
<tr>
<td>K-band</td>
<td>Varactors</td>
<td>continuous</td>
<td>[23]</td>
</tr>
<tr>
<td>Ka-band</td>
<td>PIN diodes</td>
<td>1-bit</td>
<td>[29,30]</td>
</tr>
<tr>
<td>Ka-band</td>
<td>MEMS switches</td>
<td>2-bit</td>
<td>[31]</td>
</tr>
<tr>
<td>Ka-band</td>
<td>PIN diodes</td>
<td>2-bit</td>
<td>[32–34]</td>
</tr>
<tr>
<td>D-band</td>
<td>PCM-based switches</td>
<td>1-bit</td>
<td>[35]</td>
</tr>
<tr>
<td>300 GHz</td>
<td>CMOS switches</td>
<td>8-bit</td>
<td>[36]</td>
</tr>
</tbody>
</table>

In this context, we propose to jointly optimize the T-RIS structure and the MU-MISO precoder. Our main contributions can be summarized as follows:

(i) We derive an analytical propagation model that includes a T-RIS propagation channel and constraints on the resolution of the phase shifters. The obtained model is general and can be easily adapted to most use cases by setting the array geometry, the illumination law, and the pertinent radiation patterns for the focal sources and unit cells.

(ii) The proposed model is experimentally validated through the characterization of a novel wideband T-RIS operating at Ka-band frequencies. The RIS is based on a
1-bit reconfigurable unit cell with PIN diodes. It has been specifically designed and builds on our previous works [29,38].

(iii) We propose and compare two methods to optimize the T-RIS: (i) maximization of the per-user gain (RF-only approach), and (ii) maximization of the per-user rate (hybrid approach). To this end, we fix the array geometry and the set of unit cells (1-bit design) and then we determine the optimal focal distance, i.e., the distance between the lens and the focal sources, with respect to the two aforementioned metrics.

(iv) We provide an evaluation of the two metrics for four served users with imperfect channel knowledge, with an emphasis on the impact of the beam codebook size.

The rest of the paper is organized as follows. In Section 2, the system model and the T-RIS propagation channel are described. In Section 3, the joint T-RIS structure and MU-MISO hybrid precoder optimization problem are presented. The two considered design methods are also discussed. The measurement campaign and the comparison with our model are presented in Section 4. Section 5 is dedicated to the performance evaluation and discussion of the achievable performance. Finally, the concluding remarks and perspectives for future studies are presented in Section 6.

Notations: Bold-faced lower-case and upper-case letters are used to, respectively, denote vectors and matrices; \( \angle \), \((.)^{-1}\), and \((.)^H\) represent the angle, inversion, and Hermitian (transpose conjugate) operators; and \(|.|\) and \(\|.|\|_2\) denote the modulus and Euclidean vector norms.

2. System Model

2.1. Coordinate System

A Cartesian referential \((O; \vec{O}_x, \vec{O}_y, \vec{O}_z)\) is considered in this work, where the origin \(O\) is placed at the center of the T-RIS lens. The Cartesian coordinates of any point \(P\) are denoted by \((x, y, z)\).

Spherical coordinates \((r, \theta, \phi)\) are also used, where the radial distance \(r \geq 0\) corresponds to the distance from the origin, the elevation angle \(\theta \in [0, \pi/2]\) corresponds to the angle with respect to the \(\vec{O}_z\) axis, and the azimuth angle \(\phi \in [0, 2\pi]\) is the angle between the \(\vec{O}_x\) axis and the orthogonal projection of the point \(P\) over the reference plane \((Oxy)\).

The U/V space coordinates are used for the antenna radiation patterns and gains. The relation between the two coordinate systems is given in (1):

\[
\begin{align*}
u(\theta, \phi) &= \sin(\theta) \cos(\phi) \\
v(\theta, \phi) &= \sin(\theta) \sin(\phi)
\end{align*}
\] (1)

2.2. Analytical Model: T-RIS

We consider a T-RIS equipped with \(N_s\) feeding sources and a lens composed of \(M\) unit cells as shown in Figure 1a. The feeding sources are active antennas, typically horn antennas, connected to dedicated RF chains, i.e., there are \(N_s\) RF chains. The sources are placed on a rectangular grid with distances in the \(x\)-direction and \(y\)-direction equal to \(d_s\). The lens located on the focal source plane (assuming that the lens is placed on the plane \(z = 0\), the sources are located on the plane \(z = -d_f\), where \(d_f\) is the source focal distance) and its unit cells are uniformly placed according to a rectangular grid, with \(d_x\) and \(d_y\) being the respective distances in the \(x\)- and \(y\)-directions as depicted in Figure 1b. Each unit cell is composed of a receiving passive antenna (toward the feeding sources), a passive phase shifter, and a transmitting passive antenna (toward the UEs).
On the one hand, the directivities of the feeding sources $D_{FS}$ are modeled as ideal and are expressed in (2), where $p$ is named the antenna order [12,39]:

$$D_{FS}(\theta,\phi) = 2(p + 1) \cos^p(\theta)$$

(2)

On the other hand, each unit-cell antenna is modeled as a uniform aperture element whose directivity $D_{UC}$ is expressed by (3) [39]:

$$D_{UC}(\theta,\phi) = \pi \cos(\theta)$$

(3)

In this study, we assume that the radiation patterns are symmetrical, i.e., they do not depend on the azimuth angle $\phi$. Thus, for the rest of this article, they are omitted from the directivity expressions for the sake of simplicity.

We can model the frequency propagation channel between the feeding source $n$ and the unit-cell $m$, denoted by $l_{m,n}$, as a direct line-of-sight (LOS) link, as expressed in (4) [12]:

$$l_{m,n} = \frac{\lambda \sqrt{D_{FS}(\psi_{m,n}) D_{UC}(\psi_{m,n})}}{4\pi r_{m,n}} e^{-2\pi r_{m,n}/\lambda}$$

(4)

where $r_{m,n}$ and $\psi_{m,n}$ are the distance and elevation angles between the feeding source $n$ and the unit cell $m$ and $\lambda$ corresponds to the signal wavelength. $\psi$ is used for the elevation angles between the sources and unit cells and $(\theta, \phi)$ for the UEs.

Each unit cell applies an independent phase shift $2\pi \beta_m$ to the received signal before transmitting it to the transmitting passive antenna. We assume that no signal attenuation is induced by the phase shifters.

Figure 1. (a) Geometry for the analysis of the T-RIS, and (b) Schematics of hybrid digital/analog multi-user beamformer.
2.3. Analytical Model: Over-the-Air Propagation Channel

MmWave channel models typically describe a low multi-path profile with a few dominant clusters [6,40]. We can thus assume a one-path direct line-of-sight model, as expressed in (5), between a unit cell \( m \) and a receiving UE antenna \( k \):

\[
h_{k,m} = \frac{\lambda \sqrt{D_{UC}(\theta_{k,m})}}{4\pi r_{k,m}} e^{-j2\pi r_{k,m}/\lambda}
\] (5)

Assuming far-field propagation, i.e., \( \forall m \ r_{k,m} = r_k \), \( \theta_{k,m} = \theta_k \) and \( \phi_{k,m} = \phi_k \), it is possible to derive the propagation channel expression between the T-RIS lens and the same receiving antenna \( k \), as given in (6):

\[
h_{k} = \frac{\lambda}{4\pi r_k} e^{-j2\pi r_k} \sqrt{D_{UC}(\theta_k)} h_{d}(\theta_k,\theta_k) \otimes h_{v}(\theta_k,\theta_k)
\] (6)

where the vectors \( h_{u}(\theta_k,\theta_k) = \left[e^{j2\pi d_x(m)\phi_k}\right]_m \) and \( h_{v}(\theta_k,\theta_k) = \left[e^{j2\pi d_y(m)v_k}\right]_m \).

2.4. Baseband System Model

We consider a multi-user MISO (MU-MISO) system with \( K \) users, each equipped with a single antenna, and a T-RIS with \( N_s \) feeding sources and a lens composed of \( M \) unit cells, as described in Section 2.2:

\[
y = \sqrt{P_{tx}}HF^{RF}TF^{D}x + z
\] (7)

where \( P_{tx} \) represents the maximum transmit power of each RF chain, \( x \in C^{K \times 1} \) and \( y \in C^{K \times 1} \) are the transmit and receive baseband signals, \( H = [h_0 \ldots h_{K-1}]^T \in C^{K \times M} \) is the propagation channel matrix, \( T \in C^{M \times N_s} \) is the transmit array channel whose coefficients are defined in (4), \( F^{RF} = \text{diag}(e^{j2\pi \beta_m}) \in C^{M \times M} \) is the RF precoder (i.e., the unit-cell phase shifts), \( F^{D} \in C^{N_s \times K} \) is the digital precoder, and \( z \in C^{K \times 1} \) denotes the additive white Gaussian noise of complex variance \( \sigma^2 \).

3. Problem Formulation and Precoder Design

3.1. Problem Formulation

The T-RIS structure consists of three parts: the source-lens propagation channel \( T \), which is fixed and known once the T-RIS is manufactured, and the RF and digital precoders, \( F^{RF} \) and \( F^{D} \), which jointly aim to perform the beamforming, i.e., focusing the transmit power in the direction of the users and mitigating the inter-user interference (IUI). The two precoders can be updated in real time to adapt to the instantaneous propagation channel and user locations, whereas the antenna array is fixed once manufactured.

We address the challenge of designing the T-RIS structure and the hybrid precoding scheme to maximize the achievable sum-rate \( R \), as expressed in (8) [41]:

\[
R = \sum_{k=0}^{K-1} \log \left( 1 + \frac{|h_k F^{RF} F^{D}_{k}|^2}{|\sum_{k \neq k} h_{k} F^{RF} F^{D}_{k}|^2 + \frac{\sigma^2}{P_{tx}}} \right)
\] (8)

The resulting optimization problem can thus be expressed as follows:

\[
\text{maximize} \quad R \quad \text{subject to} \quad \forall k, \|f^{D}_{k}\|_2 \leq 1, \quad \forall m, f^{RF}_{m,m} = 1, \quad \forall m, \angle f^{RF}_{m,m} \text{ belongs to a finite set}
\] (9a, 9b, 9c, 9d)
The constraints (9b) and (9c) are power limitations: (9b) enforces the maximum power (the digital precoder cannot act as a power amplifier) and (9c) means that the RF precoder does not impact the signal power (neither the signal amplification nor power loss due to the phase shifters). The phase quantization constraint (9d) assumes the use of finite-resolution phase shifters (the phase resolution depends on the unit-cell design). Therefore, the optimization problem (9) is a mixed integer programming problem [17]. Consequently, to solve this problem it is (i) necessary to have knowledge of all the channel coefficients \( \mathbf{H} \) between the \( M \) unit cells and the \( K \) UEs, and (ii) required to perform a full search over all the possible RF precoders. In practice, the digital precoder only has knowledge of the equivalent propagation channel \( \mathbf{H}_\text{RF} \mathbf{T} \), which includes the over-the-air propagation coefficients \( \mathbf{H} \), the RF precoder \( \mathbf{F}_\text{RF} \), and the T-RIS channel \( \mathbf{T} \). Therefore, to the best of the authors’ knowledge, there is no known closed-form solution to this problem.

When it comes to phased arrays, the problem is usually solved by decoupling the designs of the RF and the digital precoders \([19,37]\). When T-RIS structures are involved, there are more degrees of freedom that need to be included in the optimization problem, e.g., see the presence of \( \mathbf{T} \) in (9a). This approach requires the knowledge of the \( K \) UE locations with respect to the antenna arrays and \( N_s \times K \) equivalent channel coefficients. The constraints of the required channel knowledge are, therefore, much more relaxed. The decoupled optimization is described below.

### 3.2. RF Precoder

With multi-source structures, all the \( M \) unit cells are divided into \( N_s \) subarrays. We consider the case \( N_s = K \), which means that each feeding source is associated with a user. Each subarray thus aims at beamforming in the direction of its served user. The responding RF precoder applied to the subarray \( k \) can thus be defined as \( \mathbf{F}_\text{RF}^k = \text{diag}(e^{j2\pi\beta_k^k(m)}) \) with

\[
\forall m \in \mathcal{M}_k, \quad \beta_k^k(m) = \angle \mathbf{h}_u(\theta_k, \phi_k) - \angle t_k(m)
\]

(10)

where \( \mathcal{M}_k \in [0, M - 1] \) denotes the set of unit cells of subarray \( k \). The \( K \) sets \( \mathcal{M}_k \) are pairwise distinct, i.e., each unit cell belongs to a unique subarray and serves a unique user. This RF precoder is well known in the literature and is commonly used for MISO systems (i.e., without combiners) \([12,37]\).

The RF precoder (10) assumes infinite-resolution phase shifters. However, for practical implementations, finite resolutions must be considered. Therefore, we can define sets of phases quantized with \( B \) bits of precision by minimizing the Euclidean distance with respect to the unquantified optimal phases:

\[
\forall k \in [0, K - 1], \forall m \in \mathcal{M}_k, \beta_k^k(m) = \arg \min_{\hat{\beta}_k^k(m) \in [0,2^B - 1]} \left| \mod \left( \beta_k^k(m), 2\pi \right) - \frac{2\pi \hat{\beta}_k^k(m)}{2^B} \right| \]

(11)

In this study, we also investigate the impact of codebook size. Indeed, when it comes to practical implementation, the number of possible beams is limited. Larger codebooks contain more beams and are able to target any position with higher accuracy. However, they also suffer from beam overlapping and thus increased inter-user interference when users are close to each other. Therefore, the optimal number of beams results from a trade-off between coverage and beam overlapping. In this work, we consider codebooks that uniformly span the field of view, as depicted in Figure 2. One of our objectives is to assess to what extent the codebook size impacts the T-RIS optimal design and system performance. To this end, we considered four codebook sizes from a very limited number of beams (21) to a very high number of beams (111).
3.3. Equivalent Channel and Digital Precoder

The digital precoder aims to cancel the channel-induced distortion and IUI. The equivalent channel, denoted by $H_{eq}$, seen by the receiver includes the actual propagation channel between the Tx and Rx, the distortion induced by the RF chains, and even the RF processing. For the considered system model in this article, the corresponding expression is given in (12):

$$H_{eq} = HF_{RF}T$$ (12)

In practice, the equivalent channel is not perfectly known by the receiver and is estimated from known reference signals and time/frequency interpolation. The imperfect knowledge of the estimated channel $\hat{H}_{eq}$ can be represented by adding white Gaussian noise to the instantaneous channel, as expressed in (13):

$$\hat{H}_{eq} = H_{eq} + W_{H}$$ (13)

ZF is a commonly used digital precoder for MISO systems [37] and its unconstrained expression is given in (14):

$$F_{\mu}^D = \hat{H}_{eq}^H \left( \hat{H}_{eq} \hat{H}_{eq}^H \right)^{-1}$$ (14)
The given unconstrained expression is unlikely to satisfy the power constraint (9b). A pragmatic and sufficient solution is thus to normalize with respect to the maximum power among the RF chains, which leads to the power-constrained expression (15), where $F_D^i$ denotes the $i$th row:

$$F_D^i = \frac{1}{\max_j ||F_D^j||_2} F_D^i$$  \hspace{1cm} (15)

### 3.4. Optimization of T-RIS Structure

As shown in (9), the design of the T-RIS structure is also part of the optimization problem. Once the RF precoder is defined, the optimization problem amounts to determining the optimal source positions with respect to the lens. As mentioned, the lens is divided into $K$ subarrays and each of these is associated with a feeding source and a user. It is known that aligning the sources with the center of each subarray is optimal \[12,39\] as it minimizes the taper loss in partial illumination over the subarray. The question is thus how to determine the optimal focal distance $d_f$, i.e., the distance between the plane of the sources and the plane of the lens.

It is possible to re-express (4) as (18) in order to emphasize its dependence on the focal distance. The distance $r_{k,m}$ and the elevation angle $\theta_{k,m}$ between the feeding source $k$ located at $(s_x(k), s_y(k), -d_f)$ and the unit cell $m$ placed at $(d_x(m), d_y(m), 0)$ can be expressed, as shown in (16) and (17), by setting $X_{k,m} = d_x(m) - s_x(k)$ and $Y_{k,m} = d_y(m) - s_y(k)$. As a reminder, we assume that the lens defines the plane $z = 0$ and thus the feeding sources are all located on the plane $z = -d_f$, where $d_f$ is the focal distance.

$$r_{k,m} = \sqrt{d_f^2 + X_{k,m}^2 + Y_{k,m}^2}$$  \hspace{1cm} (16)

$$\theta_{k,m} = \tan^{-1}\left(\frac{\sqrt{X_{k,m}^2 + Y_{k,m}^2}}{d_f}\right)$$  \hspace{1cm} (17)

By applying $\cos(\tan(x)) = (x^2 + 1)^{-1/2}$ over (2) and (3), one obtains:

$$t_{m,k} = \frac{\lambda}{4 \pi} \sqrt{2\pi(p+1)} \left(\frac{d_f^{(p+1)/2}}{d_f^{(p+5)/4}} \right)$$  \hspace{1cm} (18)

In this article, we investigate and compare two methods to determine the optimal focal distance $d_f$:

- **Gain optimization:**
  The T-RIS structure $T$ is determined by maximizing the antenna array gain, as defined in (19). By doing so, $T$ is the only function of the RF precoder.

$$G(\theta_k, \phi_k) = |h^H(\theta_k, \phi_k) F_{RF} \left(\sum_{k=0}^{K-1} t_k\right)|^2$$  \hspace{1cm} (19)

The optimization problem can thus be expressed as follows:

$$\hat{d}_f = \max_{d_f} \int_{\theta_{\text{max}}}^{\theta_{\text{max}}} \int_{\phi_{\text{max}}}^{\phi_{\text{max}}} G\left(d_f\right) d\theta_k d\phi_k$$  \hspace{1cm} (20)

- **Capacity optimization:**
  Another proposed method is to directly optimize the T-RIS structure $T$ to maximize the per-user capacity (8). Contrary to the optimization of the gain, the IUI is taken
into account here and, therefore, so is the impact of the ZF precoder. The optimization problem can be stated as follows:

$$\hat{d}_f = \maximize_{d_f} R(d_f)$$ (21)

The objective of this investigation is to assess whether the optimal focal distance found by taking into account the digital precoder and the IUI (capacity optimization) is different from that determined by maximizing the antenna gain. This result has an immediate impact on the optimal T-RIS design and its manufacturing. To this end, the optimal focal distances obtained for the two optimization problems are determined in Section 5. Then, the evaluations of two indicators are presented: the antenna gain and the per-user capacity.

4. Unit-Cell Design and Characterization

In this section, we consider a specific T-RIS configuration inspired by our previous works [29,38]. The flat lens was composed of 20 × 20 unit cells and was illuminated by a single focal source. To achieve a wideband design, the unit-cell geometry was opportuneely optimized with respect to our previous works.

4.1. Unit-Cell Design and Frequency Behavior

The proposed 1-bit reconfigurable unit-cell architecture (with a periodicity of 5.1 mm) is presented in Figure 3a. The architecture consisted of four metal layers, two identical substrates of Rogers Duroid RT602 and one bonding film of Arlon CuClad 6700. The receiving layer was composed of a rectangular patch loaded with a U-shaped slot, whereas the transmitting layer contained a rectangular patch loaded with an O-shaped slot and two PIN diodes. In the electromagnetic (EM) simulator, these active devices were modeled as a lumped-element equivalent circuit (a series L-R circuit and a shunt R-C circuit) extracted from the measurements and a gallium–arsenic block [42]. The active patch was connected to the passive one with a metalized hole placed at the center of the unit cell. A ground plane occupied one of the two intermediate layers. The other inner layer contained the biasing lines.

![Unit-cell design](image_url)

Figure 3. Cont.
Figure 3. Proposed T-RIS architecture design and performance: (a) 3D sketch of the 1-bit reconfigurable PIN diode unit cell, (b) unit-cell measurement setup in a standard WR-28 waveguide, (c) simulated and measured transmission coefficients in amplitude with an embedded photograph of the measurement setup, and (d) simulated and measured transmission coefficients in phase.

In the proposed unit-cell structure, two PIN diodes (reference MA4AGP907 manufactured by M/A-COM) were selected as active devices for their low insertion loss and limited size. One was biased in the forward state with a 10 mA current. The 1.3 V threshold voltage of this diode was sufficient to maintain the other diode, which was mounted in an anti-parallel configuration, in its reverse state. Thus, only one bias line was necessary for each unit cell, thereby facilitating the layout and routing of the bias network in very large array configurations. The bias line was 100 $\mu$m wide and was connected to the active patch using two symmetric metallic connections. More details on the 1-bit unit cell operational principle, model, parametric analysis, and simulation setup are provided in [29].

The unit cell was simulated using the commercial Ansys FSS software considering periodic boundary conditions and Floquet port excitations and taking into account the mutual couplings with the surrounding unit cells. The full-wave simulations under normal incidence provided the scattering parameters calculated for the two phase states. The simulated minimum insertion loss was only 0.61 dB at 30.0 GHz with a 3 dB transmission bandwidth of 23.6–30.0 GHz (23.7% at 27.0 GHz) for the unit cells at 0° and 23.8–31.1 (26.9% at 27.0 GHz) for the state at 180°. The phase difference between the two states was around 180° on the whole frequency band. The 1-bit unit-cell bandwidth performance is synthesized in Table 2.

<table>
<thead>
<tr>
<th>Unit Cell</th>
<th>Insertion Loss (dB) at 28.0 GHz</th>
<th>−2 dB Bandwidth (% @ 27.0 GHz)</th>
<th>−3 dB Bandwidth (% @ 27.0 GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0°</td>
<td>1.50</td>
<td>22.9</td>
<td>23.7</td>
</tr>
<tr>
<td>180°</td>
<td>2.10</td>
<td>25.5</td>
<td>26.9</td>
</tr>
</tbody>
</table>

As detailed in [42], the unit cell was fabricated and characterized in a waveguide simulator based on standard WR-28 waveguides (7.11 × 3.556 mm²), as shown in Figure 3b. The measured and simulated amplitudes of the scattering parameters obtained with the waveguide setup are depicted in Figure 3c,d. A very good agreement between the simulated and experimental results was observed. Nonetheless, there was an inevitable difference between these results and the scattering parameters calculated using the periodic boundary conditions and normal incidence. In particular, we observed a 500 MHz shift toward lower frequencies. This was mainly due to the oblique incidence of the waves in the waveguide setup [29,42]. In the case of the standard rectangular waveguide WR-28, the equivalent incidence angle was equal to 45°.
4.2. Validation of the Proposed Model

The accuracy of the proposed model for the analysis of the T-RIS is demonstrated in this section by comparing the experimental and numerical results. The prototype was composed of a lens of $20 \times 20$ of 1-bit unit cells. The array was illuminated by a 10 dBi horn aligned to its center at a focal distance of 60 mm. The antenna beam was steered at different scan angles by calculating the corresponding precoding matrix at 28.0 GHz. The measured and computed gain patterns are presented in Figure 4. The numerical results, shown in blue, were computed using the model described in Section 2, considering perfectly transmitting unit cells, an ideal feed with a $\cos^4(\theta)$ pattern, and a T-RIS periodicity of 5.1 mm along both the $x$- and $y$-directions. The amplitude differences between the numerical results and measurements (red lines) were in the order of 1.5–2.0 dB depending on the scan angle, which was mainly due to the insertion loss of the unit cells. On the other hand, a strong agreement with the experimental data was observed when the theoretical model was refined considering the simulated radiation pattern of the isolated horn, as well as the patterns and transmission coefficients of the unit cell, in its two operating modes (black lines). These data were extracted from the full-wave simulations of the unit cell under normal incidence and by applying periodic boundary conditions. The beamwidth and sidelobe levels of the measured patterns were well predicted by both the theoretical and refined models. The measured and calculated gains of the broadside T-RIS are plotted as functions of frequency in Figure 5. The measured gains were 1.8 dB and $\leq0.1$ dB lower than the values obtained using the theoretical and refined models, respectively.

![Figure 4. Radiation patterns for a mono-source T-RIS equipped with $20 \times 20$ lens (28.0 GHz, focal distance 60 mm, focal source = horn antenna with 10 dBi gain).](image-url)
Figure 5. Gains of the broadside T-RIS (mono-source configuration) as a function of frequency: numerical and experimental results. The RF precoder was specified at 28.0 GHz for a focal distance of 60 mm.

5. Multi-User Performance Evaluation

5.1. Definitions of Scenarios

For the performance evaluation, we considered a multi-beam T-RIS, as described in Table 3. It was composed of the same lens as the T-RIS considered in the previous section but was equipped with four distinct focal sources in order to handle four different beams. The T-RIS was placed on the ceiling of an indoor room at an arbitrary height of 2.0 m, illuminating the space below. We considered four distinct users randomly placed according to the three different scenarios defined in Table 4. The objective was to emphasize the impact of the user density and antenna aperture limit.

For the rest of this work, we consider imperfect channel knowledge by performing a channel state estimation (13) with additional thermal noise (temperature of 25 °C).

Table 3. Multi-source multi-beam T-RIS.

<table>
<thead>
<tr>
<th>Carrier Frequency</th>
<th>Lens</th>
<th>Focal Source</th>
</tr>
</thead>
<tbody>
<tr>
<td>28.0 GHz</td>
<td>uniform square grid</td>
<td>2 × 2</td>
</tr>
<tr>
<td></td>
<td>20 × 20 unit cells</td>
<td>square arrangement with distance D/2</td>
</tr>
<tr>
<td></td>
<td>size $D \times D$ with $D = 20\frac{\lambda}{2} = 10.7$ cm</td>
<td>horn antennas 10 dBi gain ($p = 4$)</td>
</tr>
<tr>
<td></td>
<td>1-bit phase quantization</td>
<td></td>
</tr>
<tr>
<td></td>
<td>omnidirectional antennas</td>
<td></td>
</tr>
</tbody>
</table>

Table 4. Scenarios.

<table>
<thead>
<tr>
<th>Scenario</th>
<th>Desk</th>
<th>Office Room</th>
<th>Factory Hanger</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max aperture $\theta_{\text{max}}$ ($^\circ$)</td>
<td>20</td>
<td>60</td>
<td>80</td>
</tr>
<tr>
<td>Radius [m]</td>
<td>0.7</td>
<td>3.5</td>
<td>11.3</td>
</tr>
<tr>
<td>Area [m$^2$]</td>
<td>0.8</td>
<td>18.8</td>
<td>202.1</td>
</tr>
</tbody>
</table>

5.2. Optimal Focal Distance

The first objective was to determine the optimal focal distance with respect to the two performance indicators: (i) the maximization of the per-user gain (20), and (ii) the maximization of the per-user capacity (21). Because the two optimization problems (20) and (21) cannot be analytically solved, we proposed to numerically evaluate the solution.
for the T-RIS defined in Table 3 at 28.0 GHz. For this section, we did not consider codebooks for the RF precoder but we did consider the 1-bit phase quantization. The results for the two performance indicators are, respectively, depicted in Figure 6a and Figure 6b.

For each configuration, there was an optimal focal distance. Indeed, on the one hand, small focal distances imply large fluctuations in the density of the received power by the lens, i.e., high taper loss. On the other hand, considering large focal distances limits the power received by the lens due to spillover radiation [12]. The optimal focal distance, therefore, symbolizes a balance between those two effects.

One can see that the optimal focal distance was the same for the two optimization problems. This means that there was no need for a priori knowledge of the deployment scenario or to consider the presence of IUI to optimize the antenna structure. This implies that without considering the codebook limitations, the antenna system design could be performed separately from the MU-MISO precoder. For the rest of this study, we thus consider that the optimal focal distance is \( \hat{d}_f = 0.26D = 279 \) mm for the reference (i.e., without codebooks) system. One can see that the focal distance for the multi-source case was reduced with respect to the mono-source configuration (600 mm). Indeed, when multiple focal sources were involved, the lens was divided into several quadrants, one for each focal source. Each source thus needed to be placed in order to illuminate its own quadrant. Consequently, the sources needed to be placed closer to the lens in order to limit unwanted emissions over the quadrants of the other sources. Multi-source T-RIS was proposed in [39] for a point-to-point link (i.e., multi-source mono-beam) to reduce the focal distance and make the antenna system more compact. In our work, multiple sources were required in order to provide multiple beams and simultaneously serve several users.

It seems worth pointing out that the previous point does not mean that an RF-only precoder is sufficient to ensure reliable multi-user links. It just means that there is no need to consider the baseband precoder (i.e., ZF precoder) for the T-RIS optimization. However, the ZF was required to mitigate the IUI, as shown in Figure 7.

To illustrate the impact of the baseband precoder, we considered the two aforementioned performance indicators, the per-user gain and per-user capacity, and we evaluated each one for two cases: (i) the RF-only precoder, and (ii) the full hybrid precoder (RF + baseband ZF). On the one hand, one can see that the ZF reduced the per-user gain (left figure). Indeed, in order to mitigate the IUI, the ZF applied a complex factor (attenuation + phase shift) to each user stream. The total transmit power was, therefore, reduced. The gain penalty was higher when beam overlapping was more likely (i.e., \( \theta = 20 \)). On the other hand, full hybrid precoding significantly outperformed the RF precoder in terms of the achievable link capacity (right figure). Indeed, the RF precoder alone generated large IUI because of beam overlapping, even with an ideal number of beams (no codebook size limitations).
5.3. Codebook-Aware Optimization

Our second objective was to emphasize the benefits of taking into account the codebook sizes for the optimization of the antenna structure. For this section, we consider the gain optimization (20) and the codebooks, as described in Figure 2, and determine the optimal focal distance for each codebook. Then, we evaluate the resulting average per-user capacity and compare the results with the optimization when codebooks were not considered. Table 5 presents the results for the reference optimization, where the optimal focal distance was optimized without taking into account the codebook sizes (see the previous section), and the corresponding per-user capacities are provided. It also presents the results of the codebook-aware optimization (i.e., c-a opt.) and compares the results with the first approach. The performance gains are given in bold. Table 6 presents the results for the focal distances where the performance gains are also given in bold.

Table 5. Achievable per-user capacity (bits/s/Hz) for reference and codebook-aware optimizations.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>21</td>
<td>ref opt.</td>
<td>20.0</td>
<td>17.6</td>
<td>14.0 (−6.9%)</td>
<td>13.1</td>
<td>10.8 (−7.6%)</td>
<td>14.8 (−7.2%)</td>
<td>13.1</td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>20.1 (+0.5%)</td>
<td>18.2 (+5.7%)</td>
<td>14.0 (+6.9%)</td>
<td>13.1</td>
<td>10.8 (−7.6%)</td>
<td>14.8 (−7.2%)</td>
<td>13.1</td>
<td></td>
</tr>
<tr>
<td>43</td>
<td>ref opt.</td>
<td>20.4</td>
<td>18.9</td>
<td>13.8</td>
<td>13.8</td>
<td>13.8</td>
<td>13.8</td>
<td>13.8</td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>20.5 (+0.5%)</td>
<td>19.1 (+1.1%)</td>
<td>14.8 (+7.2%)</td>
<td>13.8</td>
<td>13.8</td>
<td>13.8</td>
<td>13.8</td>
<td></td>
</tr>
<tr>
<td>73</td>
<td>ref opt.</td>
<td>20.4</td>
<td>19.5</td>
<td>14.5</td>
<td>14.5</td>
<td>14.5</td>
<td>14.5</td>
<td>14.5</td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>20.5 (+0.5%)</td>
<td>19.6 (+0.5%)</td>
<td>15.3 (+5.5%)</td>
<td>14.5</td>
<td>14.5</td>
<td>14.5</td>
<td>14.5</td>
<td></td>
</tr>
<tr>
<td>111</td>
<td>ref opt.</td>
<td>20.6</td>
<td>19.85</td>
<td>15.1</td>
<td>15.1</td>
<td>15.1</td>
<td>15.1</td>
<td>15.1</td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>20.6 (+0.0%)</td>
<td>19.8 (+0.0%)</td>
<td>15.8 (+4.6%)</td>
<td>15.1</td>
<td>15.1</td>
<td>15.1</td>
<td>15.1</td>
<td></td>
</tr>
</tbody>
</table>

Table 6. Optimal focal distances $d_f/D$ for reference and codebook-aware optimizations.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>21</td>
<td>ref opt.</td>
<td>0.26</td>
<td>0.14 (+23.1%)</td>
<td>0.14 (−46.2%)</td>
<td>0.10 (−61.5%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>0.26</td>
<td>0.22 (+15.4%)</td>
<td>0.21 (−19.2%)</td>
<td>0.12 (−47.8%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>43</td>
<td>ref opt.</td>
<td>0.26</td>
<td>0.22 (+15.4%)</td>
<td>0.21 (−19.2%)</td>
<td>0.11 (−57.7%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>0.26</td>
<td>0.22 (+15.4%)</td>
<td>0.21 (−19.2%)</td>
<td>0.11 (−57.7%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>73</td>
<td>ref opt.</td>
<td>0.26</td>
<td>0.22 (+15.4%)</td>
<td>0.21 (−19.2%)</td>
<td>0.11 (−57.7%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>0.26</td>
<td>0.22 (+15.4%)</td>
<td>0.21 (−19.2%)</td>
<td>0.11 (−57.7%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>111</td>
<td>ref opt.</td>
<td>0.26</td>
<td>0.22 (+15.4%)</td>
<td>0.21 (−19.2%)</td>
<td>0.11 (−57.7%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>c-a opt.</td>
<td>0.26</td>
<td>0.22 (+15.4%)</td>
<td>0.21 (−19.2%)</td>
<td>0.11 (−57.7%)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
First, one can see that because of the codebook limitations, the achievable per-user capacities were reduced with respect to the reference performance, especially for large apertures $\theta_{\text{max}} \geq 60$. Performing the codebook-aware optimization helped to bridge the gap with respect to the reference performance: up to a $+6.9\%$ gain for codebook 21 and a minimum $+4.6\%$ gain for $\theta_{\text{max}} = 80$. The gain provided by the codebook-aware approach was significant when small codebooks and/or large apertures were considered. Additionally, one can see that the codebook-aware approach led to smaller focal distances. The benefits of the codebook-aware approach were thus multifold: (i) it helped the implementation by improving the performance with small codebooks, (ii) it increased the effective aperture of the antenna system, and (iii) it helped to reduce the form factor of the T-RIS structure.

It seems important to mention here that the presented results were from a performance evaluation for a 28.0 GHz T-RIS with low complex hybrid precoding schemes. The achieved performance strongly depended on the assumptions we made, e.g., for the codebook patterns and the RF and baseband precoders. However, the T-RIS ray-tracing propagation model is valid for all mmWave bands. The method we applied in this study can thus be easily re-used for other numerical applications. Additionally, as we mentioned, one can adapt the codebook-aware optimization methods to minimize the focal distance. For such an application, it may be interesting to consider more realistic sources and unit-cell radiation patterns, as well as near-field propagation effects; otherwise, the practice may diverge from theoretical expectations.

6. Conclusions and Perspectives

The conclusion has been reworked. In this work, we proposed the use of T-RIS for practical multi-user massive MIMO access. First, a ray-tracing channel model was derived, including wave propagation from the focal sources through the T-RIS to the UEs, considering unit cells with finite-resolution phase shifters. Low-complexity RF and baseband precoding schemes were applied. The proposed model was experimentally validated in terms of the antenna gain and radiation patterns by comparing it with the measurements of a specifically designed Ka-band T-RIS. The model is scalable and can be easily adapted to different use cases. Then, we investigated two approaches for optimizing the T-RIS focal length once the array size, focal source, and T-RIS phase profile were determined: (i) array-gain maximization, and (ii) SINR maximization. We showed that the two approaches led to the same results and, consequently, the conventional array-gain maximization was sufficient to determine the optimal focal distance in addition to being easier to perform. However, even in ideal working conditions, the analog beamforming was not sufficient to mitigate IUI and, therefore, hybrid precoding with the baseband ZF was necessary to achieve high sum rates. These results were found by considering infinite codebook sizes (no limitations on the number of beams). We again investigated codebook-aware optimization methods based on gain maximization for different codebook sizes. We obtained a significant sum-rate enhancement with respect to unlimited codebook sizes, especially when large fields of view were involved, thanks to the reduction of the focal length.

The considered approach can also be used to investigate the performance with reduced focal distances. However, for electrically short focal lengths, the model should be extended to take into account the near-field interactions of the source and the T-RIS unit cells. The proposed model and optimization methods, therefore, represent an interesting tool to assist the design of T-RIS optimizations with coarse phase resolution, as well as system designers, to help bridge the gap between antennas and system performance.

**Perspectives**

Based on the outcomes of this study, we believe that the following perspectives should be studied:

(i) Power allocation: We observed during the performance evaluation that low, complex ZF can strongly attenuate user streams when beam overlapping occurs. It
limits the TX gain and, by extension, the coverage. We believe that a solution with improved power efficiency would greatly improve system performance.

(ii) Source and unit-cell radiation pattern: One challenge for T-RIS is to limit its form factor and investigate systems with reduced focal distances. However, to do so, it is necessary to consider more accurate modeling of sources and unit-cell radiation patterns. In this case, it is possible to use models derived from measurements.

(iii) Wideband transmissions and beam split: With the rise in frequency, signal bandwidth increases and some wideband-induced effects can occur such as beam split. Beam split means that the beam direction moves with the frequency, which induces a power loss and higher IUI. This effect must be taken into account for high-mmWave and sub-THz system design.

**Author Contributions:** Modeling: D.D., F.F.M., J.-B.D., R.D. and A.C.; measurements: S.G., M.´S. and A.C.; writing—original draft preparation: D.D., F.F.M. and A.C.; writing—review and editing: all authors; project administration and funding acquisition: J.-B.D. and A.C. All authors have read and agreed to the published version of the manuscript.

**Funding:** This work has been funded by the Institut Carnot CEA-Leti (5GmmFLASH project), the French government under the “France 2030” investment program (5G-OPERA project) and the Recovery Plan (CRIIOT project), and the French National Research Agency (ANR-20-CE25 MESANGES project).

**Data Availability Statement:** Not applicable.

**Conflicts of Interest:** The authors declare no conflict of interest.

**References**

33. Diaby, F.; Clemente, A.; Sauleau, R.; Pham, K.T.; Dussopt, L. 2 Bit Reconfigurable Unit-Cell and Electronically Steerable Transmitter Array at Ka-Band. IEEE Antennas Propag. 2020, 68, 5003–5008. [CrossRef]


**Disclaimer/Publisher’s Note:** The statements, opinions and data contained in all publications are solely those of the individual author(s) and contributor(s) and not of MDPI and/or the editor(s). MDPI and/or the editor(s) disclaim responsibility for any injury to people or property resulting from any ideas, methods, instructions or products referred to in the content.