A 2D-Programmable and Scalable Reconfigurable Intelligent Surface Remotely Controlled via Digital Infrared Code

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Abstract—Reconfigurable intelligent surfaces (RISs) are promising and relatively low-cost tools for improving signal propagation in wireless communications. A RIS assists a base station (BS) in optimizing the channel and maximizing its capacity by dynamically manipulating the reflected field. Typically, RISs are based on dynamically reconfigurable reflectarrays (RAs), i.e., 2-D arrays of passive patch antennas, individually switchable between two or more reflection phases. The spatial resolution of provided reflected field patterns is governed by the aperture dimensions and the number of patches to meet the requirements of different communication scenarios and environments. Here, we demonstrate a 1 bit RIS for 5 GHz Wi-Fi band made by assembling together multiple independently operating and structurally detached building blocks all powered by the same DC source. Each block contains four separately phase-switchable patch antennas with varactor diodes and a common microcontroller extracting digital control commands from modulated infrared light. Such distributed light-sensitive controllers grant the possibility of scaling the aperture by adding or removing blocks without redesigning any control circuitry. Moreover, in the proposed RIS a full 2-D phase encoding capability is achieved along with a robust remote infrared control.

Index Terms—Control with light, patch array, reconfigurable intelligent surface (RIS), reflectarray (RA), varactor diode.

I. INTRODUCTION

MODERN and prospective wireless communications systems face a growing number of simultaneously operating user’s terminals and strong interference in urban outdoor and indoor environments. To keep reliable coverage and high channel capacity, conventional fixed antenna systems become insufficient. One of the proposed approaches to improve such systems consists in using passive structures for controlling electromagnetic field distributions. Tunable reflective meta-surfaces (MSs)—electrically dense and flat periodic structures of subwavelength scatterers) and reflectarrays (RAs—periodic arrays of passive individually tuned antennas with typically a half-wavelength spacing) used for that purpose are commonly called reconfigurable intelligent surfaces (RISs) [1], [2], [3], [4], [5]. These structures are considered in the literature as powerful and relatively cheap solutions for improving microwave and millimeter-wave wireless channels [6], [7]. In outdoor environments, a RIS can assist a fixed base station (BS) dynamically modifying the channel by creating one or several steerable reflected beams as a response to the wave impinging the RIS from the BS. In indoor environments, a RIS can optimize the link by dynamically modifying the spatial distribution of partially standing waves in a given room created by the radiation of a wireless access point, as proposed in [6]. In both cases, one or several RISs are to be strategically placed in the environment with a predefined location of the signal source.

RISs based on MSs consist of deeply subwavelength unit cells and have small periods compared to λ, while RAs typically have a period of around λ/2 (λ is the wavelength in free space). In contrast to RAs, MSs operate in the regime of strongly coupled unit cells, so they are modeled as effectively homogeneous impedance boundaries rather than arrangements of discrete elements [8], [9]. Despite MSs potentially offer more advanced reflection control such as achieving perfect reflection to angles larger than 70°–80° from the normal [10], the features of their unit cells are much finer than of RAs. For that reason, RAs better fit the relevant technological constraints that are especially important in the millimeter-wave range [11] and, therefore, serve as RISs in most papers. RAs usually consist of passive metal patch antennas placed over a common ground plane and excited by the incident wave.

To form a desirable reflected field pattern, the corresponding distribution of the reflection (scattering) phase over the aperture is to be computed and applied to every individual patch. The phase can be controlled by changing the resonant frequency of the patch, e.g., through biasing a diode connected between the patch and the ground plane [12] or between two halves of the patch. Gradual and dynamic variation of phase can be provided by varactor diodes [13], [14], but for a large amount of individually controlled patches it requires complex and expensive electronic circuits. To simplify the design and make it cheaper, 1 bit (also called digital [15]) RISs have been proposed approximating the phase profile

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with only two reflection phase states at the expense of high parasitic diffraction lobes in the scattering pattern [16]. A 1 bit RISs grant a compromising directivity for the main reflection direction provided that the number of patches is high enough (practically, several hundreds and more) [17]. Discrete phase variations can be achieved either with varactors [18], [19], [20] or positive-intrinsic-negative (p-i-n) diodes. Several 1 bit RISs switched using p-i-n diodes have been proposed in [15], [21], [22], [23], and [24]. A 2 bit version of a RIS controlled with p-i-n diodes was presented in [25]. While at frequencies below 10 GHz both varactor and p-i-n diode can be employed, at higher frequencies [26], especially in the 5G FR2 band [11], [27], [28] and above [29], p-i-n diodes have been more popular due to lower losses.

Special DC circuits are to be embedded into the design of a RIS to individually bias the diodes for setting phases of individual patches (or groups of patches). In order to decouple the DC and microwave parts of the RIS, the DC circuits including biasing lines and controllers are usually placed on the back side of a multilayer printed circuit board (PCB). On the contrary, the patches are placed on the front side. For isolation of the two sides, resonant traps (RF chokes) can be used at the corresponding vias [20], [29], [30]. Embedded biasing circuits can contain distributed controllers placed in every unit cell or usually a group of unit cells. Alternatively, a centralized control unit [21], [22], [31] can be employed. In both cases, changing the aperture dimensions and a number of patches (hereinafter referred to as scaling) requires redesigning the entire RIS.

The problem of scaling can be solved by using light to deliver bias voltages from a centralized control unit to diodes or to deliver control signals from remote control to every element of the distributed controllers. Among others, control with light is the most attractive because the reflection phase is encoded in a noncontact manner and with high switching speed [32], i.e., without any physical wire connection for transferring control data.

Controlling resonant unit cells in the microwave range with light was first proposed in [33] and further demonstrated in electromagnetic structures with various functions, e.g., in [24], [34], [35], and [36]. Light-controllable RISs have been implemented in several ways as follows. In [19], 1-D rows of a RIS were phase-controlled using varactors biased with voltages generated by p-i-n photodiodes in response to applied visible light of white light-emitting diodes (LEDs). By tuning the intensity of LEDs, the RIS was encoded to produce one or two reflected beams. In [37] a RIS with stepwise phase patterns was composed of blocks each one containing 4 × 4 patches loaded with varactor diodes. The reflection phase of each of 6 × 6 blocks was determined by the intensity of light coming from a meandered chain of LEDs placed behind the RIS. In [31] an infrared-controlled RIS capable of 1 bit phase encoding was proposed. Compared to the visible light control, this approach provided longer remote control distances and higher energy efficiency. The RIS had a centralized control unit with several phase patterns precoded in FPGA and enabled when receiving coded signals from a remote infrared transmitting circuit. Recently, a scalable RIS with full 2-D phase control via dynamically reconfigurable light field illumination from a liquid-crystal display placed behind it has been demonstrated in [20]. However, to our knowledge, none of the RISs presented in the literature combines full 2-D remote control of phase using light with scaling capability.

In this work, we propose and demonstrate a 1 bit RIS with full 2-D independent phase control via IR light. In this approach schematically shown in Fig. 1 IR light coming from a distant source is modulated with digital series control sequences and illuminates the whole aperture of the RIS. The structure is composed of identical and structurally independent building blocks all powered by the same DC source. Each block containing four independently phase-switchable patches responds only to its own precoded address from a sequence and then extracts the information about the required phase states for its four patches. In contrast to the previous approaches, the proposed RIS can be freely scaled by changing the number of blocks without redesigning any circuitry or light sources. Distant and robust digital control via IR makes the proposed RIS advantageous for communications inside a room.

The rest of this article is organized as follows. In Section II, we describe the design procedure for one building block and the entire RIS including the method for extraction of equivalent varactor’s parameters. In Section III, the results of the experimental investigation are given and compared to simulations and theoretical predictions.

II. DESIGN AND MANUFACTURING OF RIS

In this section, we describe the design procedure for one building block of the proposed scalable RIS and of an entire reflective aperture with 20 × 20 individually controllable patches with 2-D phase encoding capability operating in the
Wi-Fi 5 GHz range. Also, the design of the remote control unit is discussed.

In order to experimentally show the proposed principle the following technical requirements were set: frequency range covering Wi-Fi channels 40, 42 (5.17–5.25 GHz), vertical polarization, reflecting aperture dimensions of 600 × 600 mm (approximately 10λ × 10λ).

Four unit cells were grouped in a building block with one microcontroller separately switching the varactors of four patches between two states. The entire 1 bit RIS was composed of such building blocks all being powered from the same DC voltage. Surface-mount-device (SMD) varactors SMV2019-040LF by Skyworks Solutions [38] were chosen.

A. RIS Building Block and Unit Cell

The building block with four identical rectangular patches loaded with two varactor diodes each is schematically shown in the inset of Fig. 1. All four copper patches were printed on the top of a multilayer PCB with a common ground plane for the microwave and DC parts of the board. The substrate of the patches was made of Rogers 4003C (RO4003) material with a thickness of 1.5 mm, relative permittivity of 3.38 and dielectric loss tangent of 0.0027. The DC part was organized using multiple additional layers of FR4 behind the ground plane having a total thickness of 1.0 mm, substrate permittivity of 4.3, and dielectric loss tangent of 0.025. The RO4003 and FR4 parts were connected by 0.44 mm-thick prepreg KB-6065 with a dielectric permittivity of 4.2 and dielectric loss tangent of 0.013.

To minimize cross-polarization and simplify the interface between the microwave and DC parts of the RIS, we chose a modified version of the classical rectangular patch with dimensions L and W symmetrically loaded with two varactor diodes located near the radiating edges [39] (see Fig. 2). The patch is assumed to operate with TE polarization (XZ plane of incidence) of the incident field having only y-component of an electric field at any incidence angle. The ground plane was kept at a zero DC potential, while the patch was under a positive bias potential formed by the DC circuit. With this aim, the center of the patch was connected to the DC circuit through a via going through a hole in the ground plane. Also, two additional vias were made to connect the diodes to the ground (marked as GND in the figure). To place the diodes and properly connect them, two rectangular windows with dimensions a × b were made in the patch and two small contact plates connected to the GND vias were added. Each diode was soldered between the edge of its window and the corresponding contact plate. For TE polarization of the incident field, microwave currents are induced only on the patch and ground plane, but no current is induced along the vias. This symmetry allows avoiding RF chokes, which means saving space on the board and reducing losses.

Another advantage of the employed design is the possibility to adjust the effect of the diode’s tunable capacitance to the resonant frequency of the patch. The closer the diode and its GND via to the radiating slot, the stronger the reflection phase variation for the same bias voltage. By varying distance s between the windows with varactors it was possible to select two voltage levels corresponding to the 1 bit phase states within an available voltage range of the DC supply for the selected diode. Preferably, the difference between the voltages corresponding to both states should be as large as possible. This condition ensures that the difference in capacitance of the diode switched between two states is considerably larger than any parasitic capacitance due to soldering and mounting on a board. Moreover, that difference is to be larger than a typical capacitance deviation given in the datasheet of the diode. In our case, for stable operation of the unit cell, the difference in capacitance should exceed 0.4–0.5 pF, which was considered when choosing separation s.

For the 1 bit operation we chose “0” state with a reflection phase of −90° and “1” state with a reflection phase of +90°. Both states correspond to equal detuning of the patch resonance from the central operational frequency providing that the magnitudes of the reflection coefficient are equal (with smaller losses than at the resonance). In order to numerically predict the required voltage levels and correctly estimate the losses introduced by the RIS upon reflection, equivalent-circuit parameters of the diode were extracted from waveguide measurements as described below.

B. Extraction of Varactor Diode’s Parameters

To characterize the varactor diode as a series RLC equivalent circuit we made a set of PCBs serving as resonant waveguide terminations. Like the unit cell of the RIS, each PCB contained a resonant patch with windows and additional contact plates to place two diodes. However, the patch dimensions were changed in order to fit to the cross section dimensions of WR229 (58.17 × 29.08 mm). With patch dimensions L = 30.5 mm and W = 19.6 mm the resonance in the reflection coefficient spectrum was around 4.2 GHz with a zero voltage applied to the diodes (initial estimation was made based on the data from the manufacturer’s datasheet). The other parameters according to Fig. 2 were: b = 5 mm, a = 6 mm, w = 0.5 mm, l1 = 1.7 mm, d = 0.2 mm, s = 13.9 mm, 2r = 0.4 mm. Three different terminations were manufactured each one having three PCB layers: the top one with 1.5 mm-thick Rogers 4003C substrate, the bottom one with 0.5 mm-thick Rogers 4003C substrate, and the middle one with 0.2 mm-thick prepreg RO4450 substrate (dielectric permittivity of 3.52 and dielectric loss tangent of 0.004). The first termination shown in Fig. 3(a) had two varactors soldered, while the other two terminations used for calibration had no varactors. In one of them, the patch was disconnected from

Fig. 2. Unit cell of the proposed RIS consisting of a rectangular patch on the top layer of a multilayer PCB.
impedance of the PCB loads normalized to a characteristic waveguide spectra to the measured ones. With this aim, impedance curves their variation in the numerical model to fit the simulated voltages eventually selected for the experimental prototype used for varactor diodes.

The diode’s parameters were extracted based on the comparison between the simulated and measured spectra of the reflection coefficient at the input of the waveguide section with a length of 500 mm. All numerical simulations in this work were made in CST Microwave Studio 2020. In simulations both varactors were represented by lumped elements with sizes 0.3 × 0.475 mm. The impedance of the lumped elements was described by the series-type equivalent circuit with capacitance \( C_d(U) \) and resistance \( R_d(U) \) depending on bias voltage \( U \) as well as parasitic lead inductance \( L_d \) and parasitic capacitance \( C_{par} \) both assumed independent on \( U \). The circuit is shown in the inset in Fig. 3(a). The measurements were carried out in the range of 3–5 GHz with a step of 0.5 MHz on a vector network analyzer (VNA) Agilent E8362C calibrated using a standard coaxial kit via coaxial-to-waveguide junction [see Fig. 3(a)]. Parasitic ripples in reflection coefficient spectra caused by multiple reflections in the waveguide section between the sample and coaxial-to-waveguide transition were suppressed using the time gating technique. For the PCB with two diodes, a bias voltage \( U \) was varied from 0 to 5 V with a step of 0.2 V.

Two examples of the measured reflection coefficient spectra are given in Fig. 3(b) and (c) for \( U = 0 \) V and 3.2 V (the bias voltages eventually selected for the experimental prototype of the RIS). Equivalent-circuit parameters were obtained by their variation in the numerical model to fit the simulated spectra to the measured ones. With this aim, impedance curves of the PCB loads normalized to a characteristic waveguide impedance \( Z_0 \) were used [see Fig. 3(d) and (e)]. The frequency of the maximum real part (the resonance), the maximum level for the real part, and the slope of the imaginary part at the resonance were the values to control when fitting. Note that for different \( U \), all three criteria were met only by varying the capacitance \( C_d \) and resistance \( R_d \) of the diode.

The extracted parameters were the following: \( L_d = 0.2 \) nH, \( C_{par} = 30 \) fF. For \( U = 0 \) V the main parameters were \( C_d = 2.1 \) pF and \( R_d = 7.5 \) Ω, while for \( U = 3.2 \) V, \( C_d = 0.87 \) pF and \( R_d = 7.1 \) Ω. The same parameters were further used in the optimization of the unit cell of the RIS at 5.2 GHz.

C. Organization and Operation of RIS

Based on the extracted parameters of the diode, the unit cell was optimized via numerical simulations. The goal was to obtain “0” and “1” phase states at the central frequency of 5.2 GHz for the normal incidence and polarization along the \( y \)-axis. The unit cell was modeled with a Bloch-Floquet port in conjunction with unit-cell boundary conditions. The resulting parameters (all in mm) read: \( L = 16.6, W = 22.0, a = 2.0, b = 2.6, w = 0.475, l_1 = 0.725, d = 0.5, s = 11.58, 2r = 0.7 \). The reflection coefficient was calculated in the range of 4.9–5.5 GHz [shown with solid lines in Fig. 4(a) and (b)]. As can be seen, the chosen parameters indeed provide that “0” and “1” states realized with \( U = 0 \) V and \( U = 3.2 \) V, correspondingly, differ by approximately 180° in the reflection phase. The phase difference holds within the range of 150°–210° (maximum deviation of phase difference of 30°) at frequencies from 5.12 to 5.28 GHz. Note that this operational range covers the target bands of Wi-Fi channels 40 and 42 (5.17–5.25 GHz). The reflection coefficient magnitude of around 0.5 in the operational range can be explained by dissipation losses mostly associated with the varactors.

Four identical unit cells were combined in one building block having its own control unit that receives control sequences via infrared light regardless of its position within the RIS aperture. The control sequences were formed by the IR remote control unit. Fig. 5(a) shows its organization scheme, where the remote consists of a transmitting near-infrared (940 nm) LED, amplitude modulator, and a USB interface to a computer. The IR communication channel used amplitude
modulation and control commands to produce a series code sequence with a carrier frequency of 38 kHz. The command protocol provided 128 unique addresses (from 0 to 127) which were more than enough to control 100 blocks (i.e., 400 patches).

Within a block, all four patches with four pairs of varactor diodes, and the related components of the biasing circuit were contained within an area of around $\lambda \times \lambda$ on the multilayer PCB having both microwave and DC parts. The biasing circuit contained one TSOP34338 optoelectronic IR receiver module, one ATTiny441 microcontroller, and one driver circuit per patch. Fig. 5(b) shows the scheme of the biasing circuit of a single building block. The multilayer PCB of the block had a circular through hole in the center between the corners of four patches, which allowed the IR receiver to be sensitive to IR illuminating the face of the RIS. This solution is advantageous over previous light-controlled RISs illuminated from the back as it simplifies mounting low-profile RISs on walls for indoor applications.

Each block receives IR light illuminating the entire aperture of the RIS and extracts its own control commands from the digital code containing the block’s unique address and reflection states of four patches. As a response to the remote’s commands, the required four bias voltages are set by drivers each one containing a single field-effect transistor (FET), two Shottky diodes, and three resistors. Depending on the FET state either one or the other power supply is connected to the driver’s output via a corresponding resistor and Shottky diode. The voltages are then applied to the central vias of four corresponding patches. Four possible voltage combinations are to be stored in the controller. To adjust the voltage levels related to “0” and “1” states for the given varactor diode, each driver only required changing the voltages of two external power supplies.

To test the DC circuit, a prototype of the block was manufactured. Its bottom side (with DC biasing circuit seen) and top side (with four patches and the through hole seen) are shown in Fig. 5(d) and (e), accordingly. Elements numbering in Fig. 5(d) and (e) correspond to one in Fig. 5(b). The manufactured prototype of the IR remote unit is depicted in Fig. 5(c).

Even one single block being connected to a dual-channel DC power supply may operate as the simplest RIS with only four unit cells. However, to form a practical electrically large aperture, typically considered in the application, many independently controlled and structurally detached blocks can be assembled together. It is enough to connect all the blocks to the same DC power supply. This scaling principle allows one to build an intelligent surface with a desirable form and size. Note that adjacent blocks should have their ground planes interconnected. Otherwise, gaps between the blocks may affect the operation of the RIS due to resonant diffraction effects. Typically RISs have electrically large aperture dimensions, while RISs with small amounts of patches do not make sense. Therefore, it can be recommended to compose the entire RIS from panels each one having at least $5 \times 5$ identical proposed blocks. Such large panels can be assembled to a common RIS so that the edges of their PCBs stay close to each other without any electrical connection. The corresponding long gaps are non-resonant and will not affect the RIS operation. In the experiment, we used four square $5\lambda \times 5\lambda$ panels to compose a RIS with a total amount of $20 \times 20$ patches.

**D. Near- and Far-Field Phase Coding**

The developed RIS belongs to the class of 1 bit RAs with full 2-D phase coding capability. To create a desirable spatial distribution of the reflected field, the corresponding phases of individual patches are to be precalculated and encoded. The calculation of phases was realized in frames of self-developed software that also controlled the IR remote unit for sending commands to the RIS.

The following adaptive algorithm was developed to calculate the phases based on the specified distribution of the reflected field. Both the far-field distribution (scattering pattern) and intermediate-field distribution (field pattern on a target plane parallel to the aperture) could be set as the input data. Before starting the adaptation, the RIS is set to the initial state when all patches are in the same “1” state. However, the algorithm could start from any arbitrary phase distribution. The use of a preliminary approximation usually did not influence the final result, but could significantly speed up the adaptation process.

First, the field generated by the RIS in the target area was calculated using a simplified analytical model. The model took
to the aperture, the reflected field is

\[ E_{\text{RIS}}(k) = \sum_{m=1}^{N_x N_y} E_{\text{inc}}^m e^{j\Phi_{\text{inc}}^m} \cdot |R_m| e^{j\Phi_m} e^{-jr_{mk}} \left(1 + \frac{z_{\text{plane}}}{r_{mk}}\right) \]  (1)

where \( E_{\text{inc}}^m \) and \( \Phi_{\text{inc}}^m \) are the magnitude and phase of the incident field at the location of patch element \( m \), \( |R_m| \) — magnitude of the local reflection coefficient at patch \( m \), \( \Phi_m \) — reflection phase of patch \( m \) (\(-\pi/2 \) for state “0” and \(+\pi/2 \) for state “1”), \( r_{mk} \) — distance between patch \( m \) and point \( k \), \( z_{\text{plane}} \) — distance from the target plane to the aperture of the RIS, \( N_x, N_y \) — number of patches along \( x \) - and \( y \)-axes, respectively. In the case when the source is a plane wave, the phase of the incident field was found as

\[ \Phi_{\text{inc}} = k_0 x_m \cos(\phi_{\text{inc}}) \sin(\theta_{\text{inc}}) + y_m \sin(\phi_{\text{inc}}) \sin(\theta_{\text{inc}}) \]  (2)

where \( x_m, y_m \) are Cartesian coordinates of the center of patch element \( m \), \( \phi_{\text{inc}} \) and \( \theta_{\text{inc}} \) — azimuth and polar incidence angles in the spherical coordinate system with a polar axis being normal to the RIS, \( k_0 = 2\pi/\lambda \) — wavenumber in free space.

Second, the difference between the desired and calculated field distributions was determined. The objective function that needed to be minimized was a mean squared error (MSE) between the two magnitude distributions created on the target plane.

Third, the phase state of one of the patches (selected at random) was flipped and the corresponding change in the objective function was calculated. If the objective function decreased the new phase state was kept, otherwise, the element was returned to its original state and the process repeated.

The criteria to stop the iteration process was the absence of changes in the phase distribution for a specified maximum number of iterations. It should be noted that the number of steps depends on the number of patches and desirable field distribution. For the experimental RIS with \( N_x = N_y = 20 \) it was enough to limit the process with 50 iterations.

The above-described objective function can be useful for maximizing the reflected power concentration in the chosen region of space. If this region is small enough, and the distance to the target plane is much larger than the dimensions of the RIS (i.e., \( z_{\text{plane}} \gg N_x \lambda/2, N_y \lambda/2 \)), this process is close to simple beam steering. However, the target region may be large and its dimensions may compare to the ones of the aperture. In that case, the process can be applied to holographic field synthesis. In Section III, we consider both types of adaptation made with the experimental RIS.

**III. Experimental Results**

For the experimental characterization, we fabricated a full-sized prototype of the RIS consisting of \( N_x \times N_y = 20 \times 20 \) patch elements (i.e., \( 10 \times 10 \) IR-controlled building blocks with individual controllers inside). For simplicity of fabrication, all blocks were grouped into four identical square panels each one being a separately fabricated multilayer PCB. To support the panels and fix their edges close to each other, a special holder was made of 5 mm-thick Plexiglass. The PCBs were mounted to the holder with plastic screws. The prototype
with full dimensions of 600 × 600 mm fixed on the holder is shown in Fig. 6(a) from the front side and in Fig. 6(b) from the backside. DC supply Rohde&Shwarz HMP2030 was used to power the building blocks providing the required voltage levels.

Measurements of reflected field distributions were made using a three-axis field scanner installed in an anechoic chamber (7 × 5 × 3 m) and VNA Rohde&Schwarz ZVB20. The first port of the VNA was connected to the source antenna with vertical linear polarization ($E_z$) while the second one—to a field probe. The experimental setup for the cases of scattering pattern and reflection coefficient measurements is shown in Fig. 6(c) with the source being a linearly polarized TEM-horn antenna at the distance of 4 m from the RIS. In this case, for obtaining a scattering pattern, first, a complex magnitude of total magnetic field $H_z$ was measured over the plane parallel to the RIS at the distance of 10 mm away from the patches. Next, the map was remeasured in absence of the RIS (when all four PCB panels together with a Plexiglass holder were removed from the chamber) and was subtracted from the map measured in presence of the RIS to obtain the complex magnitude of the scattered field. Both maps were scanned within the area of 600 × 600 mm with a resolution of 15 mm. Then, the Fourier transform was applied to calculate the far-field scattering pattern from the near-field map of the reflected field using standard methods [40]. In the case of holographic field synthesis, the field was measured directly on the target plane. Depending on precalculated 1 bit phase profiles, different scan angles of an anomalously reflected TE-polarized beam were obtained in the horizontal plane. For comparison, the magnitude and phase distributions of the $x$-component of the magnetic field were recorded. The same value in the absence of the RIS and in the case where the RIS was replaced with a copper plate of the same dimensions was also measured to determine a complex reflection coefficient as described, e.g., in [41] and [42]. The effect of parasitic signal reflections between the RIS and the horn antenna was reduced using the time gating procedure [43].

The magnitude and phase of the measured reflection coefficient are compared with the simulation results for $U = 0$ V and $U = 3.2$ V in Fig. 4(a) and (b) correspondingly. The phase difference between these two states is shown in Fig. 4(c). As can be seen, at 5.2 GHz the phase difference between “0” and “1” states is precisely equal to 180°. Also, the magnitude levels of the reflection coefficient in both these states are almost equal as desired. The measured results are in good agreement with the simulated ones.

### B. Scattering Patterns in Reflected Beam Steering

Far-field scattering patterns were measured at 5.2 GHz for the normal incidence, i.e., the wave from the distant horn impinged the RIS from the direction of the normal to its aperture. Depending on precalculated 1 bit phase profiles, different scan angles of an anomalously reflected TE-polarized beam were obtained in the horizontal plane. For comparison, numerical simulations of the entire RIS illuminated with a normally incident plane wave were carried out.

The required 1 bit phase distribution for scan angle $\theta_{scf} = 30°$ is shown in Fig. 7(a). A projection of the simulated scattering pattern onto $XY$ plane, i.e., the 2-D map of the scattering pattern in dB versus coordinates $U = \cos(\phi) \sin(\theta)$, $V = \sin(\phi) \sin(\theta)$, is shown in Fig. 7(b). The corresponding measured pattern is shown in Fig. 7(c). The horizontal cuts of the simulated and measured patterns are shown in Fig. 7(d) in Cartesian coordinates versus angle $\theta$ in the horizontal plane.

### A. Reflection Coefficients

Magnitudes and phases of the complex reflection coefficient of an incident plane wave from the RIS were measured for various biasing voltage levels applied to the diodes ranging from 2.8 to 3.6 V with a step of 0.2 V. For each level, the same biasing voltage was applied to all diodes of the RIS (uniform phase distribution). Also, the case of a zero biasing voltage was considered in the measurements. The magnitudes and phases were measured indirectly through near-fields [40] in the range of 4.9–5.5 GHz with a step of 10 MHz. The required near-field maps were scanned within the area of 240 × 240 mm with a resolution of 10 mm. The scan region was chosen with smaller dimensions than ones of the aperture to reduce the effect of edge diffraction and make the obtained reflection coefficients closer to ones numerically calculated for an infinite 2-D-periodic structure. At each probe position, the magnitude and phase distributions of the $x$-component of the magnetic field were recorded. The same value in the absence of the RIS and in the case where the RIS was replaced with a copper plate of the same dimensions was also measured to determine a complex reflection coefficient as described, e.g., in [41] and [42]. The effect of parasitic signal reflections between the RIS and the horn antenna was reduced using the time gating procedure [43].

The magnitude and phase of the measured reflection coefficient are compared with the simulation results for $U = 0$ V and $U = 3.2$ V in Fig. 4(a) and (b) correspondingly. The phase difference between these two states is shown in Fig. 4(c). As can be seen, at 5.2 GHz the phase difference between “0” and “1” states is precisely equal to 180°. Also, the magnitude levels of the reflection coefficient in both these states are almost equal as desired. The measured results are in good agreement with the simulated ones.
As can be seen from the results, the scattering pattern except for the main beam exhibits a highly undesirable symmetric beam. This results in the splitting of the reflected power in two beams of almost the same level at $\theta = \pm 30^\circ$. The appearance of the second beam originates from binary phase quantization and is a common effect of any 1 bit RIS [16] visible when the reflected beam angle differs from the mirror one. The same data for scan angle $\theta_{ref} = 45^\circ$ are given in Fig. 7(e)–(h).

The simulated and measured maximum directivity as well as the simulated efficiency due to reflection losses in the RIS are given in Table I.

A slight difference between the simulated and measured results is explained by the sphericity of the phase front of the incident wave due to the relatively small distance between the source antenna and the RIS. Table II summarizes the features of our design in comparison to previously known approaches. It should be noted that there is a lack of data in the literature on the measured efficiency and directivity of RISs. Moreover, none of the RISs presented in the literature combines full 2-D remote control of phase using IR remote control with scaling capability.

### C. Spatial Distributions in Holographic Field Synthesis

In order to demonstrate the operation of the 1 bit phase coding in the regime of holographic field synthesis we aimed to create a certain magnitude distribution of magnetic field on the target plane $Z_{plane} = 700$ mm away from the RIS. The desirable shape of the distribution was the shape of...
the capital letter “I” with the sizes of 200 × 200 mm [see Fig. 7(j)].

Using the iterative adaptation algorithm, described in Section II-D, the phase distribution shown in Fig. 7(i) was computed. The simulated and measured H-field distributions on the target plane are shown in Fig. 7(k) and (l) correspondingly. The realized distribution had a blurred shape in comparison with the desirable one due to the finite aperture dimensions of the RIS which limited the spatial resolution of the holographic phase synthesis. However, the experimental and simulated distributions resembled shape “I” and looked visually similar to each other.

D. Reflected Signal Transmission Between Two Antennas

The aim of a RIS is to enhance the signal propagation between a BS and a user. To estimate the effect of the fabricated RIS in this practical scenario, we experimentally modeled a single-path wave propagation at 5.2 GHz between two horn antennas with a properly phase-encoded RIS acting as an intermediate reflector on the path. The transmission was characterized by comparing the levels of the signal transferred between the horns in the presence of the RIS and the absence of it. To satisfy the conditions of the far-field region of the RIS with respect to both horn antennas, in the measurements we used only one panel containing 10 × 10 patches. The experimental setup is schematically shown in Fig. 8(a).

It consisted of two linearly polarized broadband horn antennas placed at distances \( R_1 = 2.7 \) and \( R_2 = 4.2 \) m from the RIS. As in previous measurements, the horns created the main TE polarization with an electric field vector oriented along the y-axis. Both horns were connected to ports of VNA Rohde&Shwarz ZVB20 with calibrated 50 Ω coaxial cables. The signal level was proportional to the magnitude of the \( S_{12} \)-parameter measured in the range of 5.0–5.3 GHz.

The experimental investigation covered several scenarios with different incidence angles and reflection angles in the horizontal (XZ) plane realized with different positioning of the horns and RIS in the anechoic chamber. The following scenarios were compared:

1) RIS encoded to reflect a wave coming from a horn at angle \( \theta_{\text{inc}} = 0^\circ \) to a horn at angle \( \theta_{\text{ref}} = 45^\circ \);  
2) RIS encoded to reflect a wave coming from a horn at angle \( \theta_{\text{inc}} = -15^\circ \) to a horn at angle \( \theta_{\text{ref}} = 30^\circ \);  
3) mirror reflection with all patches set to “0” state;  
4) mirror reflection with all patches set to “1” state.  

The calculated 1 bit phase distributions for the first two (optimized transmission) scenarios are shown in Fig. 8(b) and (d), while the corresponding measured spectra of \( S_{21} \) magnitude are given in Fig. 8(c) and (e) with red lines. The obtained transmission coefficient levels were found to be in good comparison to theoretical predictions made using the radar range equation (see [44, Sec. 2.17]). The received power \( P_r \) at horn 2 can be expressed through the transmit power \( P_t \) applied to horn 1 as follows:

\[
P_r = P_t \frac{L_x L_y \cos(\theta_{\text{inc}})(1 - |S_{11}|^2)(1 - |S_{21}|^2)}{(4\pi)^3 R_1^2 R_2^2} G_t G_r G_{\text{RIS}}
\]  

(3)

where \( L_x = L_y = 300 \) mm are the dimensions of the RIS, \( \lambda = 57.7 \) mm—wavelength in free space, \( R_1 = 2.7 \) m and \( R_2 = 4.2 \) m—distances from horns 1 and 2 to the RIS, \( G_t = 15 \) dB—gain of the transmit horn, \( G_r = 15 \) dB—gain of the receive horn, \( G_{\text{RIS}} \)—gain of the RIS for the given incidence \( \theta_{\text{inc}} \) and reflection \( \theta_{\text{ref}} \) angles, \( |S_{11}| = -20 \) dB and \( |S_{21}| = -20 \) dB—return loss levels of the transmit and receive horns, respectively. It should be noted that this expression is valid only for the case, in which both transmit and receive horns are located in the far-field region of the RIS. Gain \( G_{\text{RIS}} \) was
estimated by using the magnitude of the reflection coefficient measurements (0.58 or −4.7 dB at 5.2 GHz) to find the power efficiency and the analytical estimation of the directivity value calculated from the measured scattering pattern. The directivity was estimated as $D = 20.1 \text{ dBi}$ in the first scenario and $D = 19.5 \text{ dBi}$ in the second one. The magnitude of the theoretical transmission coefficient $|S_{21}| = \sqrt{P_r/P_t}$ is given with green dashed lines in Fig. 8(c) and (e). As can be seen, the theoretical levels well coincide with the measured levels in the first two scenarios.

For comparison, the measured signal transmission levels are also shown in Fig. 8(c) and (e) for the third and fourth scenarios as well as for the case in which the RIS is absent. As follows from the comparison of the results, the presence of the properly phase-encoded RIS improves the signal transmission between the horns by 28.5 and 25.5 dB in the first and second scenarios, respectively. In the third and fourth scenarios, the RIS improves the signal only by less than 12 dB because of uniform phase distribution and non-optimal mirror reflection behavior.

### IV. SUMMARY AND CONCLUSION

A new approach to designing an optically controlled RIS was proposed and demonstrated. In contrast to previous RISs, structurally and functionally independent building blocks containing four identical patches with an embedded common microcontroller were used. Each block can operate alone or together with multiple similar blocks forming a reflective aperture that is scalable in its dimensions by changing the number of blocks. Each block can be distantly controlled via IR digital code reacting only to its own commands. The capability of full 2-D phase encoding was experimentally demonstrated in the Wi-Fi 5 GHz range in two cases: reflected beam steering in the horizontal plane and holographic field synthesis. Experimental results are in good comparison with analytical predictions and numerical simulations.

The results have shown the possibility of a distant and robust control of a scalable RIS via digital infrared code. Note that during the experiments the IR remote unit equipped with a low-power LED and was typically located a few meters from the RIS providing stable control either in an anechoic chamber or in a furnished lab room. The operational distance of the IR remote is possible to increase by using an IR laser diode or an IR repeater. Optical control is promising since it does not interfere with any radio channels but limited to use within a single room only. To improve the coverage the proposed scaling principle could be modified to employ other methods of wireless transfer of control signals to distributed controllers, such as via IoT or long-range Bluetooth protocols.

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