A 2D-programmable and Scalable RIS 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. An RIS assists a base station in optimizing the channel and maximizing its capacity by dynamically manipulating with reflected field. Typically, RISs are based on dynamically reconfigurable reflectarrays, i.e. two-dimensional arrays of passive patch antennas, individually switchable between two or more reflection phases. Different communication scenarios and environments require RISs to provide a different spatial resolution of reflected field patterns, which depends on the aperture dimensions and the number of patches. Here we demonstrate a 1-bit RIS for 5-GHz Wi-Fi band made by assembling together multiple independently operating 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 illuminating the entire RIS. Such distributed light-sensitive controllers grant the possibility of scaling the aperture by adding or removing blocks without re-designing any control circuitry. Moreover, in the proposed RIS a full 2D phase encoding capability is achieved along with a robust remote infrared control.

Index Terms—Reconfigurable Intelligent Surface, Patch Array, Reflectarray, Varactor Diode, Control with Light.

I. INTRODUCTION

Modern and prospective wireless communications systems face with 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 metasurfaces (MSs - electrically dense periodic structures of subwavelength scatterers) and reflectarrays (RAs - periodic arrays of passive individually tunable antennas with typically a half-wavelength spacing) used for that purpose are commonly called Reconfigurable Intelligent Surfaces (RISs). These structures are considered in the literature as powerful and relatively cheap solutions for improving microwave and millimeter-wave wireless channels [1], [2]. In outdoor environments, an 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, an 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 [1]. In both cases, one or several RISs are to be strategically placed in the environment with a predefined location of the signal source. Applications of RISs, their mathematical models used in wireless communications, as well as their practical realizations have been recently reviewed by several groups of authors [3–7].

In the microwave range, RISs can be implemented in two ways [7]: using MSs or RAs. MSs consist of deep subwavelength unit cells and have small periods compared to the wavelength, while RAs typically have the period of around \( \lambda/2 \), where \( \lambda \) is the operational 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 \( 70 - 80^\circ \) from the normal [10], the features of their unit cells are much finer than of RAs. For that reason, RAs better fit to the relevant technological constraints that are especially important in the millimeter-wave range [11] and, therefore, serve as RISs in most papers.

Usually RAs consist of passive metal patch antennas placed over a common ground plane. The patches are excited by the incident wave and scatter (reflect) the signal with a predefined phase delay or advance. To form a desirable reflected field pattern, the corresponding distribution of the reflection phase over the aperture is to be computed and encoded. This can be achieved by adjusting the resonant frequency of each patch according to the local value of the required reflection phase. The most important function of RISs is to reflect the impinging beam into a desirable direction. For reflection of a normally incident wave to angles of less than \( 70 - 80^\circ \) from the normal, it is sufficient to approximate a linear reflection phase profile within the aperture. Dynamical steering requires the resonant frequency of each patch to be gradually tunable. This tunability can be realized e.g. by biasing a varactor diode connected between the patch and the ground plane [12] or between two halves of the patch. However, gradual and dynamic variation of phase for a large amount of individually controlled patches requires complex and expensive electronic
Building on the back side of a multi-layer printed-circuit board (PCB). Including biasing lines and controllers were usually placed the DC and microwave parts of the RIS, the DC circuits to individually bias the diodes for setting phases of varactor diodes, a photodiode, a controller (C) and DC drive (D). One building block composed of four individually-controlled patches loaded wireless access point via infrared digital code. Inset shows the organisation of Fig. 1. Operational principle of the proposed RIS remotely controlled from a wireless access point.

Special DC circuits are to be embedded into the design of the 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 were usually placed on the back side of a multi-layer printed-circuit board (PCB). On the contrary, the patches were 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. However, in both cases RISs are to be re-designed each time when the aperture dimensions and a number of patches needs to be changed. For instance, scaling the aperture is necessary to obtain a desirable width of a reflected beam or certain spatial resolution of the reflected field distribution. 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 a remote control to every element of the distributed controllers. This approach allows one to encode the given reflection phase distribution without any physical wire connection for transferring biasing voltages or control signals providing a higher degree of freedom and helping strategically arrange RISs. Among other tunability methods for RISs, optical illumination appears the most attractive because it allows photonic functions to be reconfigured in a noncontact manner and with high switching speed \([32]\).

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 \([24], [34]–[36]\). Light-controllable RISs have been implemented in several ways as follows. In \([19]\) an RA operating around 4 GHz was composed of split patches loaded with varactor diodes in their gaps. The varactors grouped in 1D rows were commonly biased with voltages generated by PIN photodiodes operating in the photovoltaic mode in response to applied visible light from an edge row with 50 white light-emitting diodes (LEDs). By tuning the intensity of LEDs, the RIS switched between states with one or two reflected beams. In \([37]\) an RIS was composed of blocks each one containing 4 × 4 patches loaded with varactor diodes. The reflection phase of each block as a whole was determined by the intensity of light illuminating a meandered chain of 22 PIN photodiodes placed on the back panel producing the bias voltage for all 16 patches of the block simultaneously. Light from white LEDs was focused to illuminate individually 6 × 6 blocks to encode various 2D stepwise phase patterns at frequencies from 5.8 to 7.0 GHz. In \([31]\) the authors proposed an infrared-controlled RIS capable of 1-bit phase encoding for reconfigurable 1D beam steering and splitting at frequencies from 4.1 to 4.35 GHz. Compared with the visible-light control mechanism, the infrared-controlled approach can work in longer remote control distances and with higher energy efficiency. The RIS had a centralized control unit with several bias voltage combinations pre-coded in FPGA enabled when receiving coded signals from a remote infrared transmitting circuit. Recently, a light-controlled RIS operating at 6.375 GHz integrated with photoresistors and varactors has been proposed, which can independently obtain a controllable range of reflection phase modulation on each unit \([20]\). A full 2D phase encoding capability was first achieved in \([20]\) by integrating a photoresistors-based circuit on the back side of each unit cell and using the dynamically reconfigurable light field illumination created using a liquid-crystal display placed...
behind the RIS. This approach simplifies scaling the reflective aperture of the RIS due to independent analog-type biasing of identical unit cells all powered by the same DC voltage. However, it does not allow any distant light control that would be useful for realistic indoor applications.

In this work, we propose and demonstrate a 1-bit RIS with full 2D independent phase control of individual patches via IR light coming from a distant source. In this approach IR light is modulated with digital series control sequences and illuminates the whole aperture of the RIS, while the RIS is composed of identical building blocks powered by the same DC source. Each block contains four independently phase-switchable patches and equipped with one common microcontroller and one photodiode receiving the control sequences. Each block responds only to its own pre-coded address from a sequence and then extracts the information about the required phase states for its four patches and applies the corresponding combination of bias voltages. The operation of the proposed RIS is schematically shown in Figure 1. With this approach one can freely scale the RIS by changing the number of blocks without any complex wire control circuitry connecting them together and with no analog light-field sources. As a result, a distant and robust control via IR makes the proposed RIS advantageous for indoor communications.

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 $20 \times 20$ individually controllable patches with 2D 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 mm $\times$ 600 mm (approximately $10\lambda \times 10\lambda$).

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 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 inset of Figure 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 1.524 mm, relative permittivity of 3.38 and dielectric loss tangent of 0.0027. Each patch had a rectangular shape with dimensions $L$ and $W$ and a through via going through a hole in the ground plane and connecting the center of the patch to the DC circuit as shown in Figure 2. The vias were made to apply biasing voltages to the patches relative to the ground (GND).

The resonant frequency of the patch was determined by its length $L$ and a capacitance of varactor diodes connecting the patch with the ground plane at a distance $s/2$ from the center of the patch. To connect two diodes between the patch and the ground plane, two rectangular windows with dimensions $a \times b$ were made in the patch and two small contact plates connected to the ground with two additional vias were added. Each diode was soldered between the edge of its window and the corresponding contact plate. In this design, the ground plane and the contact plates were kept at a zero DC potential, while the patch was under a bias voltage formed by the DC circuit. The latter was positioned at two layers of FR4 substrate behind the ground plane. The FR4 part of the multilayer board had 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 prereg KB-6065 with a dielectric permittivity of 4.2 and dielectric loss tangent of 0.013.

In the literature, a DC biasing circuit is typically isolated from the microwave part of RISs by means of linear quarter-wavelength stubs, radial stubs, inductors or other RF chokes. Unlike in the previous works where a patch was loaded with only one diode and had an RF choke, here we used a symmetric combination of two diodes per patch. This approach removed the current induced on the central via for the normal incidence and oblique incidence of a TE-polarized wave isolating the DC circuit without any choke. This allowed us to save space on the board and reduce losses.

By varying distance $s$ between the windows with diodes one can adjust the effect of the tunable capacitance to the resonant frequency of the patch. This parameter helped us to obtain 1-bit switching between two states (“0” with a reflection phase of $-90^\circ$ and “1” with a reflection phase of $+90^\circ$). These two values of phase are convenient for operation of a 1-bit RIS as they correspond to equal detuning of the patch resonance from the central operational frequency and typically correspond to equal magnitudes of the reflection coefficient. With proper $s$ it was possible to provide the bias voltages corresponding to both phases within an available 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 $s$.

In order to numerically predict the bias voltages providing reflection phases of $\pm 90^\circ$ and correct 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. However, the patch dimensions were changed in order to fit to the cross-section of WR229 waveguide with cross-section dimensions of $58.17 \text{ mm} \times 29.08 \text{ mm}$. With patch dimensions $L = 30.5 \text{ mm}$ and $W = 19.6 \text{ mm}$ 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 Figure 2 were: $b = 5 \text{ mm}$, $a = 6 \text{ mm}$, $w = 0.5 \text{ mm}$, $l_1 = 1.7 \text{ mm}$, $d = 0.2 \text{ mm}$, $s = 13.9 \text{ mm}$, $2r = 0.4 \text{ mm}$. Three three-layers PCBs were manufactured with one 1.524-mm-thick and one 0.508-mm-thick Rogers 4003C substrate. The first PCB had two varactors (shown in Figure 3(a)) connected in the windows of the patch in the same way as discussed above for the unit cell. Two other PCBs had no varactors. In one of them the patch was disconnected from the contact plates in the windows, while in the other one short-circuiting metal strips connected the patch to the contact plates instead of the varactors.

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. Numerical simulations were made in CST Microwave Studio 2020. In simulations both varactors were represented by lumped elements with sizes 0.3 mm $\times$ 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 constant parasitic lead inductance $L_d$ and parasitic capacitance $C_{par}$ as shown in the inset in Figure 3(a). The measurements were carried out in the range of 3–5 GHz with a step of 0.5 MHz. In the measurements the waveguide section was connected to vector network analyzer (VNA) Agilent E8362C calibrated using a standard coaxial kit via coaxial-to-waveguide junction (see Figure 3(a)). The PCBs with short-circuiting metal strips and splits at the places of the varactor diodes were used to ensure the correspondence between the calculated and measured reflection coefficient spectra in the absence of the varactors and to avoid possible numerical errors. 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 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 Figure 3(b,c) for $U = 0 \text{ V}$ and 3.2 V (the bias voltages eventually taken for the experimental prototype of the RIS). The parameters were obtained by their parametric 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. Particularly, the following properties were to made the same: (1) the frequency at which the real part reaches its maximum (the resonance); (2) the maximum level for the real part of the impedance; (3) the slope of the imaginary part at the resonance. Note that for different $U$, all three criteria are to be achieved only by varying the capacitance $C_d$ and resistance $R_d$ of the diode. It was assumed that the parasitic inductance and capacitance of the diode were independent on $U$. The resulting normalized impedance curves are given in Figures 3(d,e). As can be seen from the results, good agreement between measured and simulated curves was achieved. The extracted parameters were the following: $L_d = 0.2 \text{ nH}$, $C_{par} = 30 \text{ fF}$. For $U = 0 \text{ V}$ the main parameters were $C_d = 2.1 \text{ pF}$ and $R_d = 7.5 \text{ Ohm}$, while for $U = 3.2 \text{ V}$, $C_d = 0.87 \text{ pF}$ and $R_d = 7.1 \text{ Ohm}$. Some difference from the equivalent series resistance of 4.5 Ohm given in the datasheet can be explained by the fact that the manufacturer’s data was provided for low frequencies.

Next, the extracted equivalent-circuit parameters of the varactor diode were used in optimization of the unit cell of the RIS at 5.2 GHz.

**C. Design and organization of RIS**

Based on the extracted parameters of the diode, the unit cell was optimized via numerical simulations. Varactors were modeled in the same way as in the previous subsection. The goal was to obtain “0” and “1” phase states at the central frequency of 5.2 GHz for the normal incidence (vertical polarization along y-axis). Using Frequency-Domain Solver of CST Microwave Studio, the structure depicted in Figure 2 was modeled with a Bloch-Floquet port in conjunction with unit-cell boundary conditions. The reflection coefficient was calculated in the range of 4.9–5.5 GHz. The resulting parameters (all in mm) read: $L = 16.6$, $W = 22.0$, $a = 2.0$, $b = 2.6$, $w = 0.5$, $l_1 = 0.7$, $d = 0.5$, $s = 11.6$, $2r = 0.5$. The magnitude and phase of the simulated reflection coefficient is shown with solid lines in Figures 4(a-b). As can be seen, the chosen parameters indeed provide that “0” and “1” states realized with $U = 0 \text{ V}$ and $U = 3.2 \text{ V}$, correspondingly, differ by approximately $180^\circ$ in the reflection phase. This is additionally confirmed by the plot of the phase difference for the two voltages vs. frequency (Figure 4(c)). The phase difference holds within the range of 150–210$^\circ$ (maximum deviation of phase difference of 30$^\circ$ from 180$^\circ$) at frequencies from 5.12 to 5.28 GHz. Note that this operational range covers
Fig. 3. To extraction of varactor diode’s parameters: manufactured waveguide termination with a patch and two varactor diodes (top) and measurement setup (bottom) with a waveguide section and VNA (a); measured and numerically calculated reflection coefficient magnitude (b) and phase (c); measured and numerically calculated normalized real (d) and imaginary (e) part of the normalized impedance of the waveguide load. Inset in (a) shown an equivalent circuit used for varactor diodes.

Fig. 4. Simulated (solid curves) and measured (dashed curves) reflection characteristics of the RIS with all patches in "0" state (U = 0 V) and in "1" state (U = 3.2 V): reflection coefficient magnitude (a) and phase (b) for a normally incident vertically-polarized plane wave; phase shift between two reflection states (c).

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, as shown in the simulations, was due to dissipation losses mostly associated with the varactors.

Four identical unit cells were combined in one building block of the scalable RIS. The idea behind this combination is that the block has its own control unit that allows it receiving control sequences via infrared light regardless its position within the RIS aperture. In frames of a single block, an array of 2×2 patches was placed on the top layer of the multilayer PCB (on a grounded Rogers substrate), while the biasing circuit independently forming four control voltages for four pairs of varactor diodes was positioned on FR4 substrates behind the ground plane. The microwave and DC circuits had a common ground plane as an internal layer between Rogers and FR4 substrates kept under a zero potential.

The biasing circuit contained an IR receiver, a microcontroller, and a driver circuit all mounted on the opposite side of the multilayer PCB. The PCB of the block had a circular through hole in the center between 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.

The IR receiver was based on TSOP34338 optoelectronic module that included an optical IR filter, IR photodiode (PD), a band-pass filter (BPF), amplifier with an automatic gain control (AGC), demodulator and a digital output stage for interfacing with the microcontroller. The biasing circuit used ATTiny441 microcontroller with a small footprint and low power consumption. The board also included a hardware-based serial port making interfacing with a photoreceiver simpler. CodeVision AVR Integrated development environment (IDE) was used for microcontroller software development in C programming language.

The bias voltages were applied to the central vias of four patches connecting each patch to a field-effect transistor (FET) based driver circuit. In turn, the DC circuits controlled the reflection state of all four diode pairs by switching between two different bias voltages. The voltage combinations were prescribed by the common controller receiving the IR control signal from the IR remote unit distantly placed in the same
the packet contained phase settings for four patches of the array to all-zeros or all-ones state. The second byte of blocks simultaneously. This significantly speeds up the setting to control 100 blocks (i.e. 400 patches). The zero address unique addresses (from 0 to 127) which was more than enough to zero (an address marker). Thus, this protocol provided 128 addresses.

The most significant bit (MSB) of this byte was always set to form a control packet. First byte was the address field. The carrier frequency was 38 kHz. The IR transmitter was controlled from a computer via the USB interface. The schematic organization of the IR remote is shown in Figure 5(a), while a photograph of its prototype is shown in Figure 5(b). The phase control protocol used two bytes in Figure 5(c). The phase control protocol used two bytes.

4. Remote control and IR light delivery

Each block receives IR light illuminating the entire aperture of the RIS and extracts its own control commands from the digital code with the information about the requested reflection states to be set for its four patches. Despite IR light propagation can be limited due to outdoor conditions, this method in conjunction with digital modulation for coding is beneficial for indoor applications. In the developed RIS prototype, each block had its unique address. An IR transmitter at a remote side sent commands to all blocks simultaneously followed after the corresponding addresses. As a controller of a particular block recognizes its address, it sets phases to four its patches according to the received commands by forming a set of biasing voltages using the DC drivers circuit.

A near-infrared (940 nm) LED was used as a transmitter of the IR remote. The IR communication channel used amplitude modulation and control commands produced a series code sequence. The carrier frequency was 38 kHz. The IR transmitter was controlled from a computer via the USB interface. The schematic organization of the IR remote is shown in Figure 5(a), while a photograph of its prototype is shown in Figure 5(c). The phase control protocol used two bytes to form a control packet. First byte was the address field. The most significant bit (MSB) of this byte was always set to zero (an address marker). Thus, this protocol provided 128 unique addresses (from 0 to 127) which was more than enough to control 100 blocks (i.e. 400 patches). The zero address was a broadcast address used to send phase settings to all blocks simultaneously. This significantly speeds up the setting of the array to all-zeros or all-ones state. The second byte of the packet contained phase settings for four patches of the particular block. The most significant bit of this byte was always set to one (a data marker). The phase states were encoded using bits 0 to 3 of the second byte. The presence of address (MSB set to 0) and data (MSB set to 1) markers allowed recovery in case of possible synchronization errors (due to byte loss or corruption).

Since the two required voltage levels corresponding to "0" and "1" states may require tuning, a special driver circuit was developed to equip each block of the RIS. It allowed changing the two voltage levels simply by changing the voltages of two external power supplies. The third 5V power supply was required for the microcontroller, IR receivers and additional light-emitting diodes (LEDs) used for indication of phase states. The driver consisted of a single field-effect transistor (FET), two Shottky diodes and three resistors. The FET worked as a switch that can be in either open (low resistance) or closed (high resistance) state. Depending on the FET state either one or the other power supply was connected to the driver’s output via the corresponding resistor and Shottky diode.

A graphical user interface (GUI) program was developed to encode the phase over the RIS. Open source Lazarus IDE and FreePascal compiler were used. The program allowed to set the phases of individual patches manually or load them all from a file. It also had a test mode that allowed easy testing of all elements of the array.

5. Near- and far-field phase coding

The developed RIS belongs to the class of 1-bit reflectarrays with full 2D phase coding capability. To create a desirable spatial distribution of the reflected field, the corresponding phases of individual patches are to be pre-calculated and encoded. The calculation of phases was realized in frames of the same program as one controlled the IR transmitter for sending commands to the RIS.

The following adaptive algorithm was realized to calculate the phases based on the specified distribution of the electric 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 on 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 model. The model took into account the excitation field (created by either a plane wave or a point source), the distance from each RIS element to the observation point, the phase of the array element and its individual scattering pattern. All patches were assumed to have a Huygens-like scattering pattern given by the formula: 

\[ F(\theta) = \cos^2\left(\frac{\theta}{2}\right) \]

with \( \theta \) being a polar angle measured from the normal to the aperture. If the target area is a plane parallel...
to the aperture, the reflected field $E_{k}^{\text{RIS}}$ created by the RIS at point $k$ was calculated as

$$E_{k}^{\text{RIS}} = \sum_{m=1}^{N_x N_y} E_{\text{inc}}^{m} e^{j \Phi_{\text{inc}}^{m}} |R_{m}| e^{-j \varphi_{m}} \frac{1}{2} \left(1 + \frac{z_{\text{plane}}}{r_{mk}} \right),$$

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 $m$, $\varphi_{m}$ — reflection phase of patch $m$ ($-\pi/2$ for state “0” and $+\pi/2$ for state “1”), $z_{\text{plane}}$ — distance between patch $m$ and point $k$, $r_{mk}$ — distance from the target plane to the aperture of the RIS, $N_x, N_y$ - number of patches along $x$- and $y$-axis respectively. In the case when the source is a plane wave, the phase of the incident field was found as:

$$\Phi_{\text{inc}}^{m} = k_0 \left[ x_m \cos(\phi_{\text{inc}}) \sin(\theta_{\text{inc}}) + y_m \sin(\phi_{\text{inc}}) \sin(\theta_{\text{inc}}) \right],$$

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 ones of the aperture. In that case, the process can be applied to holographic field synthesis. In the following section, we consider both types of adaptation made with the experimental RIS.

### III. Experimental results

Measurements of reflected field distributions were made using a 3-axis field scanner installed in an anechoic chamber (7 m × 5 m × 3 m) and VNA Rohde&Schwarz ZVB20. The first port of the VNA was connected to the source antenna with a vertical linear polarization ($E_y$) while the second one - to a field probe. The experimental setup for the case of beam steering is shown in Fig. 6(a) with the source being a linearly polarized TEM-horn antenna at the distance of 4 meters from the RIS. In this case, for obtaining a scattering pattern, first a complex magnitude of magnetic field was measured over the plane parallel to the RIS at the distance of 10 mm from the patches in presence and absence of RIS. Then, Fourier transform was applied to calculate the far-field scattering pattern from the near-field map using standard methods [39]. In the case of holographic field synthesis, the field was measured directly on the target plane. In both cases, SX-R 3-1 H-field Probe mounted to an arm of the scanner and

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**Fig. 5.** Organization of the IR-controllable RIS block and its fabricated prototype: schematic of the IR remote (a) and schematic of the RIS building block with four individually controlled patches (b); prototype of the IR remote circuit with an LED transmitter (c) and prototype of the RIS block shown from the back side (DC biasing circuit elements are shown) (d) and the front side (four patches with a through hole are shown) (e).
mechanically manipulated in parallel to $XY$-plane was used. The complex magnitude of the measured field was extracted from $S_{12}$ values stored by the VNA.

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 $600 \text{ mm} \times 600 \text{ mm}$ fixed on the holder is shown in Figure 6(b) from the front side and in Figure 6(c) from the back side. DC supply Rohde&Shwarz HMP2030 was used to power the building blocks providing the required voltage levels.

Experimental investigation of the RIS in different regimes included the following tasks:

1) Measurements of complex reflection coefficients from the RIS with all patches being set to "0" or "1" state;
2) Indirect determination of far-field scattering patterns in the regime of reflected beam steering by calculating from measured near fields for different phase coding profiles corresponding to different scan angles of the reflected beam;
3) Direct measurements of intermediate-field distributions in the regime of holographic field synthesis on the target plane positioned at distance $Z_{\text{plane}} = 700 \text{ mm}$ away from the RIS.
4) Measurements of the signal transmission between two antennas in the presence and absence the RIS.

The results of the corresponding tasks are presented and discussed below in the corresponding subsections.

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 $2.8 \text{ V}$ to $3.6 \text{ V}$ with a step of $0.2 \text{ V}$. 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 [39] in the range of $4.9-5.5 \text{ GHz}$ with a step of $10 \text{ MHz}$. The required near-field maps were scanned within the area of $240 \text{ mm} \times 240 \text{ mm}$ with a resolution of $10 \text{ mm}$. The indirectly calculated far-field complex magnitudes were obtained at the post-processing step as discussed in [39]. 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 2D-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 [40], [41]. The effect of parasitic signal reflections between the RIS and the horn antenna was reduced using the time gating procedure [42].

The magnitude and phase of the measured reflection coefficient are compared with the simulation results for $U = 0 \text{ V}$ and $U = 3.2 \text{ V}$ in Figures 6(a) and (b) correspondingly. The phase difference between these two states is shown in Figure 6(c). As can be seen, at $5.2 \text{ GHz}$ the phase difference between "0" and "1" states is precisely equal to $180^\circ$. 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 $5.2 \text{ GHz}$ for the normal incidence, i.e. the wave from the distant horn impinging the RIS from the direction of the normal to its aperture. Depending on pre-calculated 1-bit phase profiles, different scan angles of an anomalously reflected TE-polarized beam were obtained in the horizontal plane. The uniform phase distribution with all patches set to "0" state and the corresponding measured scattering pattern in the horizontal plane with the main reflected beam in the normal direction are given in Figures 6(d-e). The horizontal cut of the measured scattering pattern is shown in Cartesian coordinates versus angle $\theta$ in the horizontal plane with the blue dashed line. A projection of the scattering pattern onto $XY$ plane (i.e. the 2D map of the scattering pattern in dB vs. coordinates $u = \cos(\phi) \sin(\theta)$, $v = \sin(\phi) \sin(\theta))$ is shown in Figure 6(f). The same data for scan angle $\theta_{\text{ref}} = -45^\circ$ are given in Figure 6(g-i). The measured maximum directivity was $D = 31 \text{ dBi}$ for $\theta_{\text{ref}} = 0^\circ$ and $D = 24.1 \text{ dBi}$ for $\theta_{\text{ref}} = -45^\circ$.

The scattering pattern corresponding to $\theta_{\text{ref}} = -45^\circ$ exhibits a high phase quantization lobe. As a results, the reflected power is equally split between two beams at $\theta_{\text{ref}} = \pm 45^\circ$, which is an effect of any 1-bit phase-coded RISs [14] typically observed when a reflected beam scan angle differs from the mirror-reflected one.

For comparison, numerical simulations of the entire RIS illuminated with a normally incident plane wave were carried out using Time Domain Solver of CST Microwave Studio 2020. The simulated scattering patterns in the horizontal plane shown in Figures 6(c-h) with solid red lines are in good agreement with the measured curves.

C. Spatial distributions in holographic field synthesis

In order to demonstrate the capability 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 $Z_{\text{plane}} = 400 \text{ mm}$ away from the RIS. The desirable shape of the distribution was the shape of the capital letter "I" with the sizes of $200 \text{ mm} \times 200 \text{ mm}$ (see Figure 6(k)).

Using the iterative adaptation algorithm, described in Section 11-E the phase distribution shown in Figure 6(j) was computed. The simulated and measured H-field distributions...
on the target plane are shown in Figure 6 (l) and (m) correspondingly. The realized distribution had a blurred shape in comparison with the desirable one due to 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 an RIS is to enhance the signal propagation between a base station 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 characterised by comparing the levels of the signal transmitted between the horns in the presence of the RIS and in 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 Figure 7a. 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 $y$-axis. Both horns were connected to ports of VNA Rohde&Schwarz ZVB20 with calibrated 50-Ohm 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 included 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 Figure 7(b,d), while the corresponding measured spectra of $S_{21}$ magnitude are given in Figure 7(c,e) with red lines. The obtained transmission coefficient levels were found to be in a good comparison to theoretical predictions made using radar range equation (see Section 2.17 in [43]). The received power $P_r$ at horn 2 can be expressed through the transmit power $P_t$ applied to horn 1 as follows:

$$P_r = P_t \frac{\lambda^2 L_x L_y \cos(\theta_{\text{inc}})(1-|S_{11}|^2)(1-|S_{22}|^2)}{(4\pi)^2 R_1^2 R_2^2} G_t G_r G_{\text{RIS}},$$

(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 $S_{22} = -20$ 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 $D = 20.1$ dBi in the first scenario and $D = 19.5$ dB 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 Figures 7(c,e). As can be seen, the theoretical levels well coincide with the measured levels.

Fig. 6. Experimental investigation of the proposed RIS at 5.2 GHz: (a) setup for measurements of the reflected field pattern; (b) fabricated RIS composed of $N_x \times N_y = 20 \times 20$ patch elements shown from the front side (inset shows one building block with $2 \times 2$ patches); (c) back side of the RIS with electronic circuits and wire interconnections for powering building blocks from a common DC supply (inset shows one building block with $2 \times 2$ patches). Phase distributions with 1-bit coding and corresponding measured far-field patterns for the normal incidence ($\theta_{\text{inc}} = 0^\circ$); (d-f) beam steering to scan angle: $\theta_{\text{ref}} = 0^\circ$; and $\theta_{\text{ref}} = 45^\circ$ (g-i). Phase distribution (j), desirable magnitude field distribution on the target plane $Z_{\text{plane}} = 700$ mm away from the aperture, simulated (l) and measured (m) field distribution on the same plane.
in the first two scenarios.

For a comparison the measured signal transmission levels are also shown in Figure 7(c,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 scenario, respectively. In the third and fourth scenario the RIS improves the signal only by less than 12 dB because of a uniform phase distribution and non-optimal mirror reflection behavior.

IV. CONCLUSION

A new approach to design an optically-controlled RIS was proposed and demonstrated. In contrast to previous RISs, identical building blocks containing four identical patches with an embedded individual microcontroller were used. Each block can operate alone or together with multiple similar blocks forming a reflective aperture which is scalable in its dimensions by changing the amount of blocks. Each block can be distantly controlled via IR digital code reacting only to its own commands. The capability of full 2D 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. Therefore, the possibility of a distant and robust control of a scalable RIS via digital infrared code was shown to be promising for indoor wireless communications.

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