A Single-Switch-per-Bit Topology for mmWave and THz Reconfigurable Reflective Surfaces

Panagiotis C. Theofanopoulos, *Associate Member, IEEE* and Georgios C. Trichopoulos, *Member, IEEE*

Abstract— We present novel multi-bit unit-cell topologies for reconfigurable reflective surfaces –RRSs– (e.g., reflectarray antennas) with compact designs for millimeter-wave and terahertz (mmWave/THz) applications. Typically, mmWave/THz RRSs utilize one or multiple single-pole-single-throw (SPST) switches leading to single- or dual-bit modulated surfaces. These surfaces utilize the switches to manipulate the phase of the impinging waves, beamforming the reflected waves to the desired direction. As such, RRSs are leveraged either for imaging or wireless communication applications, which typically require the formation of a single beam (no grating lobes) and high gains. The gain and quantization lobe levels of an RRS is strictly related to the number of phase bits utilized in the unit-cell. Explicitly, more phase bits lead to lower quantization errors and better maximum gain/aperture efficiency. However, increasing the number of phase bits requires more SPST switches integrated within the unit-cell, leading to complex designs with high RF losses. Herein, we present, for the first time, RRSs with up to 4 phase quantization bits (16 states) that maintain one switch-per-bit topology thus retaining a low-complexity design. The proposed RRSs is presented alongside a series of analytical and full-wave simulated results.

Index Terms—Reconfigurable reflective surfaces, reflectarray antennas, millimeter-wave, terahertz

I. INTRODUCTION

Beam reconfiguration is the cornerstone of modern antenna systems and the enabler for a plethora of applications including 5G and beyond communications and security/biomedical imaging [1]–[4]. These applications typically require large apertures to form narrow steerable beams that increase signal strength and spatial selectivity for wireless communication applications, and spatial resolution for imaging systems. Typically, beam steering antenna systems are classified into four categories: 1) analog phased arrays, 2) digital phased arrays, 3) leaky-wave antennas (LWAs), and 4) reconfigurable transmission or reflective surfaces (reflectarray antennas, metasurfaces, etc.). In analog phased arrays, beam reconfiguration is achieved through a corporate feeding mechanism (analog beamforming) [5][6]. On the other hand, in digital beamforming systems, including hybrid and fully-digital

topologies (e.g., multiple input multiple output–MIMO arrays), beam reconfiguration is carried out in the post-processing domain (digital beamforming) [7][8]. Both analog and digital phased arrays systems have been widely utilized in microwave frequencies (up to 30 GHz) since they offer excellent beam steering performance and many incorporated components are available off-the-shelf (mixers, amplifiers, digital-to-analog converters, switches, filters, phase shifters, etc.). However, in mmWave/THz bands (30 GHz–1 THz), these components should be custom designed and monolithically integrated into the system, increasing fabrication cost [9], design complexity, and losses, leading to inefficient designs in terms of DC power consumption and RF losses for large apertures with hundreds or even thousands of integrated antennas/channels [10]. Alternatively, reconfigurable LWAs allow beamforming with low RF-front-end complexity and low-profile [11][12]. Reconfigurable LWAs carry out beamforming by manipulating their surface impedance with the use of cascaded switches. However, in such topologies, waves travel through cascaded lossy switches, thus the aperture size is limited by the poor efficiency, especially in mmWave/THz frequencies.

Fig.1 (a) In RRSs, a tunable switch is integrated at each antenna to modulate the wavefront of the reflected signals. The switching mechanism can be implemented using either (b) series or (c) shunt configurations.

Conversely to the aforementioned systems, RRSs (reflectarray antennas, metasurfaces, etc.) achieve beam

reconfiguration by modulating the surface electromagnetic properties of the structure (e.g., surface impedance/reflection phase), thus steering the incident waves to the desired direction [13]–[22]. Large aperture RRSs consisting of hundreds or even thousands of tunable antenna elements, retain a low profile, exhibit reduced RF-front-end complexity, and fewer losses compared to phased arrays (analog or digital) [10]. Specifically, these surfaces consist of antennas with incorporated switches, which are electrically controlled by a biasing circuitry (e.g., PIN diodes, varactors, vanadium dioxide, graphene, etc.). Moreover, the illumination source is typically a single/multi-channel feeding aperture placed outside the surface (e.g., horn antenna [13]), thus avoiding the use of a complex feeding network with multiple integrated RF components (e.g., couplers, phase shifters, power dividers/combiners, etc.). As such, by manipulating the state of the switches we can modulate the phase of the reflected waves and steer the beams in the desired directions, as shown in Fig. 1a. Besides the traditional use in reflectarray antennas, RRSs can be leveraged in the implementation of intelligent reflective surfaces (IRSs) to extend coverage and improve channel capacity in wireless communication systems [18][19]. IRSs are strategically placed in key locations and redirect the –otherwise wasted– waves from the base station to the users, increasing throughput and coverage range. As such, IRSs do not have a fixed source but rather incident waves impinging from arbitrary directions.

Fig.2 (a) The equivalent network of the switch terminated multi-port antennas. The received signal is distributed between the SPST switches (S_i) through the respective delay lines (l_i). (b) The 2-bit shunt unit-cell topology and (c)–(e) the 2-bit shunt example, in which each of the SPST switches provides two states with a phase difference of 90° , leading to the 4-state constellation diagram in (e).

RRSs typically feature single-bit phase modulation on the incident waves (e.g., [13]), thus compromising beamforming performance (aperture efficiency) due to quantization errors [20]. As such, SPST switches are integrated on the antenna elements, and by manipulating their states (e.g., forward-

ON/reverse-OFF bias), the phase of the impinging waves is modulated leading to electronic beamforming. To steer the waves toward the desired direction, a unique phase profile is required on the surface, and the reflection phases of each antenna element vary continuously within the $[-\pi, \pi]$ region. However, in realized surfaces, these continuous values are quantized based on the available phase states due to hardware limitations (e.g., for 1-bit—the unit-cell’s phase can only take the values π and 0). Thus, the errors caused during the phase quantization process, lead to reduced aperture efficiency, coarse beam steering, and higher side lobe level (SLL). These drawbacks can be alleviated by using multi-bit configurations with finer phase quantization.

Such configurations are realized either by integrating multi-state tunable switches (e.g., varactors) on the antennas [14][15], or by incorporating multiple SPST switches (e.g., PIN diodes) in the unit-cell design [16]–[22]. The first approach is currently impractical since the availability of low-loss tunable switches (e.g., varactors) is limited in the microwave region (up to 30 GHz) and the use of mmWave/THz tunable materials (e.g., liquid crystals), requires high variable voltage, increasing the biasing network complexity [23]. Thus, configurations that integrate multiple SPST switches are preferred, including solid-state switches (e.g., CMOS, GaAs, InP, etc.) or tunable 2D materials (e.g., graphene, vanadium-dioxide, etc.) [24]–[31].

The phase modulation mechanism of RRSs that use SPST switches can be classified into series and shunt topologies, as shown in Fig. 1b–c. In the series topology, the switches are separated by delay lines, thus the reflected signal’s phase is controlled by changing the effective length of the transmission line, controlling the phase of the reflected waves [16]. However, in the series configuration, the number of switches is $2^B - 1$ (B is the number of bits) and increases exponentially with the number of bits, leading to amplitude modulation, insertion losses, and complex biasing networks. On the other hand, in a shunt topology, the received signal is equally split into parallel branches with different delay lines, as illustrated in Fig. 1c. As such, the impinging waves on the antenna are equally split into branches with different delay lines. Then the reflected waves from the branches constructively interfere forming the various phase states of the unit-cell [17].

Particularly, in the shunt topology, the antenna is integrated with a splitter/combiner forming a multi-port network that distributes the received signals to the parallel branches, as illustrated in Fig. 2a. In the case of 2-bit modulation, two SPST-switch-terminated branches are needed, as depicted in Fig. 2b. Each of the branches acts as a single-bit modulator, though, if the branches have an electrical length difference of 90° (round trip), the combined signals on the antenna exhibit 4 different phase states depending on the state of the respective switches (ON/OFF), as illustrated in the constellation diagrams of Fig. 2c–e. In this manner, the shunt topology retains a single switch-per-bit topology, thus keeping the biasing network complexity low, and mitigating amplitude modulation since the signals can be equally split between the identical switches. The main drawback of this configuration is the need for the bulky multi-port splitter/combiner that distributes the signals into the

multiple branches. A few attempts to overcome this problem are found in the literature [17][18], though to the authors' knowledge, none of the existing approaches achieves a single switch-per-bit topology for multi-bit reconfigurability.

The scope of this work is to implement high-efficiency multi-bit RRSs while retaining the single switch-per-bit constraint for minimum hardware complexity. Namely, we incorporate the bulky multi-port splitter/combiner into the antenna, thus alleviating the real-estate problems arising in the use of the shunt configuration. In our designs, we use patch antennas, commonly used in mmWave/THz RRSs, and achieve up to 4-bits of phase quantization, with a single switch-per-bit topology.

The rest of this paper is organized as follows: Section II discusses the proposed antenna-splitter/combiner design based on the cavity model of patch antennas. In section III, we present the multi-bit unit-cell designs. Then, in section IV, we show full-wave results of linear RRSs accompanied by a discussion on the bandwidth constraints. Finally, in section V, we elaborate on the possible applications and extensions of the proposed RRSs.

Fig.3 The patch antenna as a power splitter: (a) the edge fields of the patch antenna TM_{001} cavity mode, (b) the multiple feed design to equally distribute the imping waves between the ports, and (c) the full-wave model of (b).

II. THE ANTENNA AS A POWER DIVIDER/COMBINER

One of the obstacles toward implementing multi-bit RRSs using the shunt topology (Fig. 1c) is the integration of the power splitter/combiner in the limited space of the unit-cell. To overcome this bottleneck, we incorporate the splitter/combiner within the patch antenna, as illustrated in Fig. 3. As such, we assume that the patch antenna is a resonating cavity (TM_{001} mode, [32]) and incorporate 4 symmetric feeding lines from the non-radiating edges. Each of these lines feeds the same cavity mode (TM_{001}), thus the fields from the imping waves are coupled to all 4 ports. The distance (y_0) of the feeding points from the radiating edge impacts the impedance matching, as explained in [32], and is optimized to achieve equal power splitting between the ports.

The antenna is fed from the non-radiating edges, instead of inset feeding, to reduce the mutual coupling between the ports as demonstrated in Fig. 4. Furthermore, the coupling between the ports is pronounced as the distance between the inset microstrip feeding lines decreases. Additionally, the inset feeding alters the current flow on the patch antennas (due to the

inset cuts), reducing the radiation efficiency, especially in the case of multiple inset feeds.

To evaluate the proposed topology, we model the antenna-splitter/combiner configuration, shown in Fig. 3c, using an infinite periodic array of $\lambda_0/2$ spacing. As such, we assign a Floquet port on the upper boundary of the unit-cell and lumped ports at the edges of the transmission lines and calculate the signal coupled from the Floquet-port (free-space- FS) to the antenna ports. The substrate used is silicon ($\epsilon_r=11.9$ and $h=20 \mu m$) and the design frequency is 275 GHz, as for the rest of this paper. All the design parameters are given in table I.

Fig.4 Examples of dual-fed patch antennas: (a) inset feeding and (b) edge-feeding. (c) The coupling between the ports (substrate height is $20 \mu m$, $\epsilon_r=11.9$, and $D=90 \mu m$).

Fig.5 The full-wave simulated scattering parameters of the multi-port antenna configurations of Fig. 3. With FS the free space (Floquet port) is denoted.

When a single port is used, all the incident power couples to that antenna port from free space, as shown in Fig. 5a. As the number of ports increases, the incoming wave is equally split at the respective ports (Fig. 5b-c). Therefore, using the proposed design we achieve equal power splitting between the antenna ports in the limited space of the unit-cell, without using any bulky components (e.g., external power splitters). Moreover,

the phase difference between ports 1 and 4, and, 2 and 3 is 180° since the mode fields have opposite directions (Fig. 3a). Finally, the isolation between the different ports is always 3dB smaller than the power splitting.

TABLE I
UNIT-CELL DESIGN PARAMETERS

Variable	Value (μm)	Variable	Value (μm)
W	215	y_o (1-bit)	50
L	147	y_o (2-bit)	57.5
h	20	y_o (3-bit)	62.5
w_{msl}	5	y_o (4-bit)	65

Fig.6 The schematic illustration of a linear RRS under plane-wave illumination.

III. MULTI-BIT UNIT-CELL DESIGN

In this section, we present the mathematical approach used to study the performance of the proposed RRSs, alongside the single- and multi-bit designs of the reconfigurable unit-cells.

A. Radiation Pattern of Linear RRSs

Typically, RRSs are comprised of a periodic arrangement of elements (unit-cells). To redirect the impinging waves to the desired direction, the reflection phases of the unit-cells need to be modulated. To derive the necessary phase modulation for each unit-cell, we use array theory [32]. As such, let us assume that a linear RRS is illuminated by an impinging wave (Fig. 6) from an incident angle (θ_{imp}) that results in phase distribution $\theta_{illum}(i)$, for $i=1$ to N , where N the total number of unit-cells. To redirect the reflected wave at an angle θ_{des} the required phased profile of the RRS is given by

$$\theta_{ele}(i) = \frac{2\pi}{\lambda_0} d(i-1) \sin(\theta_{des}) - \theta_{illum}(i) \quad (1)$$

where d is the element spacing and λ_0 is the free-space wavelength. In the case of an impinging plane-wave, the illumination phase profile is given by

$$\theta_{illum}(i) = \frac{2\pi}{\lambda_0} d(i-1) \tan(\theta_{imp}) \quad (2)$$

The phase profile $\theta_{ele}(i)$, calculated in (1), consists of continuous values within the $[-\pi, \pi]$ region, however, a limited number of phase bits are utilized at the RRS unit-cell, thus the $\theta_{ele}(i)$ profile is quantized based on the available bits. After the quantized phase profile $\theta_{ele,q}(i)$ is calculated, the far-field ($\varphi=0^\circ$) RCS pattern of the linear RRS is given by

$$E(\theta) = E_{ele}(\theta) \sum_{i=1}^N e^{j\theta_{ele,q}(i)} e^{jkd(i-1)\sin(\theta)} A_{illum}(i) e^{j\theta_{illum}(i)} \quad (3)$$

where E_{ele} is the unit-cell's far-field pattern, N is the number of unit-cells across the linear RRS, $A_{illum}(i)$ is the magnitude of the illumination beam at the i^{th} unit-cell $-A_{illum}(i)$ is constant for a source at the far-field of the RRS- and θ is the elevation angle measured from boresight.

Fig.7 (a) The single-bit unit-cell, (b) the integrated ideal planar CPW switch with the incorporated delay line, (c) the reflection phases of the unit-cell for various delay lengths at 275 GHz, and (d) the constellation diagram of the proposed single-bit unit-cell at 275 GHz for $L_{branch}=25 \mu\text{m}$.

B. Single-Bit Unit-Cell Design

In this subsection, we present the single-bit unit-cell topology that is used as the baseline for the multi-bit configurations. As such, the unit-cell is comprised of a patch antenna with an integrated ideal switch connected at the non-radiating edge, as illustrated in Fig. 7a. In an implementation scenario, the integrated mmWave/THz SPST switches could be devised using solid-state technologies (e.g., GaAs) [29] and/or tunable 2D materials, including graphene, molybdenum disulfide, vanadium dioxide, etc. [27]. Typically, these sub-mmWave switches are implemented in a planar shunt configuration, utilizing coplanar waveguides (CPW), striplines, or slotlines to acquire optimum RF performance [27].

For this case study, we use an ideal CPW shunt switch consisting of a perfect electric conductor (PEC), as shown in Fig. 7b. As such, when the PEC is shorting the signal and ground conductors of the CPW, the reflection coefficient is $\Gamma=-1$ (S.C.); on the contrary, in the absence of the PEC, the open-ended CPW is approximated by an open circuit [33], thus $\Gamma=1$ (O.C.). Instead of the PEC, we could assume any 2D material

that exhibits a tunable sheet resistance with a high ON/OFF ratio (e.g., vanadium dioxide) or a high-isolation SPST shunt switch (e.g., [28]). To couple the signals from the coplanar-waveguide switch to the microstrip line, a via-less CPW to microstrip transition is implemented [34] since mmWave/THz vias exhibit high parasitics limiting the performance. Moreover, fabricating a via through a thick silicon substrate is a challenging task, mainly due to the associated dry etching and metal filling process carried out through advanced nanofabrication techniques [35].

As mentioned in section I, to achieve multi-bit reconfiguration, multiple SPST switches need to be integrated with the antenna, where each requires an individual delay line with respect to the feeding point. For that purpose, a microstrip delay line is incorporated in the switch design, as shown in Fig. 7b. The goal of this delay line is to shift the reference of the reflected signals as discussed in section I (Fig. 2c-d). As shown in Fig. 7c, the unit-cell reflection coefficient between the two states (O.C./S.C.), acquired using full-wave simulations [36], remains close to 180° for various delay lengths. The unit-cell reflection phase is -54° when the switch is S.C. and 117° for the O.C., leading to a phase difference of 171° , close to the ideal 180° , as depicted in the constellation diagram of Fig. 7d. We define a performance metric for the single-bit unit-cell as the ratio of the achieved phase difference (171°) over the ideal single-bit phase difference (180°), leading to a 95% ideality. The phase deviation from the ideal 180° is mainly attributed to the parasitic capacitances of the open-ended CPW [33].

Fig.8 (a) The 2-bit unit-cell design with two integrated SPST switches, (b)/(c)/(d) the full-wave simulated 2/3/4-bit unit-cell constellation diagrams at 275 GHz, respectively.

C. N-bit Unit-Cell Design

In this sub-section, we present the multi-bit unit-cell designs that incorporate the ideal switch with the delay line into the

multi-port antenna-splitter/combiner of section II.

As such, the 2-bit unit-cell consists of two CPW based SPST switches integrated at each port of the dual-feed antenna as depicted in Fig.8a. Varying the states of the two switches between S.C. and O.C., we acquire four different phase states, hence 2 bits. The distribution of the four phase states alongside the unit-circle depends on the lengths of the two delay lines since these control the required phase difference as explained in section I (Fig. 1c and Fig. 2b-e). However, the multi-port antenna exhibits mutual coupling between the ports (Fig. 5), altering the expected performance. Thus, to find the optimal reflection phase constellation diagram, we vary the delay branch lengths of each switch between 25-250 μm and observe the inscribed polygon area, defined by the unit-cell's reflection coefficients, as shown in Fig. 8b. When the area of the polygon is maximized, all the reflection coefficient points are close to equispaced (ideal distribution), thus providing the optimal branch lengths. This optimization is performed by mathematically combining the scattering parameter matrices of the multi-port antennas (acquired by full-wave simulations) and the reflection coefficients of the ideal switches with the delay lines (acquired by full-wave simulations), as depicted in Fig. 2a since the full-wave optimization of the unit-cell is a computationally-intensive process. The optimized branch length values are given in table II. Afterward, we use the optimal delay lengths to design the unit-cells in the full-wave simulator and acquire unit-cell's reflection coefficients tabulated in table A-I (appendix). The ideality of our phase states distribution is 93% and is given by the ratio of the full-wave simulated polygon area (Fig. 8b) over the ideal polygon area (Fig. 2e).

Similarly, we devise the 3- and 4-bit unit-cells, by integrating SPST switches with different delay lines at each port of the multi-port antenna. Each of the single-bit switches behaves as an S.C. or O.C. termination, leading to 8 and 16 reflection phase states for the 3- and 4-bit unit-cells, respectively. The lengths of the delay branches are optimized to acquire the maximum area of the inscribed polygon (as in the 2-bit case) and the optimal values are given in table II. The simulated unit-cell reflection phase states (constellation diagram) for the 3- and 4-bit schemes are shown in Fig. 8c-d; the exact unit-cell reflection coefficients values are tabulated in table A-1 (appendix). Likewise, we compute the ideality of the acquired constellation diagrams by calculating the ratio of the inscribed area of the optimized polygon over the ideal, leading to a 95% and 97% ideality for the 3- and 4-bit unit-cells, respectively.

TABLE II
MULTI-BIT UNIT-CELL BRANCH LENGTHS AND IDEALITY

Configuration	l_1 (μm)	l_2 (μm)	l_3 (μm)	l_4 (μm)	Ideality
1-bit	25	-	-	-	95 %
2-bit	25	61	-	-	93 %
3-bit	82	187.5	127.5	-	95 %
4-bit	61	36	96	126	97 %

IV. MULTI-BIT RRS DESIGN

In this section, we analyze the scattering properties of linear RRSs using the multi-bit unit cell designs of the previous section III. We use full-wave simulations to evaluate the

performance of the proposed multi-bit RRSs when mutual coupling (e.g. surface waves) is also considered among the elements.

To demonstrate the performance of the proposed RRSs, we use the multi-bit unit-cells to devise 21-element linear RRSs ($\lambda_0/2$ spacing at 275 GHz) that are illuminated by boresight ($\theta_{imp}=0^\circ$) plane-waves and redirect the waves at 30 degrees. The phase coding profile for each bit configuration is derived by quantizing (1) to the available phase states and resulting RRS phase profiles are given in table A-2 (appendix). The simulated RCS patterns of the four arrays, alongside their calculated patterns using (3), are given in Fig. 9a-d. As clearly shown by these figures, (3) is accurately capturing the shape of the main lobe and the maximum side-lobe level (SLL), but the sidelobe positions and shape differ since the coupling between the neighboring antennas results in surface waves that alter the surface field's profile. Moreover, all the simulated radiation patterns are given in Fig. 10, normalized to the 4-bit case. As expected, the greater the number of bits, the smaller the quantization errors, leading lower SLL level. Besides better beamforming, higher bits also improve the aperture efficiency of the RRSs.

TABLE III
MULTI-BIT RRS PERFORMANCE

Configuration	f_c (GHz)	BW (%)	$\epsilon_{ap,sim}$
1-bit	273.25	1.5	0.41
2-bit	274.65	1	0.68
3-bit	274.55	0.9	0.88
4-bit	276.7	0.8	0.94

To evaluate the relative improvement in the aperture efficiency, we compare the ratio of the radar cross section (RCS) of the various RRSs with the RCS of the same sized, planar, perfect electric conductor (PEC) under normal plane-wave illumination. In this comparison, the models do not include any material losses to emphasize the improvement due to the higher quantization schemes. However, the full-wave models incorporate mechanisms that reduce the maximum RCS of the RRS including element-to-element coupling and mismatches. In the presented RRSs, the boresight impinging waves are reflected toward 30° , unlike the PEC reference plate where waves are retroreflected toward the boresight direction. This discrepancy causes the RRS to appear physically smaller than the PEC plate due to the oblique projection angle, thus the calculated aperture efficiencies are normalized by a factor of $\cos(30^\circ) \approx 0.87$ to account for this effect. The results of the normalized aperture efficiency $\epsilon_{ap,sim}$ are compared in Table III and confirm that the proposed topology can achieve more than 50% efficiency improvement with respect to the single-bit topology.

Fig.9 The full-wave E-plane RCS patterns of the 21-element RRS at their respective center frequency versus the calculated using (3), for the 1,2,3, and 4-bit topologies in (a)-(d), respectively. On the top of each graph, the top-view of the full-wave model is given.

Fig.10 The E-plane RCS patterns of the 21-element RRSs for all four configurations (full-wave), normalized to the maximum gain of the 4-bit design.

All the aforementioned results and discussions concern the center frequency of each array configuration given in table III; however, all antenna systems are required to exhibit a reasonable performance over a specific bandwidth (BW) both for imaging and communication applications. Such thin, high dielectric constant substrates are common in mmWave/THz bands and are typically fabricated, using silicon-on-insulator (SOI) wafers. The use of a thin silicon substrate reduces the 10 dB bandwidth (BW) of the patch antenna close to 1.5 % [37]. On the other hand, the designed patch antennas are not used as receivers or transmitters, though as single/multi-port power splitters/combiners (Fig. 5). Here the BW definition is expressed in terms of radiation characteristics. As such, we define the BW of our RRSs, as the frequency range in which the SLL is below -10 dB, while the impinging waves are redirected at the desired angle.

The simulated (FEM) RCS patterns of the 21-element multi-bit reflective arrays versus frequency are given in Fig. 11, where the frequency range in which the SLL remains below -10 dB is denoted. We notice that the BW of the proposed multi-bit RRSs slightly reduces as the number of bits increases. This is attributed to the frequency-dependent phase delay introduced by the different branches at each port, leading to alterations in reflection phase constellation diagrams of the unit-cells. The simulated BWs and the respective center frequencies are given in table III, where we observe the reduction in the fractional BW versus the number of utilized bits. Moreover, the center frequency of the designs is slightly shifted when modeled in the full-wave simulator due to mutual coupling effects, hence optimization is necessary to align the center frequencies toward fabrication, though this is outside the scope of this work.

Finally, in table IV, a comparison of the existing multi-bit reconfigurable surface designs is presented to emphasize the novelty of our designs. All the tabulated designs utilize multi-bit reconfiguration of the reflected waves with the use of SPST switches (e.g., PIN diodes).

Fig.11 The E-plane RCS patterns of the proposed RRSs versus frequency. With red dotted lines, the upper and lower operating frequencies are denoted.

TABLE IV
MULTI-BIT REFLECTIVE SURFACES WITH SPST SWITCHES

Ref.	f_c (GHz)	Topology	Number of Bits	Switch-per-bit
[16]	3.2	Series	2	1.5
[17]	7.3	Shunt	2	1
[18]	2.3	Unspecified	2	2.5
[21]	15.5	Unspecified	2	1
[22]	36.5	Unspecified	2	4
This Work	275	Shunt	4	1

V. DISCUSSION

The proposed multi-bit configurations could be used in reflectarray antennas (near-field illumination) and IRSs (far-field illumination). Moreover, the presented designs can be extended to other frequency bands including microwaves, where multi-bit reconfigurable surfaces are usually implemented with varactors that require high biasing voltages and complex biasing networks compared with PIN diodes. Furthermore, the proposed multi-bit approach could be implemented using other cavity-based antennas including dielectric resonators commonly used in optical frequencies, leading to reconfigurable mirrors in the far-infrared and visible range.

Additionally, although the switch topology in this work is an ideal O.C/S.C switch; the utilized topology is compatible for the use of 2D tunable materials (e.g., vanadium dioxide and molybdenum disulfide), where the sheet resistance ratio between the tunable states (biased/unbiased) is more than 1,000 [30][31]. Similarly to these tunable materials, other switching configurations could be exploited that have low losses (less than 1 dB) and are implemented with traditional solid-state technologies [28][29]. In the case of lossy switches (e.g., graphene [24]-[27]), the performance of the multi-bit configurations is expected to decline as preliminary results indicate [38], though this characterization extends the scope of this work.

Furthermore, the feeding approach proposed herein could be altered to fit the needs of other systems that require densely populated apertures (less than $\lambda_0/2$). For example, to reduce the footprint of the antenna-splitter/combiner, the feeding could be carried out from the ground plane using aperture coupling

techniques [32]. However, in our case, the coplanar feeding approach is preferred, since the use of aperture coupling is a challenging task with the use of thin wafers due to the required multiple layered substrates.

VI. CONCLUSIONS

Herein, we presented, for the first time, a multi-bit (up to 4-bits) RRS design achieving single switch-per-bit schemes. The reconfigurable multi-bit topologies incorporate the necessary power splitter/combiner in the unit-cell using the cavity model theory of the patch antenna, thus overcoming the real-estate constraints arising in such compact designs. Using these novel multi-port antennas, we split the impinging waves into the different ports and by modulating the state of the integrated SPST switches (one per port), we achieve multi-bit reconfigurability up to 4 bits (16 phase states). The performance of our designs, with respect to the ideal multi-bit, is more than 93% for all the proposed configurations. To validate the proposed topologies, we carried out full-wave simulations, that showed a good agreement between the calculated and simulated radiation patterns. Using our approach, we noticed more than 50% relative improvement in the aperture efficiency of the single-bit reconfigurable surface, which is in agreement to the expected improvement using higher-bit phase quantization. Finally, the proposed topologies can be leveraged in a variety of applications and extended in other frequency ranges including microwave and optical frequencies.

APPENDIX

In this appendix, we tabulate the reflection coefficients of the single- and multi-bit unit-cells in table A-1, and A-2 the quantized phases of the 21-element RRSs for beam steering at 30° under boresight plane-wave illumination.

TABLE A-1
THE MULTI-BIT PHASE STATES IN DEGREES (AT 275 GHz)

State Number	1-bit	2-bit	3-bit	4-bit
1	-54°	-114°	-144°	-155°
2	117°	-12°	-102°	-155°
3		59°	-53°	-131°
4		131°	-14°	-117°
5			4°	-96°
6			66°	-54°
7			115°	-41°
8			172°	-34°
9				-10°
10				2°
11				29°
12				48°
13				78°
14				111°
15				118°
16				167°

TABLE A-2
PHASE DISTRIBUTION OF THE 21-ELEMENT RRS FOR BEAM STEERING AT +30°

Antenna	1-bit	2-bit	3-bit	4-bit
1	-54°	-12°	4°	2°
2	-54°	-114°	-53°	-96°
3	117°	-114°	-144°	-155°
4	117°	131°	115°	118°

5	-54°	59°	66°	48°
6	-54°	-12°	-14°	-10°
7	117°	-114°	-102°	-96°
8	117°	-114°	-144°	-155°
9	117°	131°	115°	111°
10	-54°	59°	4°	29°
11	-54°	-12°	-53°	-41°
12	117°	-114°	-102°	-117°
13	117°	131°	172°	167°
14	117°	59°	115°	78°
15	-54°	-12°	4°	2°
16	-54°	-12°	-53°	-54°
17	117°	-114°	-144°	-131°
18	117°	131°	172°	167°
19	-54°	59°	66°	78°
20	-54°	-12°	-14°	-10°
21	-54°	-114°	-102°	-96°

REFERENCES

- [1] T. S. Rappaport *et al.*, "Wireless Communications and Applications Above 100 GHz: Opportunities and Challenges for 6G and Beyond," *IEEE Access*, p. 1, 2019.
- [2] X. Li *et al.*, "120 Gb/s Wireless Terahertz-Wave Signal Delivery by 375 GHz-500 GHz Multi-Carrier in a 2×2 MIMO System," *J. Light. Technol.*, vol. 37, no. 2, pp. 606–611, 2019.
- [3] P. C. Theofanopoulos, M. Sakr, and G. C. Trichopoulos, "Multistatic Terahertz Imaging Using the Radon Transform," *IEEE Trans. Antennas Propag.*, p. 1, 2019.
- [4] S. Vakalis, L. Gong, Y. He, J. Papapolymerou and J. A. Nanzer, "Experimental Demonstration and Calibration of a 16-Element Active Incoherent Millimeter-Wave Imaging Array," in *IEEE Transactions on Microwave Theory and Techniques*, 2020.
- [5] C. S. Merola and M. N. Vouvakis, "Massive MIMO Beamforming on a Chip," *2019 IEEE International Symposium on Antennas and Propagation and USNC-URSI Radio Science Meeting*, Atlanta, GA, USA, 2019, pp. 1477-1478.
- [6] S. Trinh-Van, J. M. Lee, Y. Yang, K. Lee and K. C. Hwang, "A Sidelobe-Reduced, Four-Beam Array Antenna Fed by a Modified 4×4 Butler Matrix for 5G Applications," in *IEEE Transactions on Antennas and Propagation*, vol. 67, no. 7, pp. 4528-4536, July 2019.
- [7] S. S. Ahmed, A. Schiessl, and L. Schmidt, "A Novel Fully Electronic Active Real-Time Imager Based on a Planar Multistatic Sparse Array," in *IEEE Transactions on Microwave Theory and Techniques*, vol. 59, no. 12, pp. 3567-3576, Dec. 2011.
- [8] Z. Hu, C. Wang and R. Han, "A 32-Unit 240-GHz Heterodyne Receiver Array in 65-nm CMOS With Array-Wide Phase Locking," in *IEEE Journal of Solid-State Circuits*, vol. 54, no. 5, pp. 1216-1227, May 2019.
- [9] http://europractice-ic.com/wp-content/uploads/2019/06/190517_MPW2019-general-v7.0.pdf
- [10] V. Jamali, A. M. Tulino, G. Fischer, R. Muller, and R. Schober, "Scalable and Energy-Efficient Millimeter Massive MIMO Architectures: Reflect-Array and Transmit-Array Antennas," *ICC 2019 - 2019 IEEE International Conference on Communications (ICC)*, Shanghai, China, 2019, pp. 1-7.
- [11] M. Esquius-Morote, J. S. Gómez-Díaz, and J. Perruisseau-Carrier, "Sinusoidally Modulated Graphene Leaky-Wave Antenna for Electronic Beamsteering at THz," in *IEEE Transactions on Terahertz Science and Technology*, vol. 4, no. 1, pp. 116-122, Jan. 2014.
- [12] Y. Cheng, L. Wu, M. Tang, Y. Zhang and J. Mao, "A Sinusoidally-Modulated Leaky-Wave Antenna With Gapped Graphene Ribbons," in *IEEE Antennas and Wireless Propagation Letters*, vol. 16, pp. 3000-3004, 2017.
- [13] H. Kamoda, T. Iwasaki, J. Tsumochi, T. Kuki, and O. Hashimoto, "60-GHz Electronically Reconfigurable Large Reflectarray Using Single-Bit Phase Shifters," *IEEE Trans. Antennas Propag.*, vol. 59, no. 7, pp. 2524–2531, 2011.
- [14] D. Rodrigo, L. Jofre, and J. Perruisseau-Carrier, "Unit-cell for frequency-tunable beamsteering reflectarrays," *IEEE Trans. Antennas Propag.*, vol. 61, no. 12, pp. 5992–5999, 2013.
- [15] S. Costanzo, F. Venneri, A. Raffo, and G. Di Massa, "Dual-Layer Single-Varactor Driven Reflectarray Cell for Broad-Band Beam-Steering and

- Frequency Tunable Applications," *IEEE Access*, vol. 6, pp. 71793–71800, 2018.
- [16] L. Li *et al.*, "Machine-learning reprogrammable metasurface imager," *Nat. Commun.*, vol. 10, no. 1, p. 1082, 2019.
- [17] C. Huang, B. Sun, W. Pan, J. Cui, X. Wu, and X. Luo, "Dynamical beam manipulation based on 2-bit digitally-controlled coding metasurface," *Sci. Rep.*, vol. 7, no. September 2016, pp. 1–8, 2017.
- [18] L. Dai *et al.*, "Reconfigurable Intelligent Surface-Based Wireless Communications: Antenna Design, Prototyping, and Experimental Results," in *IEEE Access*, vol. 8, pp. 45913–45923, 2020.
- [19] Y. Han, W. Tang, S. Jin, C. Wen, and X. Ma, "Large Intelligent Surface-Assisted Wireless Communication Exploiting Statistical CSI," in *IEEE Transactions on Vehicular Technology*, vol. 68, no. 8, pp. 8238–8242, Aug. 2019.
- [20] H. Yang *et al.*, "A Study of Phase Quantization Effects for Reconfigurable Reflectarray Antennas," *IEEE Antennas Wirel. Propag. Lett.*, vol. 16, pp. 302–305, 2017.
- [21] X. Zhang, H. Zhang, J. Su, and Z. Li, "2-Bit programmable digital metasurface for controlling electromagnetic wave," 2017 Sixth Asia-Pacific Conference on Antennas and Propagation (APCAP), Xi'an, 2017, pp. 1–3.
- [22] J. Rodriguez-Zamudio, J. I. Martinez-Lopez, J. Rodriguez-Cuevas, and A. E. Martynyuk, "Reconfigurable reflectarrays based on optimized spiraphase-type elements," *IEEE Trans. Antennas Propag.*, vol. 60, no. 4, pp. 1821–1830, 2012.
- [23] G. Perez-Palomino *et al.*, "Design and Demonstration of an Electronically Scanned Reflectarray Antenna at 100 GHz Using Multiresonant Cells Based on Liquid Crystals," *IEEE Trans. Antennas Propag.*, vol. 63, no. 8, pp. 3722–3727, 2015.
- [24] P. C. Theofanopoulos and G. C. Trichopoulos, "On-Wafer Graphene Devices for THz Applications Using a High-Yield Fabrication Process," 2019 IEEE MTT-S International Microwave Symposium (IMS), Boston, MA, USA, 2019, pp. 1107–1110.
- [25] P. C. Theofanopoulos and G. C. Trichopoulos, "Toward Large-Scale Dynamically Reconfigurable Apertures Using Graphene," 2019 IEEE International Symposium on Antennas and Propagation and USNC-URSI Radio Science Meeting, 2019, pp. 511–512.
- [26] P. C. Theofanopoulos, S. Ageno, Y. Guo, S. Kale, Q. H. Wang, and G. C. Trichopoulos, "High-yield fabrication method for high-frequency graphene devices using titanium sacrificial layers," *J. Vac. Sci. Technol. B*, vol. 37, no. 4, p. 41801, Jun. 2019.
- [27] P. C. Theofanopoulos and G. C. Trichopoulos, "Modeling and Analysis of Terahertz Graphene Switches for On-Wafer Coplanar Transmission Lines," *J. Infrared, Millimeter, Terahertz Waves*, vol. 41, no. 7, pp. 758–775, 2020.
- [28] T. Shivan *et al.*, "220–325 GHz high-isolation SPDT switch in InP DHBT technology," *Electron. Lett.*, vol. 54, no. 21, pp. 1222–1224, 2018.
- [29] F. Thome and O. Ambacher, "Highly Isolating and Broadband Single-Pole Double-Throw Switches for Millimeter-Wave Applications Up to 330 GHz," *IEEE Trans. Microw. Theory Tech.*, vol. 66, no. 4, pp. 1998–2009, 2018.
- [30] M. Kim, S. Park, A. Sanne, S. K. Banerjee, and D. Akinwande, "Towards mm-wave nanoelectronics and RF switches using MoS₂ Semiconductor," in 2018 IEEE/MTT-S International Microwave Symposium - IMS, 2018, pp. 352–354.
- [31] C. Hillman, P. Stupar, and Z. Griffith, "Scalable vanadium dioxide switches with submillimeter-wave bandwidth: VO₂ switches with improved RF bandwidth and power handling," in 2017 IEEE Compound Semiconductor Integrated Circuit Symposium (CSICS), 2017, pp. 1–4.
- [32] C. A. Balanis, *Antenna Theory: Analysis and Design*, 3rd ed. Hoboken, NJ, USA: Wiley, 2005.
- [33] N. Dib, "Comprehensive study of CAD models of several coplanar waveguide (CPW) discontinuities," *IEE Proc. - Microwaves, Antennas Propag.*, vol. 152, no. 2, pp. 69–76, 2005.
- [34] Guizhen Zheng, J. Papapolymerou and M. M. Tentzeris, "Wideband coplanar waveguide RF probe pad to microstrip transitions without via holes," in *IEEE Microwave and Wireless Components Letters*, vol. 13, no. 12, pp. 544–546, Dec. 2003.
- [35] S. J. Bleiker *et al.*, "High-Aspect-Ratio Through Silicon Vias for High-Frequency Application Fabricated by Magnetic Assembly of Gold-Coated Nickel Wires," in *IEEE Transactions on Components, Packaging and Manufacturing Technology*, vol. 5, no. 1, pp. 21–27, Jan. 2015.
- [36] ANSYS Electronics Desktop available at: <https://www.ansys.com/products/electronics/ansys-electronics-desktop>
- [37] D. M. Pozar, "Microstrip antennas," in *Proceedings of the IEEE*, vol. 80, no. 1, pp. 79–91, Jan. 1992.
- [38] P. C. Theofanopoulos and G. C. Trichopoulos, "A Novel 2-Bit Graphene Reconfigurable Reflectarray," 2020 IEEE International Symposium on Antennas and Propagation and USNC-URSI Radio Science Meeting, Montreal, Quebec, Canada, 2020.

Panagiotis C. Theofanopoulos was born in Athens, Greece, in May 1990. He received a Diploma degree and master's degree in Electrical and Computer Engineering from Democritus University of Thrace, Xanthi, Greece, in 2013 and 2015 respectively. In parallel, he participated at a European Union–Greek Government research program at the Microwaves Laboratory, Democritus University of Thrace Research project. In May 2021 he graduated with a Ph.D. from Arizona State University. His research interests are

mmWave/THz antennas, computational electromagnetics, wireless communication systems, and biometrics.

Georgios C. Trichopoulos received his Diploma degree in electrical and computer engineering from the Democritus University of Thrace, Xanthi, Greece, in 2004, the M.S. degree in biomedical engineering from the National Technical University of Athens and University of Patras, Greece (under a joint program), in 2006, and the Ph.D. degree in electrical and computer engineering from The Ohio State University, Columbus, OH, USA, in 2013.

He is currently an Assistant Professor with the School of Electrical, Computer, and Energy Engineering, Arizona State University, Tempe, AZ, USA. From 2013 to 2015, he was a Postdoctoral Researcher with the ElectroScience Laboratory, The Ohio State University. His research areas include electromagnetic theory, mmWave/THz imaging methods and systems, antenna design for mmWave/THz sensors, and high-frequency device and circuit characterization methods.