3D Beamsteering Low Complexity Reconfigurable Multilevel Antenna

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Abstract—A low complexity reconfigurable antenna (RA) capable of steering its beam into nine different directions corresponding to $\theta = \{-42^\circ, 0^\circ, 42^\circ\}$ in the $\phi = \{25^\circ, 0^\circ, 155^\circ\}$ planes, and $\theta = \pm 25^\circ$ in the $\phi = 90^\circ$ plane, using a reduced number of switches is presented. The RA operating at 3.4 – 3.8 GHz consists of a central dipole element excited by a coaxial probe, surrounded by four switchable parasitic dipoles located at different height levels above a ground plane which act as either reflector for the elements located at an upper layer or director for the ones at a lower layer, improving the radiation characteristics of the antenna. Each parasitic dipole is fed with a PIN diode as switch which enables the activation of these elements providing the desired modes of operation. The presented RA was manufactured and measured showing results with a good agreement with the theoretical and numerical simulation.

Index Terms—Broadband antennas, Reconfigurable antennas, Beam-steering antennas, Antenna beamforming.

I. INTRODUCTION

Beamforming, in its many variants, is a key spatial processing technique to improve user throughput, system capacity, system coverage as well as reducing interference [1]. Simple architectures enabling beamforming either in predefined or arbitrary directions are very desirable for the Fifth Generation of Mobile Communications (5G) to boost power efficiency. Furthermore, it is expected that the number of 5G mobile subscribers grows from 5 million in 2019 to nearly 600 million by 2023 [2], increasing traffic and connections density, which will increase the demand of capacity to the network.

Two appealing cost-efficient and low-complexity alternatives to conventional beamformers, which are generally based on phase-shifters or line-delays matrices, are the Parasitic Layer (PL) and the Electronically Steerable Parasitic Array Radiator (ESPAR), whose principles are the basis of the antenna presented in this work.

The Reconfigurable PL concept [3] is an approach to provide added functionalities to a basic antenna, e.g. the ability to adjust dynamically the matching frequency, polarization and radiation properties. The switches used in the PL can be either PIN diodes [4, 5], microelectromechanical system (MEMS) switches [6] or varactor diodes [7]. RAs using PIN diodes are widespread due to their fast switching speed and low cost

compared to MEMS, and higher energy efficiency and reduced losses when compared with varactors.

One of the successful applications of the PL concept is to add beam-steering capabilities to microstrip patch antennas. Similarly, large and medium complexity PL [8, 9] have been used to add polarization shifting to single-port antennas, although this usually comes at the expense of using a large amount of switching elements. Cost and reliability criteria have led to recent modifications of the PL concept, aiming to reduce the number of switches and thus the complexity of design. Three examples of simplified PL are [10–12]. The design outlined in [12], despite using a very reduced number of switches, is bandwidth limited.

On the other hand, ESPAR beamformers, originally proposed in [13], consist of a driven antenna element surrounded by N reactively loaded parasitic elements. With this arrangement, beamsteering as well as superdirectivity can be achieved by changing the reactive impedance of each parasitic element. Later works based on similar ideas and coining the ESPAR term can be found in [14, 15].

In comparison to the antennas presented in the literature, the novelty of the proposed antenna lies in its capability to steer its beam to up to nine different angular directions based on a novel multilevel geometry using only four PIN diodes as switches. Additionally when compared with the reconfigurable geometries based on patch antennas, the use of dipoles as proposed here result on an improved frequency bandwidth behaviour.

II. ANTENNA STRUCTURE AND WORKING PRINCIPLE

A. Geometry of the Antenna

The Reconfigurable Antenna (RA) whose structure is shown in Fig. 1, is mainly composed by a square ground plane ($W_g \times W_g$) and three layers above it. The substrate used for each layer is Rogers 3003C ($\epsilon_r = 3$) with 0.5 mm of thickness. Each layer is separated by nylon supports at h_0, h_1 and h_2 from the ground plane, as illustrated in Fig. 2. An active dipole with total length L_a and width a resonating at 3.5 GHz is placed at the middle layer, whose feeding is provided by means of a coaxial probe introduced from below the ground plane, incorporating also a quarter wavelength balun [16] for the symmetrization of branch currents.

This active element is surrounded by four parasitic dipoles which can be independently activated or deactivated. Two of them with length L_a at the upper layer, and the other ones with a shorter length L_d at the lower layer. Relevant antenna dimensions are summarized in Table I.

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Figure 1: Isometric view of the RA.

λ	$85.7mm$	a	0.14λ
L_a	0.39λ	h_0	0.123λ
L_d	0.342λ	h_1	0.21λ
d_x	0.32λ	h_2	0.25λ
d_u	0.455λ	W_a	2.4λ

Table I: Antenna design parameters.

By activating a PIN diode-based switch that connects the arms of the parasitic dipoles, a change in their electrical length occurs and a significant current is induced on it. The phase difference in the induced current in the parasitic element similarly to the Yagi-Uda concept and the multilevel geometry are responsible for steering the beam in a certain direction. When the switch is in its OFF state, the current induced on each parasitic is negligible at the frequency of interest.

The biasing lines of the diodes are connected to each reconfigurable dipole by means of $\varnothing$ 0.5 mm metallic wires from below the ground plane, where a low frequency circuit printed on a FR4 substrate was designed for controlling each diode state. Although these biasing lines are perpendicular to the currents, they still have an impact on the coupling of the parasitic elements with the active element, design optimization should consider them as they change slightly the pointing angle and the matching of the antenna.

A total amount of 19 Murata 10nH inductors (LQG18HN10NJ00D) were located at strategic positions in order to decouple as much as possible the RF signal from the DC signal, see figure 5. Also, 4 Infineon BAR64 Series diodes (BAR6402VH6327XTSA1) are used as switches. An Arduino micro-controller is placed in the back to control the different modes of operation of the RA.

B. Working Principle

The directional capability of the multilevel phasing concept, stated in the previous section, when combined with a ground plane produces additional upwards reflected beam increasing up to 9 the directional modes of operation, summarized in Table II. In this section an intuitive mathematical formulation combining the use of mutual impedances, array factor and image theory is presented and validated for all the modes of operation.

Figure 2: Side view of the RA.

Mode	$(\theta, \phi)_{max}$	S_1	S_2	S_3	S_4	$\eta(\%)$
M_0	(0, 0)	0	0	0	0	92.45
M_1	(-42, 0)	0	1	0	0	94.45
M_2	(42, 0)	1	0	0	0	94.45
M_3	(25, 90)	0	0	0	1	87.31
M_4	(-25, 90)	0	0	1	0	87.31
M_5	(42, 155)	0	1	0	1	91.1
M_6	(-42, 155)	1	0	1	0	91.1
M_7	(42, 25)	1	0	0	1	91.1
M_8	(-42, 25)	0	1	1	0	91.1

Table II: Antenna operation modes, 0 corresponds to switch-OFF state and 1 to switch-ON state.

Starting for the mode 0 (broadside radiation) where no switches are activated, the radiated field is just given by the array factor composed by the driven dipole element with its image:

$$\vec{E}_0(\theta, \phi) = 2jI_0\vec{E}_{dip}(\theta, \phi)\sin(kh_1\cos\theta) \quad (1)$$

Where I_0 and $\vec{E}_{dip}(\theta, \phi)$ are the current and the electric field in free space of the driven element.

Regarding the case of having only one parasitic element activated, which corresponds to the modes 1, 2, 3, 4, the voltages and currents in the driven dipole and parasitic ones are related by the impedance matrix Z as done in [17],[18] and [11]:

$$\begin{pmatrix} V_0 \\ I_m X \end{pmatrix} = \begin{pmatrix} Z_{00} & Z_{0m} \\ Z_{m0} & Z_{mm} \end{pmatrix} \begin{pmatrix} I_0 \\ I_m \end{pmatrix} \quad (2)$$

and

$$\frac{I_m}{I_0} = -\frac{Z_{m0}}{Z_{mm} - X} \quad (3)$$

Where X is the impedance of the diode at the frequency of operation. The 0 and m subscripts correspond to the driven and parasitic elements respectively. The mutual impedances Z_{m0} and self-impedance Z_{mm} has to be computed numerically with the presence of the ground plane, and it is observed that approaching the parasitic elements to it, increases the value of the mutual impedance.

Depending on whether the relative phase α_{m0} is positive or negative, the parasitic elements will act as either reflector or director. The position of the two parasitic dipoles at the upper layer were optimized numerically in order to act as

reflectors and the ones at the lower layer as directors by shortening the length. Since the mutual impedance of co-linear dipoles is smaller than for parallel dipoles, to achieve similar steering behaviour in the E plane and the H plane, the co-linear parasitics are closer to the ground plane than the parallel ones. Once we know the relative phase α_{m0} , by applying image theory again the radiated fields are then:

$$\vec{E}_m(\theta, \phi) = 2j\vec{E}_{dip}(\theta, \phi)(\sin(kh_1 \cos \theta) + \frac{I_m}{I_0} e^{jk\vec{d}_m} \sin(k_z h_m)) \quad (4)$$

Where $(\vec{d}_m, h_m)$ corresponds to the position of the parasitic element used for the mode of operation.

Regarding the modes 5, 6, 7, 8, where two parasitic elements are activated at the same time, their voltages and currents relations can also be calculated by the Z impedance matrix. Its corresponding radiated fields are :

$$\vec{E}_{mn}(\theta, \phi) = \vec{E}_m(\theta, \phi) + 2j\vec{E}_{dip} \frac{I_n}{I_0} e^{jk\vec{d}_n} \sin(k_z h_n) \quad (5)$$

Note that the m, n subscripts correspond to the two parasitic elements used.

In Figures 3 and 4, the comparison between the theoretical result with the simulated radiation pattern of the antenna using CST software is shown without taking into account the non-modeled elements such as the dielectric, nylon supports and polarization lines in order to check the accuracy of the model.

Figure 3: Theoretical radiation pattern (---) comparison in linear scale with simulated radiation pattern (—). ZX plane cut for modes 0,1 and 2 and ZY plane cut for modes 3 and 4.

Figure 4: Theoretical radiation pattern (---) comparison in linear scale with simulated radiation pattern (—). ZX plane cut for the left hand side and ZY plane cut for right hand side.

This theoretical approach is more accurate the bigger the ground plane is as image theory assumes an infinite ground plane. A good agreement between the computed analytical solution with respect to the simulated one is obtained for the ZX plane. However there is a difference in the pointing angle for the ZY plane corresponding to modes 3-8 because the elements responsible for steering over this plane, are closer

to the finite ground plane, hence they are more altered by its finiteness.

III. RESULTS

A prototype of the antenna, illustrated in Fig. 5, was manufactured using standard Printed Circuit Board (PCB) technology. The radiation pattern was measured into a 6x6x8 m³ far-field setup available at the UPC in order to validate the theoretical analyses and simulated results.

Figure 5: Manufactured antenna

The measurements included the reflection coefficients and realized gain patterns, that are cut at 3.5 GHz for the four principal modes of operation, that is 0, 1, 3 and 5, the rest ones are symmetric. The simulated and measured results for these modes of operation are shown in figures 6, 7 and 8.

Figure 6: Left Y axis shows the measured (---) and simulated (—) input reflection coefficient for the principal modes of operation 0,1,3 and 5. Right Y axis shows the realized gain in the shadowed region over the working frequency band.

Diodes in ON state were reproduced in the simulations by a lumped element of 2.1 Ω . Only an average of 1% radiation efficiency loss per diode was noticed in the simulations. The

Figure 7: Simulated radiation patterns for each mode of operation

Figure 8: Polar plot in dB of the realized gain comparison between simulated (—) and measured (---) radiation patterns

Reference	No. of switches	No. Modes	Bandwidth (GHz)	Average Gain (dB)	Beam-steering Angle (°)	Efficiency (%)
This work	4	9	3.4-3.8	9	42	90
[12]	4	9	2.36-2.39	7	25	NA
[11]	2	3	4.4-5.1	10	60	80
[10]	6	12	4.9-5.1	8	40	81
[5]	12	9	2.4-2.5	6.5	30	NA
[9]	12	30	1-6	6.5	60	NA

Table III: Comparison to other RA.

measurements show an increased bandwidth when compared to the simulations because a bigger hole was drilled in the ground plane for the insertion of the coaxial probe. After all, they are consistent with the simulations. The different modes cover the bandwidth 3.4–3.8 GHz (just mode 1 is 1 dB above –10 dB from 3.7 to 3.8GHz).

In general there is a trade off between the length of the parasitic dipoles, the height, steering angle and bandwidth. An important issue is the study of the gain stability as reported in [10] and [5] for all the modes of operation. The maximum realized gain of 10.6 dB is achieved at 3.5GHz, however there is also a maximum decrease of 1.6dB realized gain for modes 3 and 5 over the working frequency band, see figure 6. In this RA, again a trade off between gain stability and steering angle has to be faced. A simple solution to achieve both would be reactively loading the switches with capacitors/inductors as first mentoned by R. Harrington [13], so we can keep all the parasitic elements reflective/directive in a higher layer with more stable gain.

Comparison of the RA to other efforts reported in literature

is summarized in Table. III.

IV. CONCLUSIONS

A novel low complexity multilevel dipole RA with beam-steering capabilities into nine different directions has been proposed. In the designed RA, which is based on a new parasitic multilevel directional antenna concept, only four PIN diodes acting as switches were used to reconfigure the parasitic dipoles located at different levels, keeping the efficiency greater than 87.3% for the 9 modes of operation. The operating bandwidth of the antenna ranges from 3.4 – 3.8 GHz and can provide a broad beamwidth at broadside with ~ 7.2 dB gain and $(\theta, \phi) = (25^\circ, 90^\circ)$, $(-25^\circ, 90^\circ)$ with ~ 8.6 dB gain. The rest of directions corresponding to $(42^\circ, 0^\circ)$, $(-42^\circ, 0^\circ)$, $(42^\circ, 155^\circ)$, $(-42^\circ, 155^\circ)$, $(42^\circ, 25^\circ)$, $(-42^\circ, 25^\circ)$ provided realized gains greater than ~ 9.8 dB. The antenna is suitable for 5G New Radio (NR) Frequency Range 1 (FR-1) applications where broadband, high directivity, high efficiency and beamforming can be translated into higher capacity.