Beam Steering Using Active Artificial Magnetic Conductors: A 10-Degree Step Controlled Steering

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ABSTRACT An Active Artificial Magnetic Conductor (AAMC) is presented to steer the radiation pattern of a printed dipole working at 2 GHz. The elements that generate the phase shift are a set of Varactor Diodes, which are characterized using its spice model in order to obtain a phase shift - capacitance mapping. Overall beam steering of +/- 40° with a step size of 10° is achieved. A circuit model that describes any multilayer substrate AAMC unit cell, which uses fist form of Foster’s theorem along with transmission line theory, is proposed. Our work is suitable to be used as low profile antenna; for example, street furniture antennas, which are located on the facades of houses or buildings, so that they can be visually mixed up with signs or advertisements. Simulations have been validated using a prototype consisting of an array of 22 × 14 AAMC elements; the overall structure measures 1.9λ₀ × 1.21λ₀. This reflector will generate a phase gradient in its columns, which will modify the reflection angle of an incident electromagnetic wave in the H-Plane. Beam switching control has been achieved using suitably amplified LPF PWM signals generated by two Arduino modules. A printed dipole with a Fractional Bandwidth of 17% is designed and manufactured to illuminate the structure at a distance λ₀/4 above the surface. Far-field radiation patterns and reflection coefficients have been measured in an anechoic chamber using a spherical system. These compare favorably with simulations performed using the Time Domain solver in CST Microwave Studio.

INDEX TERMS Reflector antennas, equivalent circuits, frequency selective surfaces, artificial magnetic conductors.

I. INTRODUCTION

Artificial Magnetic Conductors (AMC) are considered to be in-phased reflectors, i.e. they reflect an incident electromagnetic wave generated by a radiating element constructively, even if the element is in the radiating zone [1]. Additionally, AMCs suppress surface current waves, this means that there is no power loss in the dielectric, thus increasing the directivity of antennas. These structures are comprised by one or multiple Frequency Selective Surface (FSS) at the top, separated by one or more substrates in the middle, a ground plane at the bottom, and vias that connect the FSS to the ground plane. The vias could be omitted in the design if the angular stability of the design is not critical due to normal incidence of the impinging electromagnetic wave [2]. The first time this type of structures were introduced with a theoretical model and validated through the manufacture of a prototype, was in the work presented by Sievenpiper, et al. in [3]. Since that publication, a great number of researches have been carried out on this topic, from which several applications have been found, being the most relevant: Electromagnetic Band Gap (EBG) [3], low profile antennas [4], Radar Cross Section (RCS) reduction [5], Luneburg Lens [6], and finally, electronically beam-steering antennas [7].

The use of AMCs as reflective surfaces has been widely investigated and many recent works show important advances, especially in increasing the operational bandwidth. The work [8] employs an AMC as a reflective surface, which provides a 90% bandwidth, for a dual-polarized dipole antenna, intended to be used in 2G/3G/4G technology. This
wide bandwidth is achieved by using a combination of air and substrate as elements that separate the FSS and the ground plane. In [9] and [10], like the previous work, uses an AMC as a reflective surface, but in this case for an array of dipoles and slot antennas, respectively. In all the works described above the main use of the AMC is to lower the profile of the antennas, but do not steer the radiation pattern by means of changing the AMC properties.

In the literature regarding beam steering using AMCs, it has been found that there are two categories of beam steering according to the position of the radiating element. The first is by placing the radiating element in the Far-Field zone and the second is by placing the radiating element in the Near-Field zone.

In the first category, a noteworthy paper is [11], which considers the manufacture of an Active Artificial Magnetic Conductor (AAMC) illuminated by a horn antenna located in the Far-Field zone. Reflectarray theory has been used to generate beam steering and thus overcoming the issues related to the illumination angle of the horn antenna on the surface. Recent work that uses the previous approach can be found in the literature. In [12], the authors use the same principle of an AAMC to steer the radiation pattern of a horn antenna intended to be used for satellite communications. In [13], the authors use an AAMC with lossy elements to reduce-redirect the impinging electromagnetic wave, thus reducing the RCS of the surface. All the previous works have their radiating elements placed in the Far-Field zone, while our contribution operates in the Near-Field zone.

For the second category, there has been significant research in the area of low profile antennas. Three of the most relevant include [14], where the authors employ an AMC as ground plane for a bowtie antenna with re-configuration capabilities to generate a beam steering in the H-plane. In [15], the authors propose to steer the radiation pattern of a slot antenna using an AMC as reflecting surface. The slot antenna has re-configuration capabilities due to a Pin Diode, which redirects the currents in the antenna. In both previous cases the antenna does the beam steering and not the AMC surface. The authors of [16] use an AAMC that is composed with a combination of liquid crystal and substrate to change the reflection phase of the structure, thus steer the radiation pattern of an array of three patch antennas. This work is only suitable for frequencies above 20 GHz due to the thickness limitation of the liquid crystal. Another draw back of this work is the high cost. Unlike the method presented in [14] and [15], this contribution generates beam steering by changing the properties of the AAMC, while our technique is cheaper to implement than [16].

It has been shown in [7] that the radiation pattern of a dipole can be redirect in different directions than broadband by changing the AMC properties. A drawback of this radiation pattern control is that all the Varactor Diodes share the voltage source; therefore, there is no progressive shift used to control the beam steering. The work presented in [17], proposes an application for the design of furniture antennas, which electronically controls the beam steering by using AAMC, while maintaining a reduced antenna profile. The drawback of this work is that the control of the AMC is the same as the previous work. In our work we control the phase shift by columns in order to generate a uniform phase gradient, allowing to have a controlled beam steering.

Our work can be used to modify the radiation pattern, not only for street furniture antennas, but also for conventional cellular base stations, wifi access points, radar stations, and thus provide greater coverage to certain areas where more power is needed. Additionally, it is applicable to satellite communications, where we could modify the radiation pattern of a very directive element and thus accentuate or reduce the footprint in a certain area where we intend to provide or deny services.

The main contributions of this paper are as follows:

1) An equivalent circuit model analysis is proposed for an AAMC, which describes its behavior even if it has multiple dielectric mediums. The results compares favorably with the full-wave simulations.
2) A beam steering analysis is presented, as well as reflectarray analysis, which allow us to have a start point for our optimization process.
3) An external biasing circuit is described and implemented for the AAMC.
4) A full structure prototype circuit that comprises the AAMC, biasing circuit, and radiating element, is manufactured and tested in an anechoic chamber.
5) A method to increase the gain of the overall structure, is detailed and simulations are presented.

This paper is organized as follows: Section II presents a circuit model of a generic AAMC, based on transmission line theory and equivalent circuit theory. The design of an AMC, AAMC unit cell, and the complete structure are introduced in Section III. Additionally, the design of the printed dipole is also presented in Section III. The procedure to calculate the phase gradient, followed by the optimization process, to achieve the steering angles, are described in Section IV. The simulations and measurements for the complete structure and the dipole antenna are discussed in Section V. A method to increase the gain of the structure is presented in Section VI. Finally, our work is summarized with a set of conclusions and future work in Section VII.

II. EQUIVALENT CIRCUIT MODEL ANALYSIS

Before presenting the design, of the AMC, antenna and the complete structure, first an equivalent circuit model will be shown, which will allow us to relate the physical components of an AMC to its frequency response.

An AMC can be modeled by means of an equivalent circuit using inductors, capacitors, resistors, and transmission lines, which represent Frequency Selective Surfaces, dielectric mediums, ground plane, vias and Varactor Diodes. A sketch of an AMC and its equivalent circuit are illustrated...
in Fig. 1. The analysis of the equivalent circuit model will be conducted through the following steps:

1) Frequency Selective Surface analysis: The FSS is modeled using Foster’s theorem, thus allowing to obtain a resonant equivalent circuit, which mainly handles the resonance frequency and part of the bandwidth of the surface.

2) Dielectric medium and ground plane analysis: Both the dielectric medium and ground plane are modeled as transmission lines and will mainly handle the bandwidth of the structure.

3) Analysis of the Varactor Diode. The Varactor Diode is modeled by its lumped elements R, L, and C.

Each of the three previously steps are detailed below and gathered to generate an equivalent circuit.

**A. FREQUENCY SELECTIVE SURFACE ANALYSIS**

Because an FSS is essentially a filter, which acts as a band-pass or band-stop, it can be analyzed by obtaining an equivalent circuit. The equivalent circuit can be classified in two groups; if the FSS acts as a band-pass filter, and using Foster’s second form, the equivalent circuit is a parallel RLC. On the other hand, if the surface is band-stop filter, and using Foster’s first form, the equivalent circuit is a series RLC [18]. Both can be used, but depending on which, the bandwidth and central frequency will be affected. The work [19], uses odd and even decomposition for equivalent circuit analysis of Planar Periodic Structures (PPS). Additionally, the previous work give us a guide of how to calculate the values of the equivalent circuit elements. In our analysis, we will ignore the losses of the conductors and dielectrics due to the low frequency operation. To obtain the values of the equivalent circuit elements, C and L, the next steps must be followed:

1) Simulate the FSS structure using full-wave electromagnetic simulation software and extract the S-Parameters.

2) Using the S-parameters, obtain the admittance as a function of frequency by (1). This relates the admittance of a two port network with its S-Parameters.

3) By means of a curve fitting method, calculate C and L that satisfy (2), which is the equivalent admittance of an LC series circuit.

\[
Y_{eq} = 3 \left\{ \frac{2S_{11}}{\eta_0 S_{12}} \right\} \quad (1)
\]

\[
Y_{eq} = 3 \left\{ j \frac{2\pi f C}{1 - (2\pi f)^2 L} \right\} \quad (2)
\]

**B. DIELECTRIC AND GROUND PLANE ANALYSIS**

The dielectrics stack can be modeled using transmission line theory, as illustrated in Fig. 1. For each dielectric, the propagation constant and line impedance must be calculated a priori, using the following formulas.

\[
\gamma_i = \sqrt{-\omega^2 \mu_i \epsilon_i} \quad (3)
\]

\[
Z_{\text{dielectric}_i} = \frac{\eta_0}{\sqrt{\epsilon_i \eta_0}} \quad (4)
\]

where, \( \omega \) is the angular frequency, \( \epsilon_i \) and \( \mu_i \) are the permittivity and permeability of the \( i \)th medium, \( \eta_0 \) is the free space impedance, and \( \epsilon_{eff} \) is the effective permittivity of the \( i \)th medium.

The ground plane can be modeled as an impedance, which is obtained using short ended transmission line formula, as follows.

\[
Z_g = \frac{\eta_N Z_{\text{short}} + \eta_N \tanh(\gamma N l_N)}{\eta_N + Z_{\text{short}} \tanh(\gamma N l_N)} \quad (5)
\]

As \( Z_{\text{short}} \) is zero; therefore,

\[
Z_g = \frac{\eta_N \tanh(\gamma N l_N)}{\eta_N} \quad (6)
\]

where, \( \gamma N \) and \( l_N \) are the propagation constant and thick of the \( N \)th medium, respectively.

Finally, the equivalent circuit of the multi-dielectric AMC is obtained by:

\[
Z_{\text{AMC}} = Z_{\text{FSS}}||Z_{eq} \quad (7)
\]

here, \( Z_{\text{FSS}} \) is the impedance of the FSS and \( Z_{eq} \) is the equivalent impedance corresponding to the stacked dielectrics and ground plane. To obtain \( Z_{eq} \) we need to use an iterative process using (8).

\[
Z_{N-1} = \frac{Z_{N-1} + Z_{N-1} \tanh(\gamma N-1 l_{N-1})}{Z_{N-1} + Z_{N-1} \tanh(\gamma N-1 l_{N-1})} \quad (8)
\]
C. VARACTOR DIODE EFFECT OVER THE FSS

The effect produced by the addition of the Varactor Diode placed between the plates, is only affected in the FSS analysis and can be represented by placing the equivalent circuit of the Varactor Diode in parallel with the equivalent capacitance of the FSS structure. In addition, a series inductance must be added to the original circuit, which models the inductance generated due to soldering and packaging, as illustrated in Fig. 2. In addition, it should be mentioned that the resistance generated by the Varactor Diode cannot be ignored due to its high value.

![Fig. 2. Equivalent circuit corresponding to a dielectric-less FSS with a Varactor Diode placed between the patch plates. The transmission lines represent the free space impedance.](image)

III. ACTIVE ARTIFICIAL MAGNETIC CONDUCTOR FULL STRUCTURE DESIGN

A. AAMC UNIT CELL DESIGN

The unit cell is designed to work at 2 GHz, as a proof of concept, and exhibits a high Figure of Merit (FoM), so that the structure is compact for the working frequency range. One method to increase the FoM is by the combination of two substrates. In this case, air has been used after the substrate that holds the FSS structure. We use the Sievenpiper structure, which is landed to the ground plane by the vias, as illustrated in Fig. 3. Later, the vias will be used for the Varactor Diode polarization. The response of the unit cell is calculated considering all the steps described in the previous section.

1) FSS DESIGN STAGE

The initial analysis of FSS is carried out in the absence of any substrate, in which an equivalent admittance will be obtained by means of Foster’s first form as indicated in Section II. The dimensions of the unit cell are: \( a = 9 \) mm, \( b = 0.5 \) mm, \( c = 10 \) mm, the metal thickness is 0.035 mm. As illustrated in Fig. 4, the curves of parameters \( S_{11} \) and \( S_{12} \), which are obtained in the Equivalent Circuit Method, compare favorably with Full-Wave Simulation Method. The values obtained by curve fitting are: \( L_{FSS} = 0.1041 \) nH and \( C_{FSS} = 107.87 \) pF.

![Fig. 3. Dimensions associated to the AMC Unit Cell. The black areas represent the vias, light gray areas represents dielectric, and dark gray areas represents metal. In addition, the Varactor Diode position is also illustrated. a) Top view b) Side view.](image)

![Fig. 4. Magnitudes of the reflection and transmission coefficients of a square patch FSS using Full-Wave Simulation (FWS) and Equivalent Circuit Simulation (ECS) methods. Results for Both simulation methods show good agreement.](image)

The effect of the Varactor Diode over the FSS structure can be analyzed by its spice model, which is a series RLC circuit. The values of the RLC circuit are: \( R_{VAR} = 2.5 \) Ohms, \( C_{VAR} = 0.8 \) fF, and \( L_{VAR} = 0.7 \) nH. The previous values correspond to a real Varactor Diode, which will be described in the next sub-section. Placing the Varactor Diode produces a substantial decrease in the transmission first null, as can be seen in Fig. 5. This decrease is generated due to the increase of the equivalent capacitance and inductance; therefore, by (9), the resonance frequency decreases. Again, it can be seen that the results for Equivalent Circuit and Full-Wave Simulation Methods are in good agreement.

\[
f_r = \frac{1}{2\pi \sqrt{LC}} \quad (9)
\]

If the substrate is added, the equivalent capacitance \( C_{FSS} \) increases to 256.23 fF. This boost is due to the fact that
the permittivity of the dielectric medium between the parallel plates has increased. The effect of the substrate in the S-Parameters is illustrated in Fig. 6.

2) AAMC DESIGN STAGE
The design of the AAMC was performed using the previously modeled FSS, a layer of air with thickness \( e = 2 \ mm \), and a ground plane. To clarify, we will call our unit cell “AAMC” from now on, due to the Varactor Diode that controls the phase response and will be feed with a power source. The mechanical stability needed to support the FSS is achieved using the same vias that bias the Varactor Diode. The final equivalent circuit of our design is presented in Fig. 1; where: the \( Y_F \) admittance corresponds to the equivalent admittance of the circuit illustrated in Fig. 2, the first transmission line is generated by the air layer, and the short circuit is generated by the ground plane. The configuration required for the simulation in CST Microwave Studio is illustrated in Fig. 7. The unit cell is simulated using the Frequency Domain Solver and unit cell boundary conditions, which automatically set the Floquet ports excitations in both, positive and negative z directions. Because our structure has a ground plane in the positive z direction, the Floquet port in that direction has been removed and replaced with boundary condition \( E_t = 0 \). Additionally, because the phase of the reflection coefficient is of interest, a reference plane must be configured. For this reason, we set a distance for the reference plane with a value of approximately dembedded \( = C_L f_r / 2 \), where, \( C_L \) is the speed of light and \( f_r \) is the zero phase crossing frequency. The Varactor Diode is simulated using a series RLC lumped elements preconfigured in CST Microwave Studio. To ensure adequate meshing, the Varactor Diode is placed within a rectangle, whose dimensions corresponds to those of a real diode.

The simulation results between both methods are illustrated in Fig. 8, which shows the phase and magnitude of the parameter \( S_{11} \). These results show excellent agreement, which proves the effectiveness of our equivalent circuit model. In theory, any AAMC or AMC structure composed of one or more dielectrics and Varactor Diodes can be modeled by using the steps described above.
3) AAMC PHASE SHIFT
The phase shift of the AAMC is generated by the variation of the capacitance in each Varactor Diode. Therefore, a mapping that relates the phase shift with the capacitance is obtained by simulating the AAMC with values of capacitance that range from 0.35 pF to 2 pF. The mapping is acquired at 2 GHz and the Varactor Diode used is the SMV2020, which has low inverse polarization voltage and high Q. The results of the Phase Shift vs Capacitance are illustrated in Fig. 9.

In addition, using the set of values available in the data-sheet of the Varactor Diode, the Voltage vs Capacitance mapping is also illustrated in Fig. 9.

FIGURE 8. Full-Wave simulation and Equivalent Circuit Method simulation of the $\mathcal{S}_{11}$ and $\overline{\mathcal{S}}_{11}$ of the AAMC. The results show good agreement in both phase and magnitude.

FIGURE 9. Phase shift vs Capacitance and Voltage vs Capacitance mapping of the proposed AAMC unit cell using the Varactor Diode SMV2020; both obtained at a central frequency of 2 GHz and at normal incidence.

B. ANTENNA DESIGN
The antenna, illustrated in Fig. 10, used in this design is a printed dipole, which was selected due to its low profile, low overall dimensions, and low shadowing capabilities over a reflector. The antenna is manufactured over a substrate with $\varepsilon_R = 4.3$ and $\tan\delta = 0.02$. The dipole is connected to a 50 Ohms balanced line by means of a printed balun over the same substrate. The dimensions of the antenna are shown in Table 1.

TABLE 1. Printed dipole and balun dimensions.

| Parameter | Value  | Parameter | Value  | Parameter | Value  |
|-----------|--------|-----------|--------|-----------|--------|
| $f$       | 14 mm  | $j$       | 6 mm   | $o$       | 9 mm   |
| $g$       | 17 mm  | $k$       | 3 mm   | $p$       | 40.3 mm|
| $h$       | 45 mm  | $m$       | 27.5 mm| $q$       | 2 mm   |
| $i$       | 20.3 mm| $n$       | 5 mm   | $r$       | 67 mm  |

The antenna was designed to resonate at 2 GHz with a 17% bandwidth, as a proof of concept. Simulation results are computed using CST Microwave Studio by the FEM method and the results are illustrated in Fig. 14.

C. FULL STRUCTURE DESIGN
1) AAMC REFLECTOR
The AAMC reflector has a total of $22 \times 14$ unit cells, which has an overall dimension of $1.47\lambda_0 \times 0.93\lambda_0$. This number of unit cells has been chosen due to the omnidirectional radiation pattern of the dipole, which is aligned to the H-plane. The dimensions of the structure that supports the FSS section of the AAMC are $u = 240$ mm and $t = 165$ mm. Additionally, the structure extends in its ground plane 30% below the AMC structure, which reduces back side lobes and increase the directivity. The dimensions of the ground plane are $v = 286$ mm and $w = 182$ mm. At the top of the structure the connections that allow us to connect the biasing of the AMCs to the Arduino module are shown. The sketch of the AAMC reflector is illustrated in Fig. 11a.

2) ANTENNA PLACEMENT
The printed dipole antenna is placed at the center of the reflector and separated a distance $r = 37.5$ mm ($\lambda_0/4$). Fig. 11b illustrates the side view of the AAMC reflector with the dipole antenna. The antenna is attached to the reflector by...
two arms, printed in 3D using ABS material. The RF feeding cable will be ducted through the printed arms.

3) AAMC Biasing

The biasing of the Varactor Diodes is handled by an external circuit placed at the rear of the AAMC reflector, as illustrated in Fig. 11b. The polarization circuit features four stages, as illustrated in Fig. 12a. The first one, is the stage in which 22 Pulse Width Module (PWM) outputs are generated. Two Arduino Mega 2560s have been programmed for this purpose, where each one allows the generation of 11 PWM outputs. The PWM cycle values will be programmed to generate the desired DC voltage to bias the Varactor Diodes. In the second stage, the PWM signals are converted to DC signals by using an RC filter, which is a low pass filter, thus extracting the DC component. The third stage, amplifies the DC signal to values greater than 0 V and less than 20 V, which are the operating voltage of the Varactor Diodes. The voltage amplification is performed by operational amplifiers operating in a non-inverting configuration. Finally, the last stage consists of 22 capacitors, each connected to one of the DC voltage lines. These capacitors block the RF component that may enter to the bias circuit from the AAMC. The Varactor Diodes will be inverse polarized in order to control the capacitance through the voltage applied and its connection is described as follows. The first Varactor Diode (left to right), illustrated in Fig. 12b, is polarized connecting the ground plane to the first vias and the +V to the next vias, which will be isolated and do not touch the GND. The second Varactor Diode is inverted with respect to the first in order to use the second vias to connect it to +V; the third vias is connected to the GND. The next Varactor Diodes will be polarized using the same procedure and each of them will have the same voltage due to the parallel connections.

IV. Beam Steering Analysis

To understand the beam steering application offered by an AAMC, a mathematical insight of its phase reflection, as well as a progressive phase shift, is described.

First, A point source generates an electromagnetic wave, which is separated a distance \( d \) from the AAMC. The vertically polarized E-Field incident to the AAMC can be represented by (10).

\[
\vec{E}_i = E_0 e^{-jkd} \hat{y}
\]  

(10)

where, \( E_0 \) is the magnitude of the E-Field, and \( k \) is the wavenumber. A reflected E-Field due to the incidence in the AAMC is generated and can be expressed by (11).

\[
\vec{E}_r = \Gamma(x, y) E_0 e^{-jkd} \hat{y}
\]  

(11)

The reflection coefficient \( \Gamma(x, y) \) is a function of the position in the case of a tunable surface. The reflection coefficient can
be expressed as function of the impedance by (12), which is also function of position.

\[
\Gamma(x, y) = \frac{Z_s(x, y) - \eta_0}{Z_s(x, y) + \eta_0} \tag{12}
\]

In this expression, \(Z_s(x, y)\) is the surface impedance at different positions. Combining (12) and (11) we obtain (13).

\[
\vec{E_r} = \frac{Z_s(x, y) - Z_0}{Z_s(x, y) + Z_0} E_0 e^{\frac{-jkd}{\eta_0}} \hat{y} \tag{13}
\]

If the impedance of the surface is \(Z_s \ll Z_0\) the reflected E-Field will have a phase shift of \(\pi\), reflecting out of phase. This case is known when an incoming electromagnetic wave is incident to a PEC. On the other hand, if \(Z_s \gg Z_0\) the reflected E-Field will have a zero phase shift. This means that all the electromagnetic waves incident to the surface will be reflected with a zero phase shift. This case is known when an incoming electromagnetic wave is incident to a HIS. We can work (12) to obtain an expression that englobe the reflection phase in one term, as seen in (14);

\[
\vec{E_r} = E_0 e^{\frac{-j2\tan^{-1}\left(\frac{|Z_s(x, y)|}{Z_0}\right)} + k\theta - \pi} \hat{y} \tag{14}
\]

thus, if the reflector is a HIS, the argument of \(\tan^{-1}\) will be near \(\pi\), which leave us with the other term \(kd\). If \(d \ll \lambda\) this term is almost negligible and the expression will have a zero reflection phase. This means that an antenna could be placed very close to the surface without shorting it out. The reflection phase, as a function of the surface position, is illustrated in (15).

\[
\Phi(x, y) = -j2\tan^{-1}\left(\frac{Z_s(x, y)}{Z_0}\right) + k\theta - \pi \tag{15}
\]

The previous equation states that the phase of the reflected electromagnetic wave and, thus, the steering of the beam can be controlled by changing the impedance of the surface, in other words, it can be controlled by a phase gradient, in both \(\hat{x}\) and \(\hat{y}\) directions [4].

In order to derive the values of capacitance for the Varactor Diodes, first we analyze the case of a planar wave front. Fig. 13 represents the AAMC, and the position of each element and the reflection angles in azimuth and elevation. The phase gradients that can be generated in the previously presented structure are defined by Equations (16) to (18).

\[
\Delta \psi(x_i, y_i) = -k \sin \theta_i (d_x \cos \phi_x \hat{x}_i - d_y \sin \phi_x \hat{y}_i) \tag{16}
\]

\[
\psi(x_i, y_i) = -\pi + \Delta \psi(x_i, y_i) / N, \quad i = 0, 1, 2, \ldots, M - 1 \tag{17}
\]

\[
\psi(x_i, y_i) = -\pi + \Delta \psi(x_i, y_i) / N, \quad i = 0, 1, 2, \ldots, N - 1 \tag{18}
\]

where; \(\Delta \psi(x_i, y_i)\) are the phase gradient in the \(\hat{x}\) and \(\hat{y}\) directions, respectively; \(k\) is the wavenumber in free space; \(\theta_i\) and \(\phi_x\) are the steering angles; finally, \(N\) and \(M\) are the number of elements in the \(x\) and \(y\) directions, respectively. Once we have the phase gradient, we can use Fig. 9 in order to attain the equivalent capacitance. As an example, if our AAMC has 8 elements, with: \(p = 10\) mm, \(\phi_x = 40^\circ\), \(\theta_i = 0^\circ\), and \(f = 2\) GHz; then \(\psi = [-180^\circ - 164.6^\circ - 149.1^\circ - 133.7^\circ - 118.3^\circ]\)

In addition to the formula (16), we need to add the term that represents the distance between the reflective surface and the phase center of the radiating element. This term is presented in 19 as follows:

\[
\Delta \psi(x_i, y_i) = -k (\sin \theta_i (d_x \cos \phi_x \hat{x}_i - d_y \sin \phi_x \hat{y}_i) + r_i) \tag{19}
\]

where, \(r_i\) is the distance from the phase center of the dipole. Finally, since the wave front of the dipole is not flat due to its location in the near field zone, the formulas will not be suitable for an exact calculation of the phase gradient. But we can take it as a starting point in order to find the appropriate values that generate the desired steering through an optimization process. The optimization process was set in order to vary the capacitance values of each column, to obtain steerings starting from \(\theta = 0^\circ\) to \(\theta = 40^\circ\). To obtain a negative steering angle the previously calculated values need to be placed starting in the last column. The values of the capacitance of each column are presented in Table 2.

V. MEASUREMENTS
A. S_{11} MEASUREMENTS
1) ANTENNA WITHOUT AAMC REFLECTOR
The coupling of the dipole antenna is measured without the AAMC reflector, which is illustrated in Fig. 14. As can be seen, the results between measurement and simulation are quite consistent, except for a reduction in bandwidth. This reduction is attributed to the manufacturing tolerance of the antenna and the effect of the feeding cable which is not considered in the simulation.
TABLE 2. Capacitance values for column 1 to 22, to generate a beam steering from 0° to 40°. To generate negative steering the column order must be inverted.

| Column | 0° | 10° | 20° | 30° | 40° |
|--------|----|-----|-----|-----|-----|
| 1      | 1.20 pF | 0.95 pF | 0.95 pF | 0.95 pF | 0.83 pF |
| 2      | 1.20 pF | 1.00 pF | 1.00 pF | 1.00 pF | 0.90 pF |
| 3      | 1.20 pF | 1.05 pF | 1.05 pF | 1.05 pF | 0.95 pF |
| 4      | 1.20 pF | 1.10 pF | 1.10 pF | 1.10 pF | 1.00 pF |
| 5      | 1.20 pF | 1.15 pF | 1.15 pF | 1.15 pF | 1.05 pF |
| 6      | 1.20 pF | 1.20 pF | 1.20 pF | 1.20 pF | 1.10 pF |
| 7      | 1.20 pF | 1.25 pF | 1.25 pF | 1.25 pF | 1.15 pF |
| 8      | 1.20 pF | 1.30 pF | 1.30 pF | 1.30 pF | 1.20 pF |
| 9      | 1.20 pF | 1.35 pF | 1.35 pF | 1.35 pF | 1.23 pF |
| 10     | 1.20 pF | 1.40 pF | 1.40 pF | 1.40 pF | 1.30 pF |
| 11     | 1.20 pF | 1.45 pF | 1.45 pF | 1.45 pF | 0.90 pF |
| 12     | 1.20 pF | 1.50 pF | 1.50 pF | 1.50 pF | 0.90 pF |
| 13     | 1.20 pF | 1.55 pF | 1.55 pF | 1.55 pF | 0.90 pF |
| 14     | 1.20 pF | 1.60 pF | 1.60 pF | 1.60 pF | 0.90 pF |
| 15     | 1.20 pF | 1.65 pF | 1.65 pF | 1.65 pF | 0.90 pF |
| 16     | 1.20 pF | 1.70 pF | 1.70 pF | 1.70 pF | 0.90 pF |
| 17     | 1.20 pF | 1.75 pF | 1.75 pF | 1.75 pF | 0.90 pF |
| 18     | 1.20 pF | 2.00 pF | 2.00 pF | 2.00 pF | 0.90 pF |
| 19     | 1.20 pF | 2.00 pF | 2.00 pF | 2.00 pF | 0.90 pF |
| 20     | 1.20 pF | 2.00 pF | 2.00 pF | 2.00 pF | 0.90 pF |
| 21     | 1.20 pF | 2.00 pF | 2.00 pF | 2.00 pF | 0.90 pF |
| 22     | 1.20 pF | 2.00 pF | 2.00 pF | 2.00 pF | 0.90 pF |

FIGURE 14. Simulated and measured reflection coefficient of a printed dipole antenna without the AAMC reflector. The antenna have the dimensions described in Table 1 and is feed using a 50 Ohm line.

2) ANTENNA OVER THE AAMC REFLECTOR

The AAMC reflector is measured in a spherical anechoic chamber, as illustrated in Fig. 15. The surface is clamped using the ground plane, which is isolated from the metal contacts of the bracket using plastic washers. Power and control of the AAMC is provided outside the anechoic chamber via a serial connector.

The dipole is measured in vertical polarization over the AAMC reflector at a distance \( r \), as illustrated in Fig. 11. The simulated and measured \( S_{11} \) parameter for different steering angles are illustrated in Fig. 16. The simulation and measurements shows that the dipole still resonate at 2 GHz for all the steering angles. A reduction in the coupling with respect to the case when the dipole is without the reflective surface is observed, which is of around -5 dBi for the measured case. This reduction is generated due to the narrow distance between the dipole and the reflective surface. There is a small discrepancy between the results with regard to the bandwidth, which can be attributed to manufacturing tolerances and the feeding cable.

3) RADIATION PATTERN

The Full Structure radiation pattern, measured and simulated at 0°, 10°, 20°, 30°, and 40° are presented in Fig. 17- Fig. 21, respectively. The negative beam steerings are not illustrated due to the symmetry of the results. The DC feeding values are controlled outside the anechoic chamber and are set after the finish of each measurement.
FIGURE 17. Measurements and simulations of the co-polarized and cross-polarized realized gain of the complete structure. The Varactor Diodes are polarized in order to generate a phase gradient that steers the main lobe an angle of 0°.

FIGURE 18. Measurements and simulations of the co-polarized and cross-polarized realized gain of the complete structure. The Varactor Diodes are polarized in order to generate a phase gradient that steers the main lobe an angle of 10°.

FIGURE 19. Measurements and simulations of the co-polarized and cross-polarized realized gain of the complete structure. The Varactor Diodes are polarized in order to generate a phase gradient that steers the main lobe an angle of 20°.

FIGURE 20. Measurements and simulations of the co-polarized and cross-polarized realized gain of the complete structure. The Varactor Diodes are polarized in order to generate a phase gradient that steers the main lobe an angle of 30°.

As expected, when the dipole is placed in front of a reflector it loses its omnidirectional radiation pattern, redirecting and concentrating it mainly in the backscatter direction. On the other hand, its realized gain increases, in this case from a value of 1.7 dBi to around 7 dBi in its main lobe at 0° of steering. As the steering is being generated, it can also be seen that the main lobe has a lower realized gain, reaching a value of approximately 6 dBi at 40° of steering. From 0° and 30° steering the simulated and measurement co-polarized results have acceptable agreement in the main lobe. The steering is more visible after the 20° result, because of the non-directive beam of the dipole. The back lobe results differ between simulation and measurements. This variation is due to the absorbent material and the circuitry placed at the back of the structure which extends the perfil of the antenna around 10 cm. In all the results there is a notch in the main lobe, which is generated by the feeding cable of the dipole. The cross-polarized results show less agreement specially at 20°, 30°, and 40°. This can be explained due to soldering and encapsulated bumps generated due to the Varactor Diodes and vias, which modifies the surface flatness making it ribbed.

VI. GAIN ENHANCEMENT

There are several options to increase the gain of our design, for example, replacing the dipole with a more directive antenna such as a horn, patch, Vivaldi or an antenna array.
FIGURE 21. Measurements and simulations of the co-polarized and cross-polarized realized gain of the complete structure. The Varactor Diodes are polarized in order to generate a phase gradient that steers the main lobe an angle of 40°.

FIGURE 22. Realized gain simulation from −40° to 40° of an array of 4 dipoles placed over an AAMC reflector. The dipoles are distributed in the z-axis, separated form each other a distance $\lambda_0/4$.

From all the options described above, the only one that allows keeping the low profile of the structure and avoid shadowing is using a dipole array. To elucidate the gain increase by using an array, we have arranged 4 dipoles placed along the z-axis separated a distance of $\lambda_0/4$ from each other. The simulation was carried out without modifying the structure, instead we have used the array factor tool of CST Microwave Studio. Fig. 22, illustrates the simulations of our design at different steering angles. As can be seen, there is a gain increase of about 6 dBi for each angle. To have a clear image of the gain increase, Fig. 23 illustrates, for different types of array elements, the steering at +40°. By simple inspection, as the number of array elements increases the gain increases and the main lobe narrows in the H-plane, as expected.

FIGURE 23. Radiation pattern simulation results for an angle of 40° using the proposed structure. The number of dipoles used are from one and up to 8 elements.

VII. CONCLUSION

An AAMC has been proposed as a reflective surface that allows to modify the radiation pattern of a printed dipole located at a distance of $\lambda_0/4$ above the surface. The beam steering of the printed dipole is generated by adjusting the capacitance of each Varactor Diode through the variation of the potential of the power supply network. This variation in capacitance generates a progressive phase shift throughout the structure, which steers the radiation pattern of an impinging electromagnetic wave. Additionally, an equivalent circuit model for an AAMC is proposed. The validation of the effectiveness of the proposed equivalent circuit has been performed by comparing the unit cell S-Parameters obtained by electromagnetic simulation of the unit cell, with the simulation of the proposed equivalent circuit in Advance Design System (ADS). One advantage of modeling an AAMC through an equivalent circuit is to understand the relationship between resonance frequency, bandwidth, and admittance with the physical elements of the unit cell, such as thick of the substrate(s), resonant element shape, dimensions, and even the effect of lumped elements. The full structure has been tested and measured in an anechoic chamber using the spherical system, where the measurements obtained are in good agreement with simulations, thus validating our work.
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