Design of Triangular-Latticed Subarray Antenna Fed by Hexagonal Radial Line for K-band Applications

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Abstract — A circularly polarized double-stacked patch subarray antenna fed by a hexagonal radial line with internal circular coupling patches is proposed. The antenna works at K band (19.7 GHz – 20.2 GHz) with LHCP. On the one hand, the design approach consists of a coupling study based on the simulation of a periodic parallel-plate waveguide as a first approximation of the radial line. On the other hand, a phase compensation method by rotation is applied in order to adjust the radiating phase of each element of the subarray, which are separated 0.7λ0. Thus, the subarray is uniformly fed in terms of amplitude and phase. An axial ratio below 1 dB is achieved for the entire frequency band, with a gain of 24 dB and a total efficiency of 87%. The radial line has been modelled as a hexagon to perform this analysis in larger arrays in future design steps.

Keywords — Radial Line Antenna, Triangular Lattice, Double-Stacked Microstrip Patch, Internal Circular Coupling Patch, Phase Compensation Rotation

I. INTRODUCTION

Over the last decades Parallel-Plate Waveguide (PP-WG) technology has been widely studied for array antenna feeding schemes due to their inherent low-profile property [1]. If their height is properly defined below λc/2, a low-loss TEM mode is propagated inside the PP-WG, leading to a radiation efficiency and gain enhancement.

Besides, different manners to propagate the TEM mode are possible within this technology. One possibility is to generate a planar wavefront inside de PP-WG. There are several ways to perform planar wavefronts in PP-WG, such as coupling windows fed in turn by rectangular WG’s in the same layer [2] or from the layer below [3]. Other possibility is to generate a cylindrical wavefront inside the PP-WG. The means to perform this are considerable and usually imply a coaxial input structure whose core penetrates inside one of the plates. This latter kind is usually called Radial Line (RL) in literature.

RL’s have been widely used to feed arrays with different kind of radiating elements such as slots. A lot of studies have been carried out with this structure (Radial Line Slot Antenna or RLSA), from spiral [4], [5] to circular [6], [7] disposal of the slots, in addition to the way the input coaxial feeding structure is used [1], [4], [5], [6], [7]. Nevertheless, slots perform the coupling and the radiating role at once. This leads to a very constraining design scheme in which the feeding coefficients are very difficult to control.

Recent works proposed the use of microstrip patches (Radial Line Patch Antenna or RLPA) fed by internal coupling patches and connected by microstrip compensation lines to perform a more agile design scheme to control the feeding coefficients and the polarization performance [7], [8]. However, the RLPA in [8] consists of a circular RL that feeds several concentric circular arrays. This may suppose a significant bandwidth reduction due to phase errors when the size of the RLPA is increased, even if the design is non-resonant.

To overcome this problem, this paper presents the design of a non-resonant Hexagonal RLPA (H-RLPA) that is suitable to be integrated in larger arrays fed by wideband feeding networks. The Radiating Elements (RE) of the subarray are disposed in a triangular lattice that fits better on a hexagonal surface. A coupling study and a phase compensation process by rotation of the RE are applied to set a uniform feeding scheme of the subarray. The developed work is appropriate for K-band applications such as satellite communications, in which the classical reflector antenna offers a more voluminous and less efficient solution.

II. SUBARRAY ANTENNA ARCHITECTURE

The proposed antenna structure is depicted in Figure 1(a). It consists of 90 RE’s placed in a triangular lattice. Besides, this lattice is sub-divided in 5 concentric hexagonal rings in a non-resonant array design that is short-circuited at the end. Separation between elements is of R = 0.7λ0, at the central frequency f= 19.95GHz. The subarray is designed to operate from 19.7GHz to 20.2GHz.

Figure 1(b) shows the layer scheme of the antenna. The radiating elements are fed by the RL defined below through an internal circular coupling patch, that is calculated in order to perform a uniform distribution of the signal over the entire array. The radiating element is connected to the coupling patch through metallic pins of radii pin r = 0.2mm, and an aperture surrounding it of radii ap r = 0.55mm. The height of the RL is h1+h2 = 1mm+0.508mm = 1.508mm, that is below half the wavelength inside the RL at the fc, to do not generate high-order modes apart from the desired TEM mode. The coupling patch substrate is stacked to the active patch substrate with a Rogers RO4450B prepreg (εr = 3.3, tanδ=0.04) of h3 = 0.091
mm, the active and passive patch substrates are $h_5 = h_7 = 0.203\text{mm}$ respectively, separated by a layer of foam material of $h_6 = 1\text{mm}$. Coupling, active, and passive substrates are Rogers RO4003C ($\varepsilon_r = 3.38$, $\tan\delta=0.027$). Copper layers are all $h_4 = 36\mu\text{m}$ height.

In Figure 1(c) the RE structure is detailed. It consists of a stacked microstrip patch structure. The active patch is symmetrically cut on the edges to achieve good polarization purity. The cut length is $cut \ell = 0.4\text{mm}$ and the radii of the active patch is $active \ r = 2.3\text{mm}$. Equivalently, passive patch radii is $passive \ r = 2.645\text{mm}$. The metallic pin that connects the active patch with the coupling patch is separated from the centre of the active patch $probe_{sep} = 1.6\text{mm}$. The coupling factor is regulated by the separation between the metallic pin and the centre of the coupling patch $coupl_{sep}$, and the radii of the coupling patch $coupl \ r$. All the different coupling patches are orientated to the propagation direction of the TEM mode, with the coupling patches moved backwards from the metallic pin.

Phase errors are inherent in this kind of structures, so a phase compensation method is applied to maintain antenna performance. It consists of a rotation of the elements with respect to the active/passive patch axis. Thus, a partial sequential rotation is set in the antenna feeding scheme. This way, polarization purity is enhanced.

In the rest of the paper the coupling analysis (Chapter III) and the phase compensation method (Chapter IV) are discussed. Once applied, the simulation results of the final subarray prototype are presented in Chapter V. Chapter VI adds future work to be done to enhance the proposed subarray performance.

### III. COUPLING ANALYSIS

As mentioned before, the RE’s of the subarray are uniformly fed from the RL through the coupling structure. Therefore, a first approximation of the RL is assumed in order to calculate the coupling patches dimensions per hexagonal ring. This approximation consists on simulating coupling structure ended in a coaxial port inside an infinite parallel plate waveguide. This is only true for the elements that are placed far enough from the input port, but as a first design step it is faster to simulate (see Figure 2(a)).

Figure 2(b) shows the considered effects that are taken into account for the calculation. This process is used and explained in detail in [8], and only assumes an outgoing wave and not any possible reflected ingoing waves. Thus, if losses are neglected in the analysis, the coupled power to a RE of the $k$-th hexagonal ring is completely radiated ($P_{rad,k}$). In the present design it is also assumed that all the coupling patches are at the same distance from the centre of the subarray. This is also not true but again, as a first design step, it will work.

The applied equations (1), (2), (3) appear in the analysis performed by authors in [8]. It considers that the incident power to the $(k+1)$th ring $P_{inc,k+1}$ is determined by the incident power of the previous ring $P_{inc,k}$ and the radiated power of the same ring $P_{rad,k}$ (1).

Figure 2. Coupling analysis process: (a) infinite parallel-plate waveguide approximation, (b) coupled power in the proposed coupling structure, (c) internal circular coupling patches dimensions vs. theoretical and simulated transmission and reflection parameters.
Figure 2(c) shows the coupling patch dimensions per ring and the theoretical and simulated results for the transmission and reflection parameters of the structure depicted in Figure 2(a). It can be deduced that the bigger the coupling structure is, the more reflective it becomes.

IV. PHASE COMPENSATION METHOD

In order to reduce possible phase errors in the feeding coefficients of the subarray, it is necessary to compensate the phase differences between patches that are uniformly separated. This compensation is performed by rotating the RE’s along the active/passive patch axis until the radiated phase of each element is the same. It is a trigonometric process and it only requires a reference element to start with and the position of each RE of the subarray.

Figure 3 shows the iterative process that every patch has followed in order to compensate phase errors in the feeding network. The first step is to place the RE in the correct position, by rotating it β degrees and by displacing from r₁ to r₂ position. Equation (4) calculates the rotation angle α needed to compensate the effect of rotating and displacing the RE’s. Once the RE is rotated α degrees, it is necessary to reorient the coupling patch to the centre of the subarray, so another rotation of the coupling patch of γ degrees is necessary.

\[ -k \cdot r_1 = -k \cdot r_2 + \beta \rightarrow -k \cdot (R + \text{probe sep}) = -k \cdot \sqrt{r_2^2 + \text{probe sep}^2} + 2 \cdot r_2 \cdot \text{probe sep} \cdot \cos \alpha + \alpha - \beta \]  

Where k is the propagation constant of the RL.

This fact forces to the coupling patches of the elements of the last ring to be not equally spaced from the short-circuit. Then an intermediate solution that is able to be matched for every possible position of the RE’s in the last ring is performed. As seen in Figure 1(a), the short-circuit has an additional step that enhances the matched load behaviour (short \( l_1 = 0.5 \) mm, short \( w_1 = 7.75 \) mm, short \( l_2 = 1 \) mm, short \( 2 \text{el} \text{ sep} = 9 \) mm).

V. SIMULATION RESULTS

Once the coupling structure is fully characterized and the phase compensation method is applied, a full wave simulation of the proposed subarray antenna is performed.

Figure 4 shows the main simulation results of the subarray. The input reflection parameter is below -15dB in the defined frequency band, and it is below -10dB from 19GHz to 21GHz (Figure 4(a)). The radiation pattern depicted in Figure 4(b) presents a good behaviour with a maximum directivity of 24.8dBi and with a reasonable agreement with the theoretical uniform feeding distribution due to the symmetry of the second sidelobe. Figure 4(c) depicts the axial ratio of the antenna for the 4 principle cuts of the radiation pattern (\( \phi = 0^\circ, 45^\circ, 90^\circ, 135^\circ \)) for the lower, higher and central frequencies (\( f_L, f_H, f_C \)). It becomes evident that the proposed phase compensation method within this paper is enhancing the polarization behaviour of the subarray, since this method applies a partial sequential rotation to the radiating elements.

In Figure 4(d) it is depicted the frequency behaviour of the

\[ P_{\text{inc},k+1} = P_{\text{inc},k} - P_{\text{rad}} \]  

Under these assumptions, the complete radiated power can be considered as the same value of power that is introduced to the antenna. The parameter \( t \) takes into account the residual power that impinges the matched load.

\[ P_{\text{input}} = P_{\text{rad}} = t + \sum_{i=1}^{M_k} P_{\text{rad},k} \]  

As in [8], the coupling coefficients per element and per ring \( |S_{0,k}| \) are calculated with (3) to generate the desired amplitude delivery \( F_{0,k} \). In this paper (like in [8]), a non-resonant design approach is set, so the \( t \) parameter is also put to zero. Therefore, the last ring of the design is a matched load. Unfortunately, the rotating process forces to design a matched load that is adapted for any possible position of the patches.

\[ |S_{0,k}| = \frac{P_{\text{rad},k}}{P_{\text{inc},k}} = \frac{\sum_{i=1}^{M_k} P_{\text{rad},k}(\phi_i)}{\sum_{i=1}^{M_k} N_{\text{elem},k} P_{\text{rad},k}(\phi_i) + \sum_{i=1}^{M_k} N_{\text{elem},k} P_{\text{rad},k}(\phi_i)} \]  

Figure 3 shows the step-by-step phase compensation method.
subarray axial ratio at the boresight along with its directivity, gain and efficiency. The axial ratio does not seem to be degraded with the frequency. The subarray exhibits a good behaviour in terms of efficiency (almost 90% of total efficiency in the whole frequency band and with a very constant tendency) and gain, with a maximum value of 24.5 dB).

VI. CONCLUSION AND FURTHER RESEARCH

In this paper, a triangular-latticed hexagonal subarray antenna has been presented. The proposed coupling analysis and phase compensation method are able to generate a good behaviour in terms of efficiency (up to 90%) and axial ratio (below 0.5 dB for the boresight in the whole frequency band). Fabrication of the proposed subarray is the next research step to validate the design workflow. Different technologies may be involved such as classic CNC, 3D printing and copper coating, DMLS and PCB fabrication techniques.

It is also proposed a bigger array in which a corporate waveguide feeding network feeds some of the proposed subarrays. The input port may be changed for a TM${}_{01}$-TE${}_{10}$ circular to rectangular waveguide transition to feed properly the RL and enhance the efficiency of a bigger array due to waveguide corporate feeding network.

![Simulated Radiation Subarray Parameters](image)

Figure 4. Subarray principle results: (a) input reflection parameter, (b) 3D radiation pattern at the central frequency, (c) axial ratio for the 4 principle cuts at the extreme and central frequencies, (d) directivity, gain, efficiencies and axial ratio at broadside vs frequency.

ACKNOWLEDGMENT

Simulations done in this work have been performed using CST Microwave Studio Suite 2018 under a cooperation agreement with Computer Simulation Technology (CST). This work was supported in part by the Spanish Government, Ministry of Economy, National Program of Research Development and Innovation under the projects ENABLING-5G “ENABLEING INNOVATIVE RADIO TECHNOLOGIES FOR 5G NETWORKS” (project number TEC2014-55735-C3-1-R); FUTURE-RADIO “Radio systems and technologies for high capacity terrestrial and satellite communications in an hyperconnected work” (project number TEC2017-85529-C3-1-R) and the Madrid Regional Government under the project SPADERADAR “Space Debris Radar” (S2013/ICE-3000).

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