Non-homogeneous Fabry-Pérot Antenna Design Process to Improve Aperture Efficiency

Pablo Mateos-Ruiz, Elena Abdo-Sánchez, Carlos Camacho-Peñalosa
Instituto de Telecomunicación (TELMA), Universidad de Málaga - Andalucía Tech, 29010 Málaga, Spain
{pablomr, elenaabdo, ccp}@ic.uma.es

***Abstract*—A novel technique to design Fabry-Pérot antennas with non-homogeneous partially reflective surfaces (PRSs) is described. It uses a transmission line circuit model to efficiently obtain all necessary unit cell designs that satisfy the cavity resonance condition. This method allows an increase in directivity without reducing the bandwidth for a given footprint. Some design examples in the Ku-band are presented, showing the evolution from a simple single-layer PRS to a non-homogeneous two-layer one. The latter achieves an increase of about 3 dB in directivity while maintaining the bandwidth in electromagnetic simulations. This way, the gain-bandwidth product is improved from a value of 5 to almost 9, effectively raising the antenna efficiency.**

***Index Terms*—Fabry-Pérot Antenna, Partially Reflective Surface, Circuit modelling, Wideband, Non-homogeneous**

I. INTRODUCTION

In 1956, Trentini [1] explored the capabilities of a new type of antennas based on Fabry-Pérot resonators making use of the ray theory. From then on, these Fabry-Pérot Cavity Antennas (FPCAs) have gained popularity as a cost-effective technique to increase the directivity of a primary radiation source. This source is placed between two parallel surfaces, one being a ground plane and the other a Partially Reflective Surface (PRS). These antennas have the advantages of being passive, low profile, and highly directive, only needing a single feeding point. However, one of their main drawbacks is their narrow bandwidth, a problem which has become the main focus of the research related to FPCAs [2], [3].

Another challenge of these antennas is their lateral size. As the PRS is more reflective, more electromagnetic field is diffracted at the antenna edges. Consequently, a larger antenna is commonly needed to prevent that increase in lateral radiation, which implies a lower directivity. However, once the antenna lateral size is fixed, the PRS characteristics can be varied throughout its length in order to better adequate its properties to the field density underneath it, and consequently increasing the antenna efficiency. For example, in [4], a certain degree of heterogeneity is introduced in the PRS design, combining two unit cells with slightly different responses to obtain a more desirable overall performance.

In this article, the PRS non-homogeneity is explored in greater depth as a technique to both increase bandwidth and directivity. The goal is to better utilize the available area and achieve a higher antenna efficiency. First, a brief overview on wide-band PRS unit cell design and modelling is provided. An homogeneous PRS design is then carried out to contrast the procedure with a non-homogeneous (NH) design employing a novel technique based on the previously explained circuit model. Finally, these designs are simulated and compared to ascertain the improvement in the last one. A similar work was already presented in a Spanish symposium [5].

II. PRS DESIGN

From the simplified ray model, it is derived that, essentially, for a FPCA to work at a desired frequency, the PRS reflection phase ϕ must satisfy the resonance condition given by [1] at that wavelength λ:

$$\phi + \pi - \frac{4\pi}{\lambda} D = 0 \tag{1}$$

being D the cavity height, and assuming that the other reflective surface is a PEC. In addition, the PRS reflection magnitude governs the resonance intensity which, in turn, is proportional to the directivity of the antenna. On this basis, a simple design can be carried out with any PRS design that satisfies (1) for a single frequency.

However, a common way of improving the antenna bandwidth is to optimise the PRS to better satisfy the resonance condition around the design frequency. As proposed in [6], (1) can be rearranged, obtaining the ideal PRS reflection phase to have maximum directivity within a certain frequency range:

$$\phi = \frac{4\pi D}{c} f + (2k-1)\pi, \quad k \in \mathbb{Z}. \tag{2}$$

In this sense, achieving a positive-gradient reflection phase usually requires the PRS structure to have a (partial) resonance around the design frequency [2]. A simple circuit model is usually used to design the PRS frequency response, for which at least two degrees of freedom are needed to obtain the required resonance. In this case, a multi-layer approach is chosen as it allows a simple design of each layer while facilitating a comprehensive transmission line modelling of the structure.

III. TRANSMISSION LINE MODEL

A well-known transmission line (TX) model is used to model the multi-layer PRS and synthesize its desired frequency response. Each PRS layer is modelled as an impedance in parallel, while the spaces between them are modelled with transmission lines characterised by the length of the gap between both layers and the permittivity of the dielectric used.

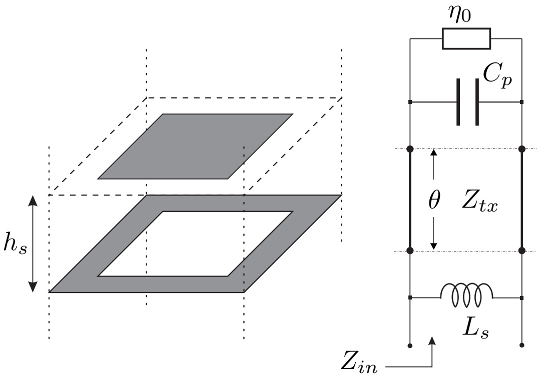


Fig. 1. Patch-Slot unit cell and equivalent Transmission Line model.

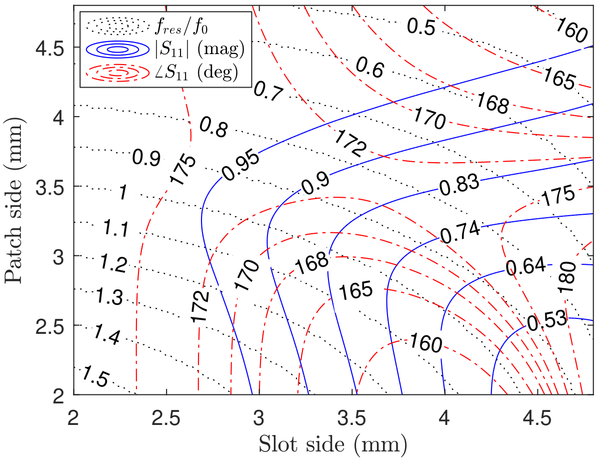


Fig. 2. Reflection magnitude and phase for a mesh of possible PRS designs at $f_0 = 14\,\text{GHz}$ [5].

The vacuum intrinsic impedance must be added after the last layer impedance to account for the air after it. This way, the total reflection coefficient can be obtained from the input impedance of the circuit.

In order to illustrate the design procedure, two PRS layers are selected. The inner layer is composed of squared slots, which can be modelled as an inductance L_s, and the outer layer is made of squared patches, modelled as a capacitance C_p [3], as shown in Fig. 1. Both layers have a periodicity of $5\,\text{mm}$. The FPCA design is carried out around $14\,\text{GHz}$ using a RO4350B dielectric sheet where each layer can be assembled from one of the metallized sides. The chosen dielectric has a permittivity of $\varepsilon_r = 3.66$ and a thickness of $h_s = 60\,\text{mils}$.

The synthesizable values of C_p and L_s can be obtained from simulations varying the sides of the PRS layers d_p and d_s, respectively. Using these reactances in the analytical expression of the input impedance Z_{in} from Fig. 1, the frequency response of every possible design can be obtained. From them, the reflection magnitude and phase values at the desired frequency (14 GHz) can be extracted and plotted in a visually comprehensive way using an isolines diagram, shown in Fig. 2. In addition, for each design, its normalized resonance frequency (minimum reflection coefficient) respect to the desired frequency is also represented, as it helps understanding how close to the positive phase gradient each design is.

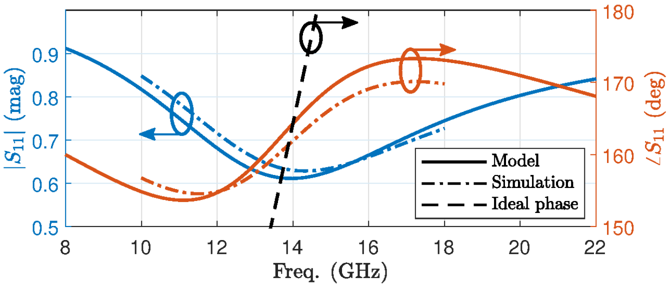


Fig. 3. Reflection coefficient for the designed homogeneous PRS.

IV. DESIGN PROCEDURE

A. *Homogeneous design*

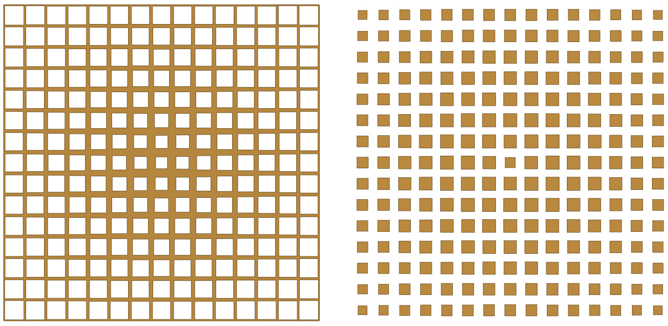
A PRS reflection magnitude of 0.611 is selected first for this example design. From Fig. 2, the required design parameters can be extracted from the intersection of the corresponding reflection magnitude isoline and the unit-value normalized-resonance-frequency one, thus maximizing bandwidth. In addition, the reflection phase value ($\phi = 164.3°$) is directly obtained and, using (1), the needed cavity height $D = 10.24\,\text{mm}$ is computed.

The frequency response is represented in Fig. 3, where the ideal phase from (2) is also illustrated with a dashed line (Fig. 3). In addition, the results from the electromagnetic simulation of this design are also shown, proving that the model is accurate enough. The slight discrepancies between the model and the simulations can be explained by the unmodelled layers parasitic effects, coupling effects between the layers and numerical errors from the simulation. Nevertheless, it can be perfectly used as a base for these designs.

B. *Non-homogeneous design*

As a type of leaky-wave antenna, FPCAs share the characteristic of a decaying electromagnetic field along its radiant elements. This effect causes a non-uniform PRS illumination, which translates into a lower aperture efficiency. To solve this problem, a similar approach to the increasing leakage factor of leaky-wave antennas can be followed for FPCAs. The resonance condition only depends on the PRS reflection coefficient phase (1) for a fixed cavity height, while from Fig. 2 it is derived that different reflection magnitudes can be obtained along the same phase isoline. This way, the resonance condition can be met while radially decreasing the reflection coefficient of the PRS. Consequently, the transmission through it would increase to compensate for the field decay, achieving a more homogeneous PRS illumination.

A NH design is carried out following the previous guidelines. It presents reflection coefficients whose magnitude decays linearly with the distance from the PRS center ($s_{11} = 0.95$) to its corner unit cells ($s_{11} = 0.45$). In this case, the used phase isoline is 170°, since its unit cells resonance frequencies



(a) Slots layer.

(b) Patchs layer.

Fig. 4. Designed layers of the Non-homogeneous Patch-Slots PRS.

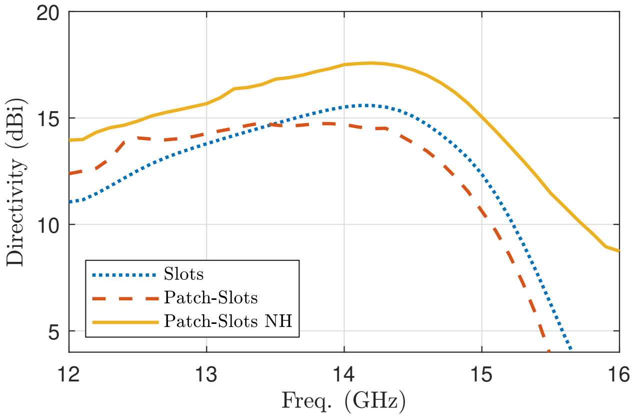


Fig. 5. Simulated directivity at broadside ($\theta = 0°$) for the different designs.

are relatively close to 14 GHz (Fig. 2). Thus, the different PRS designs will be better centered in the positive phase gradient areas, maintaining the bandwidth increase effect. Both PRS layers, with 15 unit cells of lateral size, are shown in Fig. 4.

V. SIMULATION RESULTS

First of all, a slots-only PRS is also designed with a reflection coefficient magnitude of 0.61 to have a base for the comparison. All the described designs are simulated in HFSS using a simple horizontal half-wavelength dipole as the primary source inside the cavity.

The results for the broadside directivity are presented in Fig. 5. All designs effectively have their maximum directivity around 14 GHz. In addition, a clear evolution can be observed from the simpler slots-only to the NH design. Qualitatively, the homogeneous patch-slots design achieves a greater bandwidth (extends below 12 GHz) than the slots-only one, having a more constant directivity over a wider frequency range, thanks to the positive phase gradient technique. The increment in directivity is significant when the non-homogeneity is incorporated, taking into account that the bandwidth is even slightly bigger than the one of the slots-only design.

Some specific figures of merit are shown in Table I such as aperture efficiency (ϵ_{ap}), half-power bandwidth (BW), fractional bandwidth (FBW) and gain-bandwidth product (GBP), defined as the FBW multiplied by its average directivity. The

TABLE I. FIGURES OF MERIT OF THE DIFFERENT DESIGNS.

	Slots	**Patch-Slots**	**Patch-Slots NH**
$\boldsymbol{\epsilon_{ap}}$	23.5 %	19.5 %	37.2 %
BW	2.45 GHz	$>$ 2.87 GHz	2.74 GHz
FBW	17.8 %	$>$ 21.4 %	20.0 %
GBP	5.05	$>$ 5.44	8.85

improvement in the GBP of the NH design is noticeable, as a higher directivity has been achieved without worsening the BW. Although this design could achieve even greater directivity values making the PRS bigger, the main focus in this study is how to exploit the same PRS size to obtain higher aperture efficiencies.

VI. CONCLUSIONS

In this contribution, a comprehensive process for designing non-homogeneous FPCAs which increases the aperture efficiency has been presented. The progression from a single layer PRS to a NH multi-layer one has been clearly shown, also demonstrating how the positive phase gradient approach successfully broadens the BW.

In order to obtain greater bandwidths, the unit cell design could be further improved using geometries with higher quality factors. On the other hand, the optimum reflection decay to obtain maximum efficiency is still to be studied. However, a simple linear decay with reasonable bounds has already increased the maximum directivity in almost 3 dB, proving the potential of this technique. As immediate future work, a realistic primary radiation source should be designed and the final prototype should be fabricated and measured.

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