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**Published on:** 02 Aug 2012 - IEEE Transactions on Antennas and Propagation (Institute of Electrical and Electronics Engineers (IEEE))

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# Electronically Steerable 1D Fabry-Perot Leaky-Wave Antenna Employing a Tunable High Impedance Surface

Raúl Guzmán-Quirós, *Student Member, IEEE*, José Luis Gómez-Tornero, *Member, IEEE*, Andrew R. Weily, *Member, IEEE*, and Y. Jay Guo, *Senior Member, IEEE*

**Abstract**— A novel fixed-frequency electronically-steerable one dimensional (1D) leaky-wave antenna is presented. The antenna is based on a parallel-plate waveguide loaded with a planar partially reflective surface and a tunable high impedance surface (HIS), which creates a 1D Fabry-Perot leaky-waveguide. The tunable HIS consists of printed patches loaded with varactor diodes that allow the electronic tuning of the cavity resonance condition. Using a simple Transverse Equivalent Network, it is theoretically shown how the variation of the varactors' junction capacitance allows the scanning of the antenna pointing angle from broadside towards the endfire direction at a fixed frequency. Experimental results of an antenna prototype operating at 5.6 GHz are reported, demonstrating that the new reconfigurable leaky-wave antenna can provide electronic beam scanning in an angular range from 9° to 30°.

**Index Terms**— Leaky-wave antennas (LWA), reconfigurable antennas, electronic beam scanning, high impedances surfaces, partially reflective surfaces, electromagnetic band gap structures, frequency selective surfaces.

## I. INTRODUCTION

SCANNING of the main beam direction is an inherent property in one-dimensional Leaky-Wave Antennas (1D LWAs), which is usually achieved by sweeping the frequency of the input microwave signal [1-3]. For many applications, however, operation at a fixed frequency is required [4]. Various technologies and topologies have been proposed in recent decades to create fixed-frequency electronically-steerable 1D LWAs [5-24]. In all cases, an active device is employed to produce the electronically controlled change in the leaky-line boundary conditions, thus altering the leaky mode (LM) complex propagation constant ( $k = \beta - j\alpha$ ) and producing the associated change in its pointing direction (noting  $\sin \theta_{RAD} \approx \beta/k_0$ , where  $\theta_{RAD}$  is measured from the broadside direction) [2]. For instance, p-i-n diodes have been added to dielectric rod LWAs [5,10] or microstrip LWAs [19], obtaining a discrete change in  $\theta_{RAD}$ . Similar discrete changes

are reported when introducing photosensitive switches in dielectric slab LWAs [9], or electronic switches in multi-port microstrip LWAs [20],[21]. Continuous scan of the main beam has been obtained in corrugated ferrite LWAs [6],[7], microstrip LWAs over a ferroelectric substrate [17], photoinduced plasma grating LWAs [8], and microstrip LWAs loaded with photosensitive silicon substrates [14], by respectively using magnetic/electric fields or optical signals to control  $\theta_{RAD}$ . However, the most used active device to electronically control 1D LWAs in the microwave range is the varactor diode, which has been applied to multitude of leaky lines as the slotline [11], the coplanar waveguide [12], the first higher-order mode microstrip line [13], the composite right-left handed microstrip line [15],[16], the microstrip log-periodic line [18], the half-mode microstrip line [22],[24], and the half-mode substrate-integrated waveguide [23].

In this paper, we present a new configuration of reconfigurable 1D LWA to achieve fixed-frequency beam steering, which is based on a 1D Fabry-Perot (FP) leaky-waveguide recently proposed in [25]. In the antenna, varactor diodes are added to the High Impedance Surface (HIS) substrate in order to achieve electronic control of the pointing angle. A simple transverse equivalent network (TEN) is used to analyze the LWA and provide a theoretical foundation for the proposed mechanism to steer the beam. An experimental prototype of the antenna operating at 5.6 GHz was designed and tested. Good agreement between theoretical predictions and experimental results are obtained.

The paper is organized as follows. The working principle of the new reconfigurable 1D LWA is described in Section II. It is shown that, using the transverse equivalent network (TEN), the dispersion of the associated LM (please spell it out as it is the first time this is used) is a function of the varactor junction capacitance. The dispersion curves prove essential for the computer-aided design of the antenna, and particularly for the optimization of the scanning range. All the TEN analysis results are validated with full-wave simulations. Section III reports experimental results obtained from a fabricated prototype operating at 5.6 GHz, showing an electronically-controlled scanning region from  $\theta_{RAD} = 9.2^\circ$  to  $\theta_{RAD} = 34.2^\circ$  as the varactors' biasing DC voltage is varied from 4.5V to 18.2V. Practical limitations of the proposed antenna for scanning angles close to endfire are also discussed. Finally, Section IV summarizes the main conclusions of this work.

Manuscript received November 11, 2011. This work has been supported by Spanish National project TEC2010-21520-C04-04, Spanish Regional Seneca project 08833/PI/08, and Australian project XXXX.

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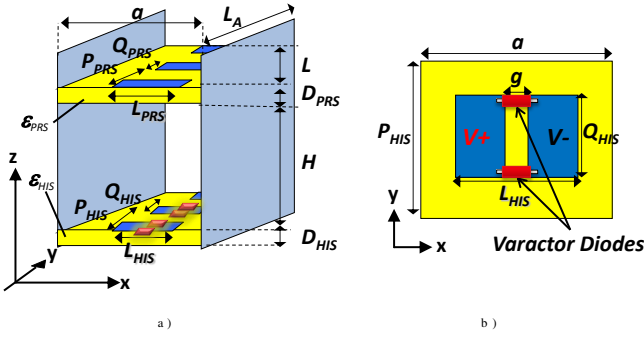


Fig. 1.a) Scheme of the proposed Reconfigurable 1D LWA b) Detail of tunable High Impedance Surface. ( $a=25\text{ mm}$ ,  $H=25.2\text{ mm}$ ,  $L=13.6\text{ mm}$ ,  $D_{HIS}=1.524\text{ mm}$ ,  $\epsilon_{HIS}=3$ ,  $P_{HIS}=30\text{ mm}$ ,  $Q_{HIS}=14\text{ mm}$ ,  $L_{HIS}=17\text{ mm}$ ,  $g=1\text{ mm}$ ,  $D_{PRS}=0.8\text{ mm}$ ,  $\epsilon_{PRS}=4.4$ ,  $P_{PRS}=20\text{ mm}$ ,  $Q_{PRS}=18\text{ mm}$ ,  $L_{PRS}=22\text{ mm}$ )

## II. RECONFIGURABLE 1D LEAKY-WAVE ANTENNA

The proposed configuration of 1D reconfigurable LWA and its main geometrical parameters are shown in Fig.1-a. The structure is inspired by the passive antenna presented in [25], where a parallel-plate waveguide (PPW) was loaded with two passive printed-circuit boards (PCBs): a top Partially Reflecting Surface (PRS) and a bottom High Impedance Surface (HIS) (see Fig.1-a). Both PCBs are formed by periodic arrays of resonant metallic patches printed on thin dielectric substrates. The two PCBs are separated at a distance  $H$ , comprising a one-dimensional Fabry-Perot (FP) resonant cavity which allows the propagation of a perturbed  $TE_{01}$  leaky mode. Since these types of antennas combine a host PPW with PCBs, they are known as hybrid waveguide printed-circuit LWAs [25],[26]. As demonstrated in [25], this 1D FP PRS-HIS LWA allows the leakage rate to be adjusted by changing the resonant length of the PRS patches ( $L_{PRS}$  in Fig.1), while the resonant length of the HIS patches ( $L_{HIS}$  in Fig.1) controls the pointing angle at a fixed design frequency. Using these previous results, a reconfigurable version of this 1D LWA is envisaged by introducing varactor diodes in the middle of the HIS resonant patches, as illustrated in Fig.1-b. Similar types of tunable HIS have been used to design reconfigurable reflectors [27],[28], reconfigurable reflectarrays [29], or tunable Frequency Selective Surfaces [30]. Also, an electronically tunable HIS was applied in [31] to transform a surface-wave into a backward or forward leaky-wave, thus achieving beam scanning from a two-dimensional textured surface. In [32],[33], a tunable HIS was applied to tune the operating frequency of low profile 2D FP LWAs, and similar reconfigurability was demonstrated in [34] for a tunable HIS fed by a bow-tie antenna. To the best of the authors' knowledge, it is the first time that a tunable HIS is applied to the design of a fixed-frequency electronically-scanned one-dimensional Fabry-Perot LWA.

In the following subsection, a simple but accurate TEN is developed to obtain the complex propagation constant of the leaky-mode which operates in the proposed reconfigurable LWA, as a function of its main parameters. These dispersion curves give physical insight into the operating principle and design rules of the proposed LWA, and lay a theoretical foundation for the steerability of the antenna main beam. Analysis of reconfigurable LWA using a TEN

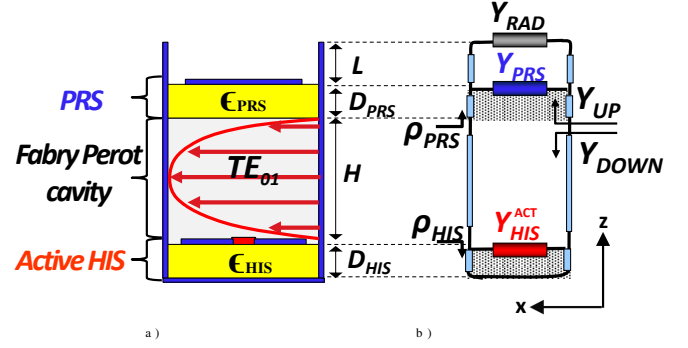


Fig. 2.a) Cross section of the reconfigurable 1D FP-PRS-tunable HIS LWA b) Transverse Equivalent Network (TEN) of the structure.

The Transverse Resonance Method (TRM) is a simple and powerful analysis tool widely used in the design of LWAs [2],[35],[36]. Particularly, as it was demonstrated in [25], the TRM allows efficient analysis and design of the passive version of the 1D FP PRS-HIS LWA. For the analysis of the reconfigurable version of this LWA, it is necessary to extend the TRM approach proposed in [25]. The key feature of the TRM is the development of an appropriate Transverse Equivalent Network (TEN) that accurately models the cross section of the antenna, which in our case is shown in Fig.2. The constituent parts of this TEN are well described in [25], with the most important parts being the equivalent admittances which model the top PRS and the bottom HIS ( $Y_{PRS}$  and  $Y_{HIS}$  in Fig.2-b). In this work, the necessary novelty required to model the tunable HIS admittance as a function of the varactors' junction capacitance  $C_j$  is introduced in the TEN formulation.

As explained in [37], the pole-zero method proposed by Maci et al. in [38] can be expanded to take into account more optimization parameters beyond frequency in the design of 1D LWA. For instance, the effect of the resonating length of the PRS patches ( $L_{PRS}$  in Fig.1-a) can be efficiently modeled by using the following analytical pole-zero expansion for  $Y_{PRS}$  [37]:

$$Y_{PRS}(k_y, L_{PRS}) = j \frac{L_{PRS} [L_{PRS} - L_{PRS_z}(k_y)] \dots [L_{PRS} - L_{PRS_{zn}}(k_y)]}{[L_{PRS} - L_{PRS_p1}(k_y)] \dots [L_{PRS} - L_{PRS_{pn}}(k_y)]} \quad (1)$$

where the location of the poles and the zeros depend on the longitudinal wavenumber  $k_y$ , which is related to the leaky-mode incidence (or radiating) angle by  $\sin \theta_{inc} = \text{real}(k_y)/k_0$ .

Fig.3 illustrates the magnitude and phase of the reflection coefficient presented by the PRS ( $\rho_{PRS}$  in Fig.2-b) as a function of  $L_{PRS}$  for the design frequency of  $5.6\text{ GHz}$ , and for three different incidence angles. As explained in [25],  $L_{PRS}$  must be properly chosen to tune the PRS reflectivity in order to control the antenna radiation efficiency. The rest of the geometrical parameters of the PRS are fixed, so patches of width  $Q_{PRS}=18\text{ mm}$  separated at a distance  $P_{PRS}=20\text{ mm}$ , are printed on FR4 substrate ( $\epsilon_{PRS}=4.4$ ,  $D_{PRS}=0.8\text{ mm}$ ). As shown in Fig.3, good agreement is observed between the analytical pole-zero model of the PRS admittance (1) and full-wave results obtained with commercial Finite Element Method (FEM) solver HFSS.  $L_{PRS}=22\text{ mm}$  is chosen to provide a PRS reflectivity above 0.9 for all scanning angles, as shown in Fig.3. The main novelty of the proposed TEN is in the

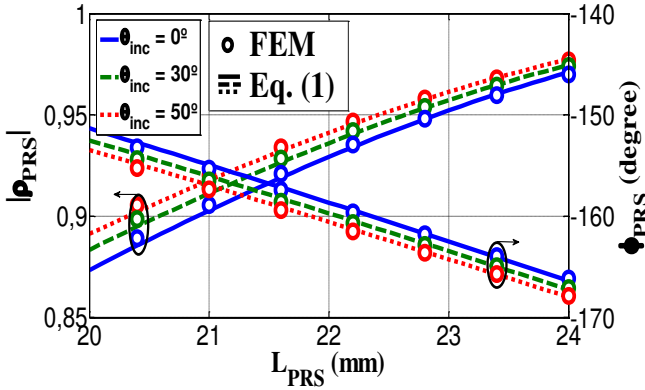


Fig. 3. Reflection coefficient of the PRS (magnitude and phase) seen by a plane wave for different incidence angles as a function of  $L_{PRS}$  at  $5.6\text{ GHz}$ .

modeling of the tunable HIS. In this case, the following pole-zero expression is employed for  $Y_{HIS}$ , which is now dependent on the varactors' junction capacitance  $C_j$ :

$$Y_{HIS}(k_y, C_j) = j \frac{[c_j - c_{j_{zn}}(k_y)] \dots [c_j - c_{j_{zn}}(k_y)]}{[c_j - c_{j_{p1}}(k_y)] \dots [c_j - c_{j_{p1}}(k_y)]} \quad (2)$$

In the antenna design, we use a one dimensional version of the tunable HIS configuration proposed in [29,32], so that a row of metallic patches of width  $Q_{HIS} = 14\text{ mm}$ , length  $L_{HIS} = 17\text{ mm}$ , separated at a distance  $P_{HIS} = 30\text{ mm}$ , are printed on Rogers RO 4230 substrate ( $\epsilon_{rHIS} = 3$ ,  $D_{HIS} = 1.524\text{ mm}$ ). Each HIS patch is divided in the  $x$ -direction by two equal patches separated by a gap ( $g = 1\text{ mm}$ ) where two varactor diodes are placed as illustrated in Fig.1-b.

As was demonstrated in [25], the boundary condition presented by the changes drastically as the physical resonating length  $L_{HIS}$  is varied, thus resulting in a direct mechanism to control the  $TE_{01}$  leaky-mode cut-off frequency and therefore the pointing angle at a fixed frequency. Yet, this control could only be mechanically achieved in [25], as the whole PCB should be replaced in order to vary  $L_{HIS}$ . The proposed reconfigurable LWA overcomes this disadvantage, since the effective resonant length of the HIS patches can be varied by properly modifying the varactors' junction capacitance  $C_j$  [29]. This is shown in Fig.4-a, where the reflection phase presented by the tunable HIS ( $\phi_{HIS}$  in Fig.2-b) provides PMC condition ( $\phi_{HIS} = 0^\circ$ ) at different frequencies ( $5.15\text{ GHz}$ ,  $5.65\text{ GHz}$  and  $6.15\text{ GHz}$ ) as  $C_j$  is varied ( $0.35\text{ pF}$ ,  $0.23\text{ pF}$ , and  $0.15\text{ pF}$ ). These results were obtained using the pole-zero analytical expression of  $Y_{HIS}$  as a function of frequency, for different values of  $\theta_{inc}$  and  $C_j$ . When the frequency is fixed to  $5.6\text{ GHz}$ , the behavior of  $Y_{HIS}(k_y, C_j)$  can be analytically modeled using (2), obtaining the HIS reflection phase as a function of  $C_j$  for different  $\theta_{inc}$  shown in Fig.4-b. Very good agreement with FEM results is observed for all the range of  $C_j$  (from  $0\text{ pF}$  to  $0.35\text{ pF}$ ) and for all incidence angles. The HIS behaves as a transparent sheet for  $C_j = 0\text{ pF}$  ( $\phi_{HIS} = 150^\circ$ ), changing to PMC sheet ( $\phi_{HIS} = 0^\circ$ ) for  $C_j \approx 0.23\text{ pF}$ , and to PEC sheet ( $\phi_{HIS} = 180^\circ$ ) for  $C_j = 0.35\text{ pF}$ . According to the results reported in [25], increasing  $C_j$  corresponds to a larger  $L_{HIS}$ , thus showing a direct relation between  $C_j$  and the effective resonant length of the HIS patches. Since  $C_j$  can be controlled

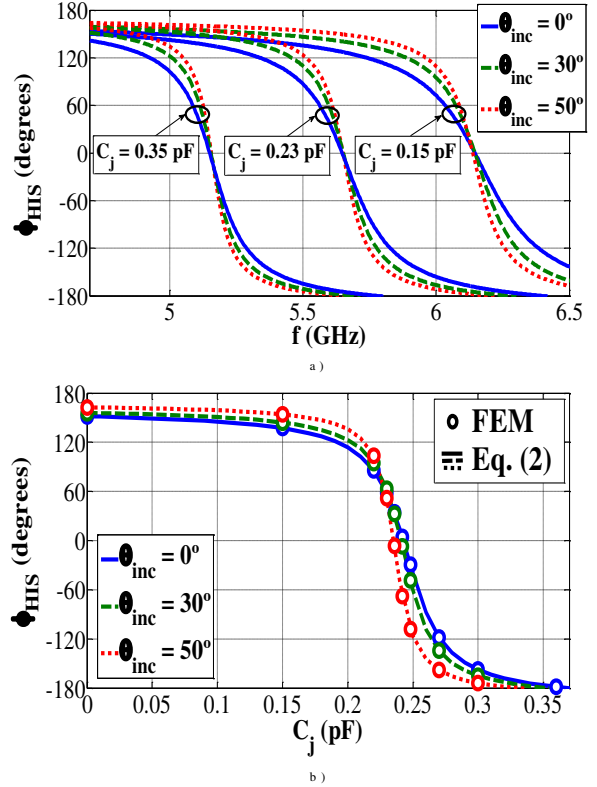


Fig. 4. Reflection phase of the tunable HIS seen by an incident plane wave for different incidence angles a) as a function of frequency for different values of  $C_j$ , b) as a function of  $C_j$  at  $5.6\text{ GHz}$ .

by the varactors' reverse polarization voltage [27]-[34], electronic control of the pointing angle at a fixed design frequency should be possible by combining the 1D FP leaky cavity and the tunable HIS. To theoretically confirm this fact, the behavior of the radiating leaky mode must be studied as a function of  $C_j$ .

#### A. Effect of $C_j$ in the leaky-mode dispersion curves.

Obtaining the leaky-mode dispersion curves as a function of  $C_j$  is essential for the electronic beam steering application presented. The unknown leaky-mode complex wavenumber  $k_y = \beta_y - j\alpha_y$  can be obtained from the TEN shown in Fig.2, by solving the corresponding Transverse Resonance Equation (TRE) [25],[37]:

$$Y_{UP}(f, k_y, L_{PRS}) + Y_{DOWN}(f, k_y, C_j) = 0 \quad (3)$$

Fig.5 shows the frequency dispersion for the normalized phase ( $\beta_y/k_0$ ) and leakage ( $\alpha_y/k_0$ ) constants as  $C_j$  is varied from  $0.1\text{ pF}$  to  $0.23\text{ pF}$ . As expected, the  $TE_{01}$  leaky-mode cutoff frequency decreases as  $C_j$  increases. This results in a continuous rise of the leaky-mode phase constant at a fixed frequency as  $C_j$  is increased, as it is shown in Fig.5-c for  $5.6\text{ GHz}$ . Again, very good agreement is obtained between TRM and full-wave FEM results, validating the proposed TEN. As it can be seen in Fig.5-c,  $\beta_y/k_0$  is varied from values close to zero when  $C_j = 0.1\text{ pF}$ , to values close to one when  $C_j = 0.23\text{ pF}$ . Since the leaky-mode pointing angle is given by  $\sin\theta_{RAD} \approx \beta_y/k_0$ , it is expected that fixed-frequency beam-scanning from broadside to endfire can be realised by controlling  $C_j$  in the aforementioned range [ $0.1\text{ pF}$ ,  $0.23\text{ pF}$ ].

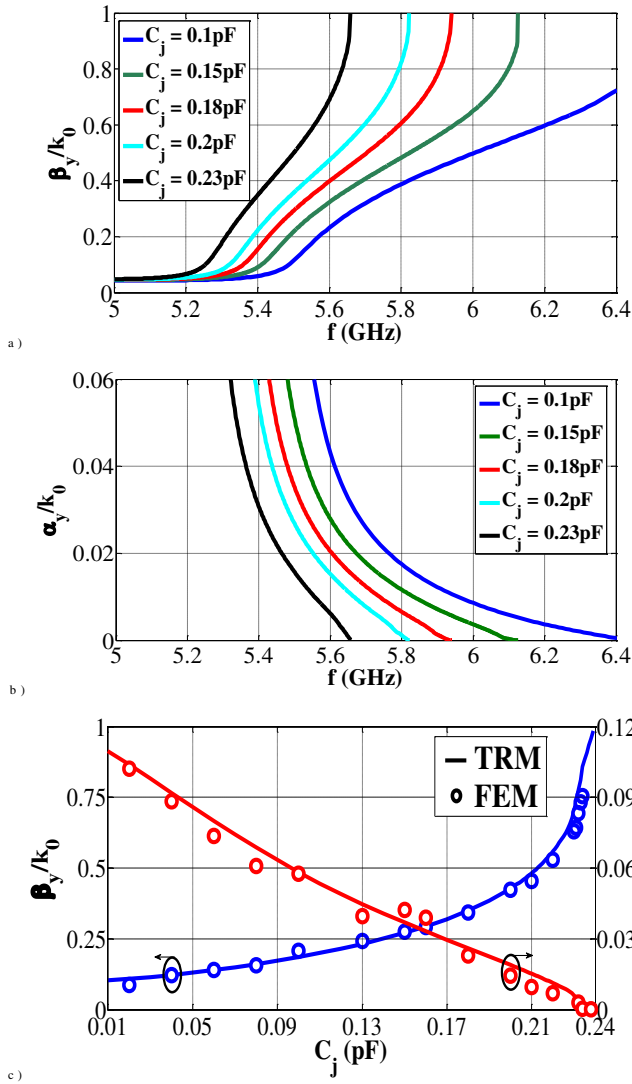


Fig. 5. Leaky-mode dispersion curves as a function of  $C_j$  ( $L_{PRS}=22\text{ mm}$ ) a-b) Dispersion with frequency c) Dispersion with  $C_j$  at  $5.6\text{ GHz}$ .

To give more physical insight into the working principle of the proposed reconfigurable LWA, Fig.6 illustrates the  $TE_{01}$  leaky-mode electric field distribution in the cross section of the FP cavity for different values of  $C_j$  at the design frequency of  $5.6\text{ GHz}$ . For low values of  $C_j$  the HIS behaves as a transparent sheet as shown in Fig.4-b and the leaky-mode is resonating in the metallic cavity of height  $H$ , providing maximum horizontal field at  $z=H/2$ , as it can be seen in Fig.6 for  $C_j=0.01\text{ pF}$ . However, as  $C_j$  is increased to  $0.23\text{ pF}$ , the HIS tends to behave as a PMC sheet which induces maximum field at its interface, as shown in Fig.6. The boundary conditions seen by the leaky-mode resonating in the FP cavity are strongly changed as  $C_j$  is varied. As a result, the leaky-mode transverse wavelength  $\lambda_z$  is modified from  $\lambda_z=2H$  when  $C_j=0.01\text{ pF}$  to  $\lambda_z=4H$  when  $C_j=0.23\text{ pF}$ , as illustrated in Fig.6. This enlargement in the transverse wavelength  $\lambda_z$  implies the subsequent reduction in the longitudinal wavelength  $\lambda_y$ , and therefore a rise in the leaky-mode longitudinal phase constant ( $\beta_y=2\pi/\lambda_y$ ) as  $C_j$  is increased, as predicted by the results shown in Fig.5-c.

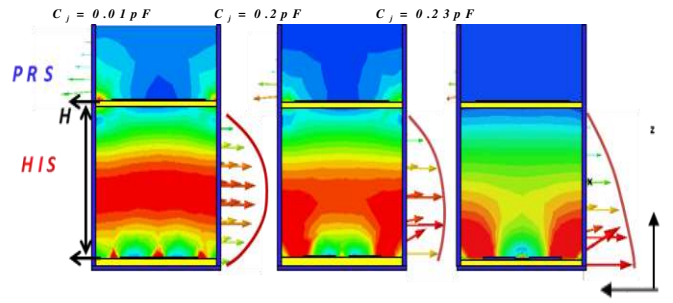


Fig. 6. Leaky-mode electric-field pattern inside the FP PRS-tunable HIS cavity (obtained from FEM) at  $5.6\text{ GHz}$  for different values of  $C_j$ .

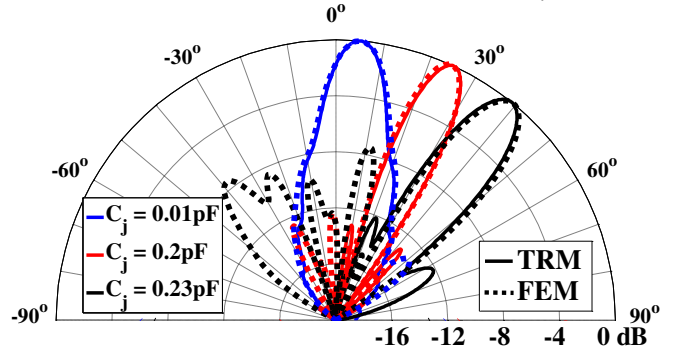


Fig. 7. Computed H-plane normalized radiation patterns for the proposed reconfigurable LWA at  $5.6\text{ GHz}$  for different  $C_j$  (antenna length  $L_A=5\lambda_0$ )

Once the leaky-mode complex propagation constant has been obtained as a function of  $C_j$ , the associated H-plane radiation pattern can be directly obtained [2], as shown in Fig.7 for the case of a LWA of length  $L_A=5\lambda_0$  at  $5.6\text{ GHz}$ . As expected, the scanning angle  $\theta_{RAD}$  is swept from nearly broadside towards the endfire direction as  $C_j$  is increased, following the  $\beta_y/k_0$  curve plotted in Fig.5-c. Particularly, Fig.7 shows the results for  $C_j=0.01\text{ pF}$  ( $\beta_y/k_0\approx 0.09$ ,  $\theta_{RAD}\approx 5^\circ$ ),  $C_j=0.2\text{ pF}$  ( $\beta_y/k_0\approx 0.42$ ,  $\theta_{RAD}\approx 25^\circ$ ), and  $C_j=0.23\text{ pF}$  ( $\beta_y/k_0\approx 0.65$ ,  $\theta_{RAD}\approx 40^\circ$ ). The beam direction and the beam width predicted from the TRM are in very good agreement with full-wave FEM simulations performed in a 3D model of the whole reconfigurable LWA. A strong reflected lobe appears for  $C_j=0.23\text{ pF}$ , due to the very low leakage-rate associated with high values of  $C_j$  (see Fig.5-c), which results in poor radiation efficiency and a large amount of energy being reflected at the far-end of the LWA. This fact limits the maximum scanning angle of the proposed reconfigurable LWA to  $40^\circ$ , as it will be explained in detail in Section III.

#### B. Optimization of the Fabry-Perot cavity height $H$ .

The physical height of the FP cavity ( $H$ ) is a key geometrical parameter which determines the scanning range of the proposed reconfigurable LWA. All previous results were computed using the optimum value  $H=H_{opt}=25.2\text{ mm}$ . As can be seen in Fig.8, a wider scanning range is obtained in this case, so that the cavity height makes the leaky-mode cut-off condition ( $\theta_{RAD}\approx 0^\circ$ ) to be coincident with the minimum value of  $C_j$ . In this situation, the dynamic range of the varactors is fully used and the minimum scanning angle is close to broadside. On the other hand, as discussed in [39], the maximum achievable pointing angle in this type of 1D LWA



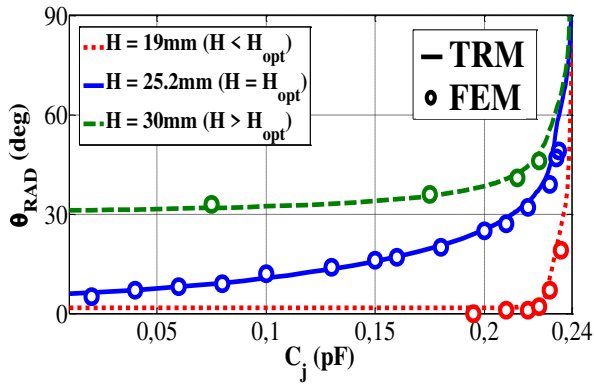


Fig. 8. Leaky-mode scanning-angle as a function of  $C_j$  for different values of the Fabry-Perot cavity height,  $H$ , at  $5.6\text{ GHz}$ .

is produced when the HIS presents a PMC resonance (which in our case corresponds to the value  $C_j = 0.23\text{ pF}$ ).

Above the HIS PMC resonance, a sudden fall of the leakage rate and a rapid divergence of the phase constant as shown in Fig.5-c limit the scanning range to higher angles. Fig.8 shows that for  $H = H_{opt}$  the scanning angle can be swept from  $\theta_{RAD} = 5^\circ$  for  $C_j = 0.01\text{ pF}$  to  $\theta_{RAD} = 40^\circ$  for  $C_j = 0.23\text{ pF}$ , as previously described. However, if  $H > H_{opt}$ , the minimum value of  $C_j = 0$  does not correspond to the leaky-mode cut-off point ( $\theta_{RAD} \approx 0^\circ$ ) but to a higher scanning state. As a result, the minimum scanning angle is located further from broadside ( $\theta_{RAD} = 30^\circ$  at  $C_j = 0.01\text{ pF}$  in Fig.8 for  $H = 30\text{ mm}$ ). Since the maximum pointing angle given by the HIS PMC resonance ( $C_j = 0.23\text{ pF}$ ) is  $\theta_{RAD} = 40^\circ$ , the scanning range is reduced to  $[30^\circ, 40^\circ]$  instead of  $[5^\circ, 40^\circ]$ . On the other hand, for shorter FP cavities ( $H < H_{opt}$ , shown in dotted red line in Fig.8) the leaky mode is below cutoff for most of the range of variation of  $C_j$ . As it can be seen in Fig.8 for  $H = 19\text{ mm}$ , the leaky mode is below cutoff from  $C_j = 0\text{ pF}$  to  $C_j = 0.2\text{ pF}$ , missing this region of the varactors' dynamic range. As a result, the scanning starts at  $C_j = 0.2\text{ pF}$  with  $\theta_{RAD} = 0^\circ$ , but then it is very soon limited to a maximum angle of  $\theta_{RAD} = 20^\circ$  at  $C_j = 0.23\text{ pF}$ . The scanning range has now been reduced to  $[0^\circ, 20^\circ]$  instead of  $[5^\circ, 40^\circ]$ . Figure 8 also shows that the optimum cavity provides the most linear dependence between  $\theta_{RAD}$  and  $C_j$ . Once more, excellent agreement is found in the validation of the TRM dispersion curves and FEM results. In the next sections, the beam scanning mechanism that has been theoretically predicted is demonstrated by experimental results.

### III. EXPERIMENTAL RESULTS

The following section describes the experimental results carried out on a prototype fabricated at *CSIRO ICT Centre*, shown in Fig.9. The antenna is designed to operate at the fixed frequency of  $5.6\text{ GHz}$ , and the radiating aperture has a length of  $L_A = 5\lambda_0$ . As illustrated in Fig.9-a, three constituent parts in the reconfigurable LWA can easily be distinguished: a parallel-plate waveguide, a passive PRS printed-circuit, and a tunable HIS. To excite the  $TE_{01}$  leaky-mode of the tunable 1D FP cavity, a horizontal coaxial probe is used as proposed in [25]. The dimensions of this feeding probe were optimized to obtain maximum input matching at the operating midpoint of

$C_j = 0.15\text{ pF}$  in order to obtain good matching across most of the dynamic range of  $C_j$ .

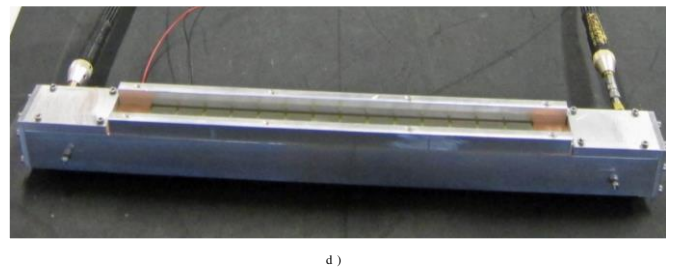
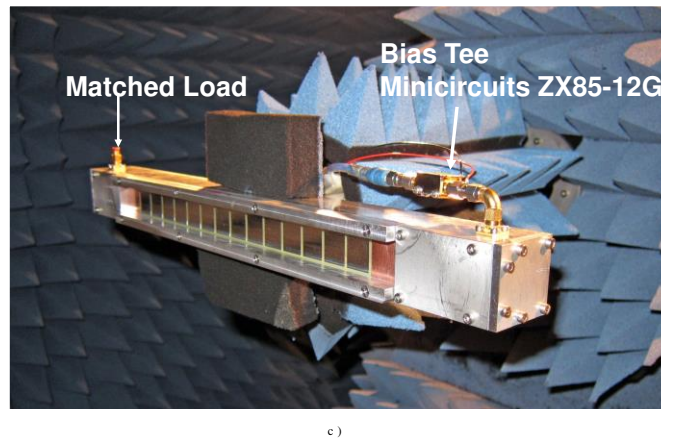
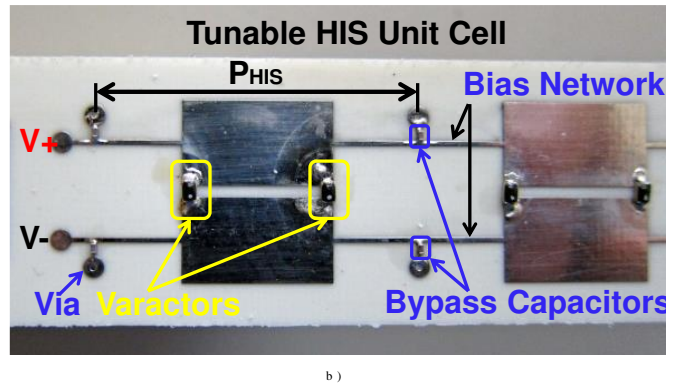
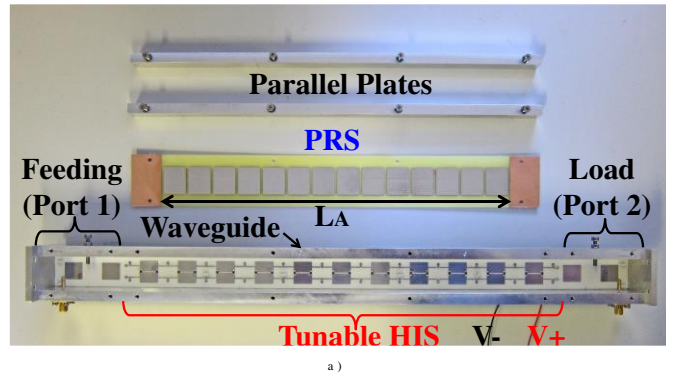


Fig. 9. Photographs of a) Manufactured reconfigurable LWA prototype b) tunable HIS phase agile cell c) Radiation pattern measurement experimental set-up and d) S parameters measurements set-up.

A detailed photograph of the phase agile cell used for the tunable HIS, including the biasing network proposed in [32], is depicted in Fig.9-b. Metelics' MGV125-08 varactor diodes (0805 package) are used for the tunable HIS, providing a range in  $C_j$  from  $0.055\text{ pF}$  to  $0.6\text{ pF}$  as reverse bias voltage  $V_{ris}$

tuned from  $20V_{DC}$  to  $2V_{DC}$ . The varactor's tolerance for  $C_j$  is  $\pm 0.05pF$  according to the manufacturer datasheet.

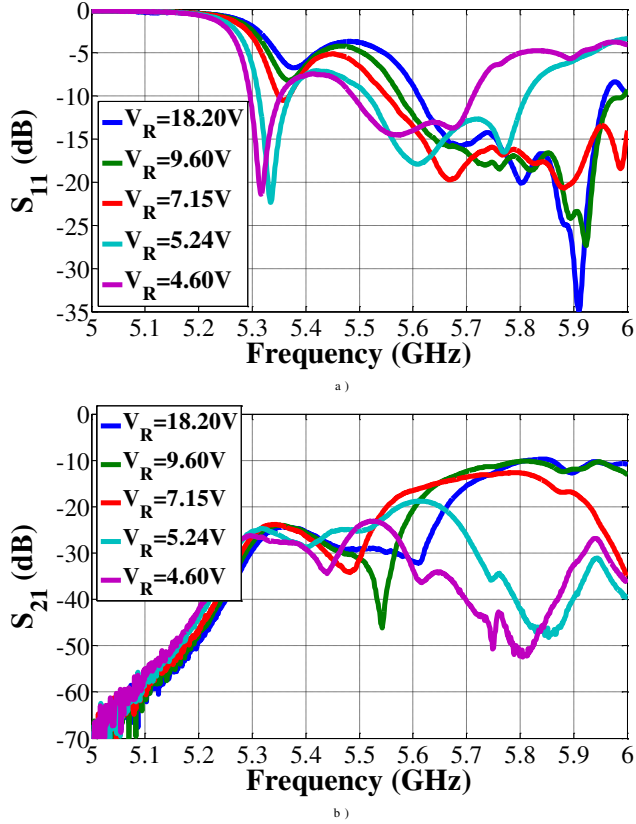


Fig. 10. Measured S parameters vs frequency for different  $V_R$ . a)  $S_{11}$ , b)  $S_{21}$ .

The measured S parameters in the  $5GHz-6GHz$  frequency range are presented in Fig.10 for different values of bias voltage  $V_R$ . As expected, the matching band is shifted to lower frequencies as  $V_R$  is decreased (and  $C_j$  is increased), in accordance with the expected theoretical shift in the leaky-mode frequency dispersion presented in Fig.5. These results are also consistent with data reported in previous reconfigurable LWAs using varactors[27-34]. At the fixed operation frequency of  $5.6GHz$ , Fig.11 shows the simulated and measured S parameters as a function of  $V_R$ . As expected, the matching ( $S_{11}$ ) and the transmission ( $S_{21}$ ) coefficients are affected as  $V_R$  ( $C_j$ ) is varied. Simulated data predict optimum operation with  $S_{11} = -23.7dB$  and  $S_{21} = -9.54dB$  for  $V_R = 7.17V_{DC}$  (which corresponds to  $C_j = 0.15pF$ , according to the datasheet).

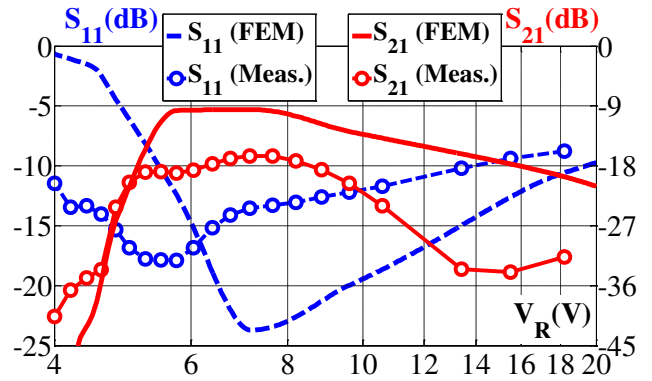


Fig. 11. Measured S parameters vs  $V_R$  at  $5.6GHz$ .

As  $V_R$  is varied from this value of  $7.17V_{DC}$ , the  $TE_{01}$  leaky-mode field distribution is strongly perturbed (as it was theoretically illustrated in Fig.6), thus decreasing the matching between the coaxial probe and the FP cavity. It is important to note that, as previously mentioned, the coaxial probe dimensions were optimized for the aforementioned operating point  $C_j = 0.15pF$  ( $V_R = 7.17V_{DC}$ ). As a result, poorer matching is obtained at other operation points of the varactor's dynamic range. In particular, the mismatch increases up to  $S_{11} = -2dB$  for  $V_R = 4.5V_{DC}$  ( $C_j = 0.24pF$ ) and to  $S_{11} = -10.6dB$  for  $V_R = 18.2V_{DC}$  ( $C_j = 0.06pF$ ). Experimental data report a similar tendency for the measured S parameters as a function of  $V_R$ , obtaining  $S_{11} = -17.9dB$  and  $S_{21} = -19dB$  at the optimum tuning point  $V_R = 5.75V$  ( $C_j = 0.19$ ) and similar deterioration for other values of  $V_R$ . As can be seen, measured data is shifted to lower  $V_R$  values with respect to simulations. This shift is attributed to the varactors' tolerance errors, as demonstrated next.



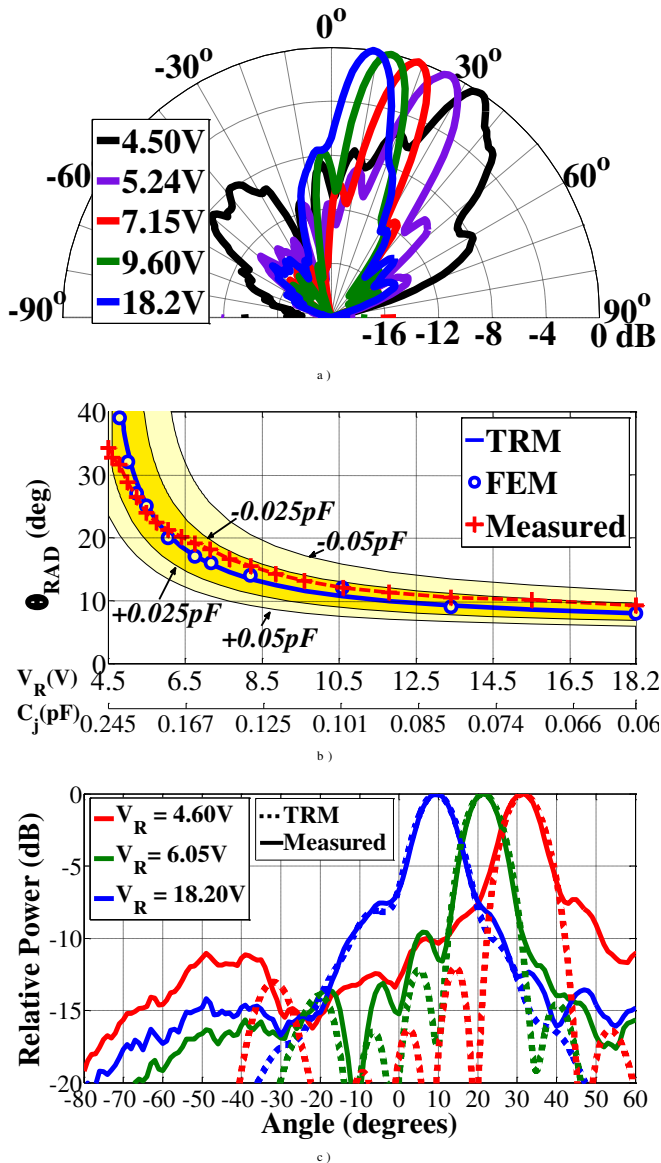


Fig. 12. Radiation response of reconfigurable LWA at 5.6 GHz vs  $V_R$ . a) Measured radiation patterns b) Pointing angle scanning c) Comparison between measured and theoretical radiation patterns.

The electronic control over the LWA radiation pattern is shown in Fig.12. Fig.12-a depicts the measured normalized radiation pattern at 5.6 GHz for different values of the applied bias voltage  $V_R$ . As it can be seen, the main beam is scanned from  $\theta_{RAD} = 9.2^\circ$  for  $V_R = 18.2 V_{DC}$  ( $C_j = 0.06 pF$ ) to  $\theta_{RAD} = 34.2^\circ$  for  $V_R = 4.5 V_{DC}$  ( $C_j = 0.245 pF$ ). Figs.12-b and 12-c show the agreement between theory and experiments. In particular, Fig.12-b plots with red crosses the measured scanning response ( $\theta_{RAD}$  vs  $V_R$ ) of the proposed reconfigurable LWA, confirming the continuous electronic scanning from  $34.2^\circ$  to  $9.2^\circ$  as  $V_R$  is increased from  $4.5 V_{DC}$  to  $18.2 V_{DC}$ . The theoretical leaky-mode scanning response extracted from the TRM is plotted with a continuous blue line, while FEM simulations are represented with blue circles. Very good agreement is observed between these three set of results. However, the measured scanning response is slightly shifted with respect to TRM and FEM results, showing discrepancies in  $\theta_{RAD}$  that are below  $\pm 2^\circ$ . This can be attributed to the aforementioned

$\pm 0.05 pF$  varactors' tolerances in  $C_j$ . To demonstrate this fact, error curves computed by applying a shift in  $C_j$  of  $\pm 0.025 pF$  and  $\pm 0.05 pF$  are represented with shaded zones in Fig.12-b. As can be seen, the measured scanning response lies inside the  $\pm 0.025 pF$  region, well below the diodes' tolerance ( $\pm 0.05 pF$ ).

Finally, Fig.12-c compares the theoretical leaky-mode and measured radiation patterns for three illustrative values of  $V_R$  covering the entire scanning range. Not only the main beam direction is accurately coincident between theory and experiments for all  $V_R$ , but also the 3dB beam width and the secondary lobes level and position are well predicted, thus confirming the leaky-mode radiation mechanism for the reconfigurable antenna. To obtain the results of Fig.12-c, the measured pointing angle  $\theta_{RAD}$  and the requested experimental tuning voltage  $V_R$  and associated  $C_j$  (according to the nominal  $V_R$ - $C_j$  response of the varactors' datasheet) have been used as references for each case. In order to show the discrepancies between the experimental and the theoretical electronic scanning, Table I summarizes the pointing angle  $\theta_{RAD}$  obtained for three operation points inside the scanning region. As it is shown, maximum error is obtained at lower voltages that correspond to larger scan angles ( $V_R = 5 V_{DC}$ ,  $\Sigma \theta_{RAD} = -5.3^\circ$ ).

TABLE I  
BIAS VOLTAGE AND POINTING ANGLE FOR THREE OPERATING POINTS OF THE RECONFIGURABLE LWA

Case	Theory (TRM)	Experiments	Error
$V_R (V_{DC})$	$\theta_{RAD}$	$\theta_{RAD}$	$\Sigma \theta_{RAD}$
5.0	$34.1^\circ$	$28.8^\circ$	$-5.3^\circ$
6.0	$20.5^\circ$	$21.2^\circ$	$+0.7^\circ$
20.0	$8^\circ$	$9.2^\circ$	$+1.2^\circ$

Finally, Fig.13 summarizes the experimental performance of the reconfigurable LWA in terms of gain and efficiency. Fig.13-a shows the variation of the gain as a function of  $V_R$ . Maximum measured gain of  $G = 12.95 dBi$  has been obtained for  $V_R = 10.6 V_{DC}$  ( $C_j = 0.1 pF$ ,  $\theta_{RAD} = 12^\circ$ ), while simulations predict maximum  $G = 12.6 dB$  for  $V_R = 9.6 V_{DC}$  ( $C_j = 0.11 pF$ ,  $\theta_{RAD} = 11^\circ$ ).

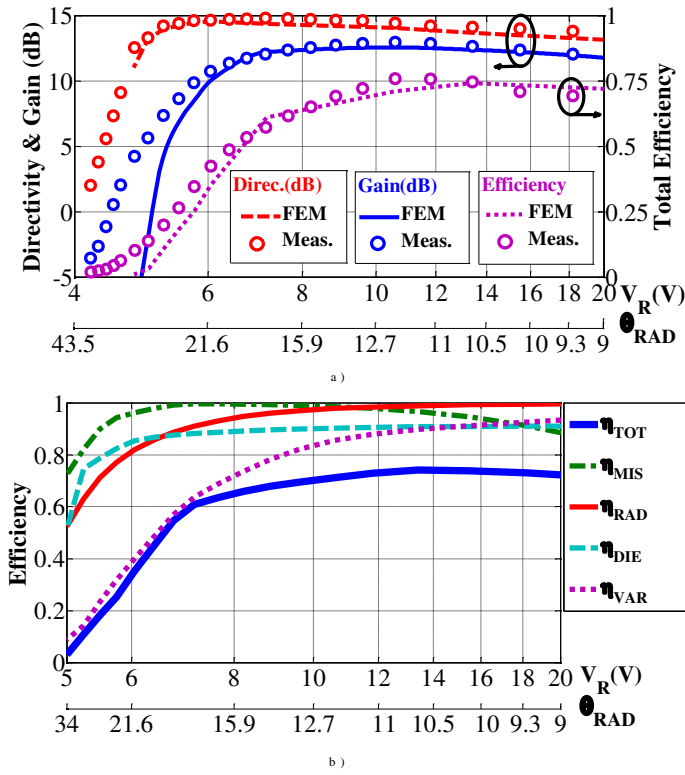


Fig.13. Results at 5.6 GHz vs  $V_R$  a) Measured gain, directivity and total efficiency and b) Estimated efficiency and subefficiencies from Full-Wave simulations.

Moreover, as  $V_R$  is decreased from this optimum point, a significant drop in the gain is observed. The minimum value of gain measured is  $G = -3.55 \text{ dB}$  for  $V_R = 4.2 \text{ VDC}$  ( $C_j = 0.26 \text{ pF}$ ), which corresponds to an angle close to  $\theta_{RAD} = 35^\circ$ . Similar behavior is obtained with simulations, thus making it difficult to obtain large scan angles. As it was previously explained, this fact is caused by higher mismatch (see Fig.11) and by the fall of the leakage rate (see Fig.5-c) that take place at high values of  $C_j$  (lower  $V_R$ ), due to the associated PMC resonance of the HIS for high  $\theta_{RAD}$  (see Fig.6). However, the LWA shows a stable gain above  $G = 10 \text{ dB}$  for  $V_R > 6 \text{ VDC}$ , which corresponds to scanning angles below  $\theta_{RAD} = 25^\circ$ . As also plotted in Fig.13-a, this gain translates into a total efficiency  $\eta_{TOT} = G/D > 50\%$ , reaching  $\eta_{TOT} = 75\%$  for  $V_R = 10-14 \text{ VDC}$  ( $\theta_{RAD} = 13^\circ-10^\circ$ ). Measured and simulated total efficiency ( $\eta_{TOT}$ ) in Fig.13-a are in concordance, observing  $\eta_{TOT} > 60\%$  in the range  $V_R = 8-20 \text{ VDC}$  ( $\theta_{RAD} = 16^\circ-10^\circ$ ), whereas  $\eta_{TOT}$  tends to 3% for  $V_R = 5 \text{ VDC}$  ( $\theta_{RAD} > 30^\circ$ ). To understand the gain and efficiency response of the LWA, Fig.13-b separates  $\eta_{TOT}$  into its different constitutive efficiencies [41,42]:

$$\eta_{TOT} = \eta_{MIS} \cdot \eta_{RAD} \cdot \eta_{DIE} \cdot \eta_{VAR} \quad (4)$$

where the mismatch efficiency is computed from the measured input reflections ( $\eta_{MIS} = 1 - |S_{11}|^2$ ) [41], the leaky-mode radiation efficiency is estimated from the theoretical leakage rate ( $\eta_{RAD} = 1 - e^{-2\alpha y L_A}$ ) [42], and the ohmic efficiency has been attributed to two contributions: one part due to the dielectric losses and another term due to dissipation in the varactors series resistance:  $\eta_{\Omega} = \eta_{DIE} \cdot \eta_{VAR}$  [41]. Since the separated ohmic efficiencies (due to dielectrics and varactors)

are difficult to compute from experiments [41], they have been estimated from full-wave simulation data. Equation (4) gives, as a first order approximation, a good insight into the origin of the gain drop for large scan angles. Particularly, it can be seen in Fig.13-b that the mismatch efficiency  $\eta_{MIS}$  is quite high (over 85%) for the entire dynamic range. On the other hand, the leaky-mode radiation efficiency  $\eta_{RAD}$  strongly decreases from 100% at  $V_R = 20 \text{ VDC}$  ( $\theta_{RAD} = 9^\circ$ ) to 52% at  $V_R = 5 \text{ VDC}$  ( $\theta_{RAD} = 30^\circ$ ), being consistent with the drop of the leakage rate predicted in Fig.5-c for high values of  $C_j$  (i.e. high values of  $\theta_{RAD}$ ). The dielectric efficiency results predict a fall from  $\eta_{DIE} = 90\%$  at  $\theta_{RAD} = 9^\circ$  ( $V_R = 20 \text{ VDC}$ ) to  $\eta_{DIE} = 50\%$  at  $\theta_{RAD} = 30^\circ$  ( $V_R = 5 \text{ VDC}$ ), which is caused by the electric field concentration at the HIS substrate for the PMC regime (see Fig.6 for  $C_j = 0.23 \text{ pF}$ ). Due to the same high concentration of the modal fields, an increase in the current density flowing across the diodes occurs at this PMC operational point, raising the heat dissipation at their series resistance. As a result, the varactors' ohmic efficiency  $\eta_{VAR}$  strongly drops for scanning angles tending to endfire, observing a decrease from  $\eta_{VAR} = 94\%$  at  $\theta_{RAD} = 9^\circ$  ( $V_R = 18 \text{ VDC}$ ) to  $\eta_{VAR} = 9\%$  at  $\theta_{RAD} = 30^\circ$  ( $V_R = 5 \text{ VDC}$ ). This result agrees with other published works that report significant increase in the HIS ohmic losses when operating close to the PMC resonance, and which has even been widely applied to create planar microwave absorbers [44-46]. Therefore, it can be concluded that the gain drop for large scan angles is unavoidable and it is due to two main reasons: the decrease in the leaky-mode radiation efficiency, and the increase in the varactors resistance losses.

#### IV. CONCLUSIONS

A new type of electronically scannable one-dimensional LWA, which is based on a tunable Fabry-Perot (FP) cavity formed by a Parallel-Plate Waveguide (PPW), a Partially Reflective Surface (PRS), and a varactor-loaded active High Impedance Surface (HIS), has been presented in the paper. The proposed reconfigurable LWA topology allows to electronically modify the resonant condition of the  $TE_{01}$  leaky mode inside the FP cavity, resulting in a variation of the pointing direction at a fixed frequency as the varactor bias is changed. Leaky-mode dispersion curves have been obtained from a simple but accurate Transverse Equivalent Network, which rigorously takes into account the effect of the varactors' junction capacitance in the phase response of the active HIS. These modal curves are very useful for the design of the main parameters of the antenna, in order to optimize the beam-scanning range. Leaky-mode results are in very good agreement with full-wave simulations and experiments, which have shown a continuous scanning from  $9^\circ$  to  $30^\circ$  in a prototype operating at the fixed frequency of 5.6 GHz, as the varactors' bias voltage is varied from 20 VDC to 5 VDC. The scanning range is limited in the endfire region due to the drop in the leakage rate and the increase of thermal losses, which are associated with the Perfect Magnetic Conductor (PMC) condition of the tunable High Impedance Surface. An interesting advantage of this new reconfigurable topology resides in its potential application to two-dimensional FP LWAs [32],[33],[35],[40] and to new metasurface LWAs [47],[48], so that the electronic scanning could be extended for

both elevation and azimuthal planes. This type of electronically steerable LWA presents a low-cost potential alternative to more expensive phased-arrays [48].

#### ACKNOWLEDGMENT

The authors would like to thank Mr Carl Holmesby for fabricating the waveguide, Mr Rob Shaw for soldering the surface mount components, and Mrs. María García-Vigueras for her insightful and interesting comments.

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