Beam Scanning Antenna Using a Reflectarray as Sub-Reflector

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Abstract **—** In this paper, a dual-reflector antenna based on a main parabolic reflector and a reconfigurable reflectarray as subreflector is proposed for beam scanning applications. The beam deflecting is achieved by modifying the phase introduced by each element of the subreflectarray. The required phase distribution for each scan angle is obtained through a synthesis technique based on the analysis of the antenna in receive mode. The design technique has been applied to the particular cases of beam scanning in azimuth and elevation planes. Patches aperture-coupled to delay lines, which provide low losses and cross-polar levels, are proposed as elements for the reflectarray subreflector, allowing the easy implementation of electronic control devices in the microstrip delay lines. The results show that the beam can be scanned in a range ±6º by inserting switches on the delay line to provide a 3-bit quantization.

Index Terms **—** Beam scanning, dual-reflector antenna, reconfigurable antenna, reflectarray.

I. INTRODUCTION

Antennas with beam scanning capabilities are required in a number of applications, as radar, emergency communications, or earth observation from space missions, [1-2]. Usually, these applications also require high gain antennas (narrow beams). Different solutions have been proposed, based either on mechanic, electronic, or hybrid mechanic-electronic scanning of the beam [3-23]. Lenses [3], reflector antennas [4-6], phased arrays [7-12], reflectarrays [13-20], and array fed reflectors [21-23] have been studied and proposed as solutions for beam steering applications requiring a narrow beam. The solutions based on lenses usually imply the use of mechanic devices [3]. The motion of the antenna components can be reduced by the use of complex antenna optics based on three reflectors [4]. Mechanic scanning options are not optima in terms of volume, mass and power consumption, being preferred electronic beam scanning. Beam steering can also be achieved by feeding the reflector by a horn array and using two shaped reflectors [5-6]. However, the cost of a shaped reflector manufacture process is high because of the moulds. Two possibilities for achieving electronic beam scanning are phase-arrays and reflectarrays. Phased arrays have been used as reconfigurable and wide angle beam steering antennas [7-8]. The reconfiguration of the beam is achieved by introducing active elements [9]. Combined electronic and mechanic beam control is proposed for applications with stringent scanning requirements [10]. Conventional phased arrays with large apertures would require transmit-receive (T/R) modules due to the losses of the feeding network [7], requiring very complex control circuitry [11-12]. On contrary, reflectarray is spatially fed, without any feeding network, and the ohmic losses are much lower than in phased arrays. Then, reflectarrays can provide a low-cost low-complexity solution for beam scanning by simply inserting low-loss phase-shifters in each element [13-16]. Some recent works have shown different implementations for phase control in reflectarray antennas, based on electronic devices as MEMS or diodes [17-20].

If high gain and electronic reconfigurability or scanning of the beam is required, a dual-reflector antenna with an electronically controllable reflectarray subreflector can be used (see

Fig. 1). Reflectors with an array as feed have been proposed for beam steering applications [21-23]. Different optics have been proposed for this configuration, including single parabolic [21] or spherical [22] reflectors and Cassegrain dual reflectors configurations [23]. These options reduce the complexity, mass, volume, and power consumption of the control circuitry compared to a large reconfigurable phased array. However, the array as feed solution exhibits the same inconveniences as single phased arrays: transmit-receive (T/R) modules and complex feeding networks are required. The capability of a subreflectarray–main parabolic reflector configuration for beam scanning has been demonstrated for a passive reflectarray [24], where the beam was deflected by introducing a progressive phase along the y-axis of the subreflectarray, according to Fig. 1. However, the results in [24] showed that the beam suffered also a slight deflection in the vertical direction, because the reflectarray subreflector was tilted. To avoid this problem, a more elaborate phase-synthesis is required to provide beam scanning, than simply introducing a progressive phase on the sub-reflectarray.

A phase-synthesis technique is proposed in this paper to provide beam scanning in a dual-reflector configuration using a reflectarray subreflector. The synthesis technique has been applied to design a dual-reflector antenna for beam scanning in the principal planes at 11.7GHz. After defining the geometry and elements of the antenna, the beam scanning performance is evaluated.

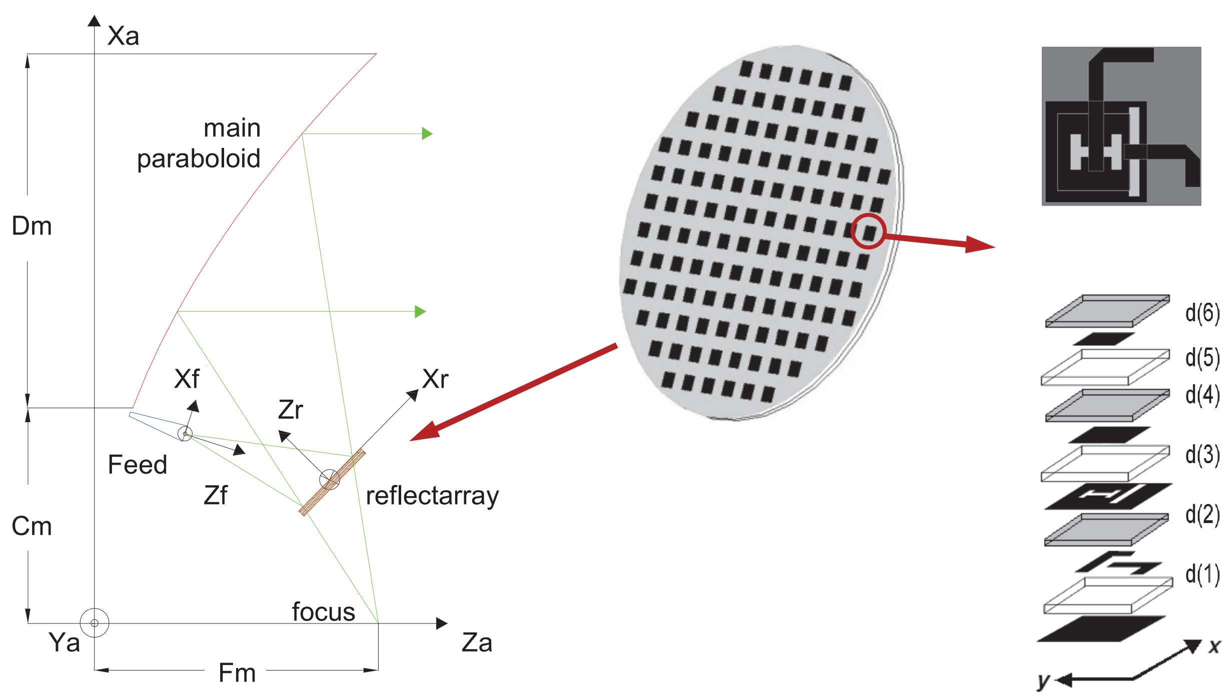


Fig. 1. Scheme of the dual-reflector configuration.

II. ANALYSIS AND SYNTHESIS TECHNIQUES

A scheme of the configuration of the antenna is shown in

Fig. 1, which includes three main elements: a primary feed (horn antenna), a reflectarray as subreflector, and a main parabolic reflector.

A. Antenna analysis

The antenna analysis technique is described in [25]. It combines several techniques for the analysis of the different elements of the antenna. For simplicity, an ideal feed-horn model based on $\cos^q(\theta)$ function is used. However, the near-field of the primary feed, obtained from measurements or full-wave simulations, can be used [26]. The elements of the subreflectarray are analyzed through a convenient full-wave tool, considering local periodicity and the real incidence angle of the wave coming from the feed. Once the field reflected by the elements of the subreflectarray has been obtained, the parabolic reflector is analyzed through physical optics (PO).

The far field produced by each reflectarray element, considered as a small rectangular aperture, is used to compute the PO printed currents on the parabolic surface. The equivalent currents on the main reflector are computed by adding the contributions from all the elements of the reflectarray, and the final illumination (amplitude and phase of the incident field) on the parabolic surface depend on the phase-shift introduced by the reflectarray cells. Finally, the radiation pattern is computed using an FFT-based algorithm, which is applied to the electric field on the antenna aperture. This technique has been validated by comparing the simulated and measured radiation patterns of a breadboard in the 94GHz band, see [24].

In this antenna, the beam is deflected by adjusting only the phase-shift on the reflectarray elements. Note that the amplitude is not changed on the reflectarray subreflector, except for small variation in the ohmic losses. However, the resulting illumination on the main reflector is different for each scan angle, because the currents induced by all the reflectarray cells are superposed.

B. Phase synthesis for beam scanning

A technique based on PO is applied to synthesize the required phase distribution on the reflectarray subreflector for a required scan angle. The synthesis is carried out by analyzing the antenna in receive mode, as in [27]. In a receiving antenna, the incident field is a plane wave that propagates in a direction forming an angle with the paraboloid Z_a-axis, see Fig. 1. In a dual reflector antenna, the incident wave defines an electric field distribution on the antenna aperture (the flat surface defined by the reflector edge) with a progressive phase according to the incidence angle. Then, the inverse process to that used for the analysis of the transmitting antenna is carried out. First, the PO currents on the reflector surface are calculated. Second, the incident electric field on each element of the subreflectarray produced by the printed currents in the main reflector is computed, being the phase of this field retrieved on each reflectarray element (m,n) denoted as $\phi_{retr}(m,n)$. The phase of the field reflected by the reflectarray in receive mode will be:

$$\phi_{RA}(m,n)+\phi_{retr}(m,n), \qquad (1)$$

where $\phi_{RA}(m,n)$ is the phase-shift introduced by the reflectarray element (m,n). The fields reflected on all the elements (m,n) should converge to the focal point where the feed is located, after propagating along the corresponding paths. This propagation produces the phase delay $\phi_{feed}(m,n)$ from the element (m,n) to the feed. By reciprocity, $\phi_{feed}(m,n)$ is the same phase of the incident field on the reflectarray element (m,n) coming from the feed when the antenna is in transmit mode. Therefore, assuming the antenna in reception mode, the condition to concentrate all the signals to the focal point when a field is received from a given scan angle, is:

$$\phi_{RA}(m,n)+\phi_{retr}(m,n)+\phi_{feed}(m,n)=C, \qquad (2)$$

where C is a constant. Then, the required phase-shift on the reflectarray elements $\phi_{RA}(m,n)$ is:

$$\phi_{RA}(m,n)=-\left(\phi_{feed}(m,n)+\phi_{retr}(m,n)\right)+C. \qquad (3)$$

Finally, the phase-shift $\phi_{RA}(m,n)$ has to be implemented in each reflectarray element. If a discrete phase control is used, the phase shift at each reflectarray cell will be approximated by a finite number of phase values, defined by

$$\phi_{RA}'(m,n) = (p-1)\frac{\pi}{2^{k-1}}, \qquad p \in \mathbb{N}\left[1, 2^k\right], \tag{4}$$

where k is the number of control bits in each element and a uniform quantization is assumed. Thus, the value of the phase $\phi_{RA}'(m,n)$ is chosen at each cell (m,n) to minimize the phase error.

III. RESULTS

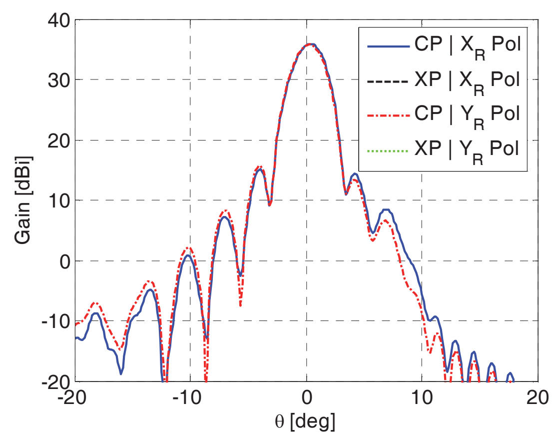
A. Antenna definition

A general scheme of the dual-reflector antenna is shown in

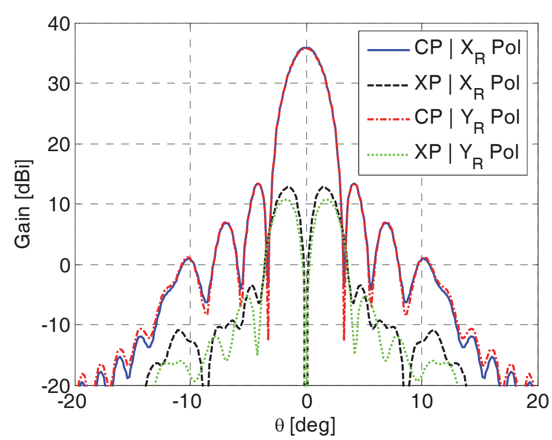
Fig. 1. The antenna geometry is defined by the parameters given in Table 1. The design and analysis is carried out at 11.7GHz. In this antenna, the main parabolic reflector is oversized in order to obtain beam deflecting without significant reduction in the gain because of spillover. Note that the change in beam pointing is obtained by scanning the illumination on the main reflector surface. The antenna optics has been defined to fulfill two conditions. First, a beam in the boresight direction should be obtained when the phase-shift is constant along the subreflectarray surface. Second, the beam is scanned in a ±6degree range both in elevation and in azimuth without subreflector blockage. As shown in Table 1, the reflectarray is elliptical with 24 and 22 elements along X_R and Y_R axis, respectively. Considering that the size of the reflectarray cell is 12.5mm × 12.5mm, the axes of the resulting surface are 300mm and 275mm. In this case, the reflectarray is illuminated with a taper at edges of -10dB. The radiation patterns of the boresight beam have been computed on the principal planes considering an ideal phase distribution for dual linear polarization, see Fig. 2. The computed antenna gain is 35.8 dBi. According to the antenna optics, the spillover efficiency is quite high, ε_s=0.81. However, the oversized main reflector produces a low aperture efficiency, ε_t =0.31, being the antenna efficiency the product of the two terms, $\varepsilon=\varepsilon_s*\varepsilon_t$=0.25.

Table 1: Main parameters of the antenna

Parabolic Reflector	
Aperture diameter (Dm)	989mm
Clearance (Cm)	595mm
Focal Distance (Fm)	792mm
Subreflectarray	
Center	(390, 0, 640)mm
Periodic cell size	12.5mm×12.5mm
Reflectarray size	24 × 22 elements
Direction cosines	$\begin{pmatrix} \sqrt{2}/2 & 0 & \sqrt{2}/2 \\ 0 & -1 & 0 \\ \sqrt{2}/2 & 0 & -\sqrt{2}/2 \end{pmatrix}$
Feed-horn *(in Sub-RA coordinate system)*	
Phase center	(-174, 0, 375)mm
Pointing	(0, 0, 0)mm



(a)



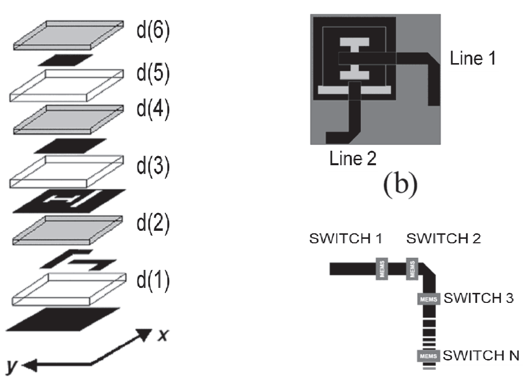
(b)

Fig. 2. Radiation pattern of the boresight beam. Main cuts in elevation (a) and azimuth(b).

B. Sub-reflectarray periodic cell definition

The proposed reflectarray element is based on the patches aperture-coupled to delay lines, as that shown in Fig. 3 and can be used for dual linear polarization. With the aim of providing a broadband element, two square patches have been stacked (dimensions of the upper patch are scaled by a factor of 0.68 with respect those of the lower patch which is 8.0mm × 8.0mm). The room restriction for placing the two slots has been solved using an H-shaped slot for the X-polarization (the width of all the branches is 0.8 mm, the main branch length is 4mm and the secondary branches length is 2.8mm), while a rectangular slot is used for the Y-polarization (0.8mm × 7mm). The width of both delay lines is 1.18 mm. The period for the element has been fixed to 12.50mm×12.50mm, while the dielectric materials used in the design are summarised in Table 2.

These elements exhibits very low cross polarization levels and allow the implementation of electronic phase control devices, as MEMS [28] or diodes [29]. Because of the presence of the ground plane, the electronic switches and their control network are on the opposite side of the radiating patches, avoiding spurious radiation. Fig. 3(c) shows one option for implementing series switches between different segments of the microstrip line, allowing a change in the length of the delay line and therefore a variation in the phase of the reflected wave, [30].



(a) (c)

Fig. 3. Dual-polarization reflectarray element based on patches aperture-coupled to delay lines. (a) Expanded view. (b) Upper view. (c) Implementation of series switches in the delay line.

Fig. 4(a) and (b) show respectively the phase and amplitude response of the element when a plane wave with the electric field oriented in the X-axis direction impinges on top of the element. The surfaces have been generated as a function of both line lengths using the frequency domain solver of CST Microwave Studio® [31] with Floquet conditions. The element has been analyzed for an incident field impinging at an angle $\theta=\varphi=30^o$, assuming that for smaller angles the cross-polar component will be lower. As can be seen, in all the cases the average losses produced at 11.70 GHz are around 0.2 dB.

For X-polarization, the phase is practically a linear function of the delay line L1 (for different values of the second delay line, L2), as shown in Fig. 4(c). These phases are compared with that produced by an ideal phase-shifter where the phase is equal to $-2\beta L$, being β the propagation constant for the microstrip line, at the working frequency of 11.70GHz, and L the length of the delay line. More than 360º of phase delay can be obtained if the line is bent, allowing true-time delay (TTD) [32]. The average cross-polar levels are better than –25 dB, except for the length L1=5.8 mm, where increases until –18 dB, as can be seen in Fig. 4(d). The phase response of this element remains very similar for incidence angles up to $\theta=30^o$, with variation in phase smaller than 30º. As a first approach, the subreflectarray can be designed considering the data from normal incidence, however the real angles of incidence should be taken into account for a more accurate design.

The electronic control implementation in these kinds of elements has been previously validated with PIN diodes [33] and MEMS devices [28].

Table 2: Dielectric slabs values

Layer	**d(1)**	**d(2)=d(4)=d(6)**	**d(3)=d(5)**
Material	Vacuum	Arlon	Eccostock
thickness (mm)	6.400	0.508	1.000
ε_r	1.0054	3.3800	1.0400

tanδ	0.0002	0.0035	0.0010

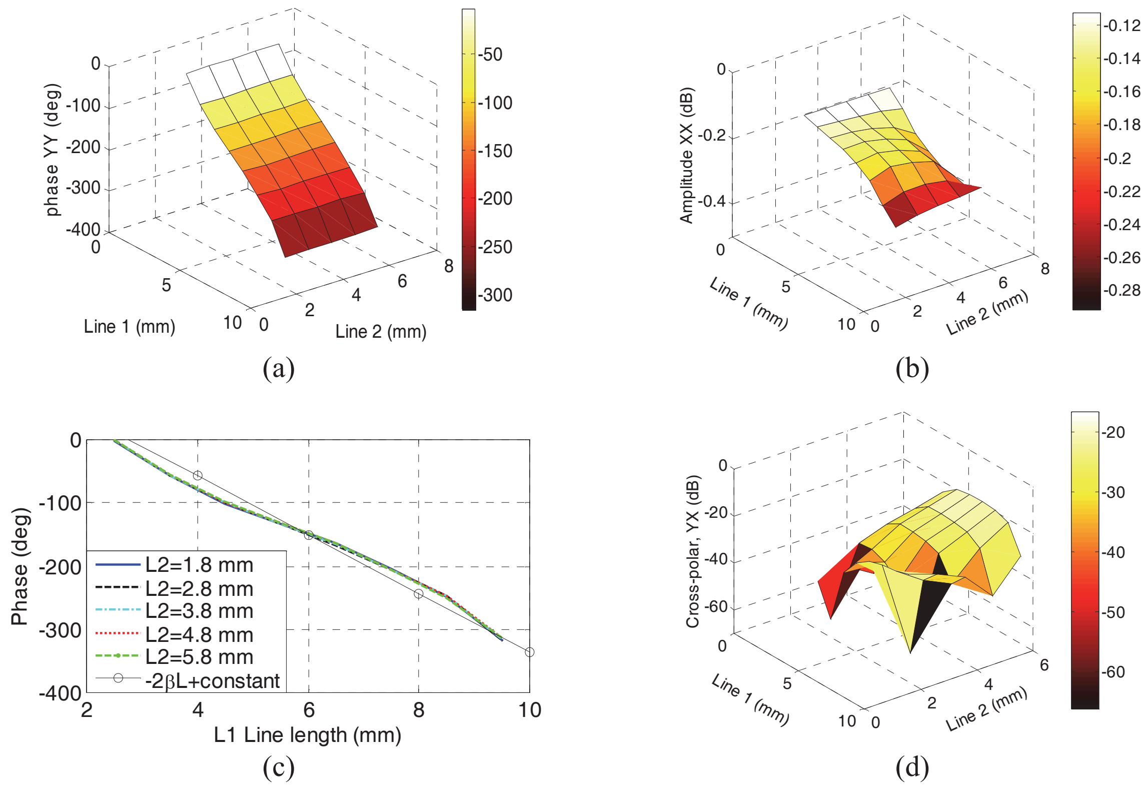


Fig. 4. Response for the proposed reflectarray element, at 11.70 GHz, when impinges an X-polarization plane wave. (a) Copolar phase response. (b) Losses. (c) Phase cut as a function of line L1. (d) Amplitude of the cross-polar component.

C. Beam scanning performance

For the antenna geometry under study, the phase distribution has been synthesized on the subreflectarray for two scan angles and the antenna is analyzed considering X_R polarization according to

Fig. 1 shows that the first case corresponds to the synthesis of a beam deflected –6 degrees in elevation. The phase distribution obtained by equation (3) is shown in Fig. 5(a). These phases are introduced in the analysis tool to compute first the electric field on the antenna aperture and then, the radiation pattern of the entire antenna. The amplitude and phase of the electric field on the antenna aperture are shown in Fig. 5(b) and (c), respectively. According to these results, the beam deflecting is achieved by scanning the field coming from the subreflectarray on the main reflector surface. The progressive phase is not obtained along the entire aperture of the antenna, see Fig. 5(c). However the required phase is achieved in the area with high illumination level, see Fig. 5(b) and (c), and therefore the beam is satisfactorily scanned. The zone of the aperture with low illumination will introduce a small distortion in the radiation pattern.

The second case corresponds to a beam scanned in the orthogonal dimension: +6 degrees in azimuth. As in the previous case, the required phase distribution on the subreflectarrray surface is first obtained and then introduced in the model to compute the electric field at the antenna aperture, see Fig. 6. Now, the illumination is scanned along the azimuth direction as shown in Fig. 6(b). As in the former case, the suitable progressive phase is achieved on the area of the antenna aperture with significant illumination level, see Fig. 6(b) and (c). Note that in this case the required phase-shift distribution shows some variation along the X_{RA} axis according to Fig. 6(a). The reason for this result is that the subreflectarray is tilted. If a progressive phase would be introduced along the Y_{RA} axis (with no variation on the orthogonal direction), the scanned beam will suffer a slight deflection in the vertical dimension, as shown in [24].

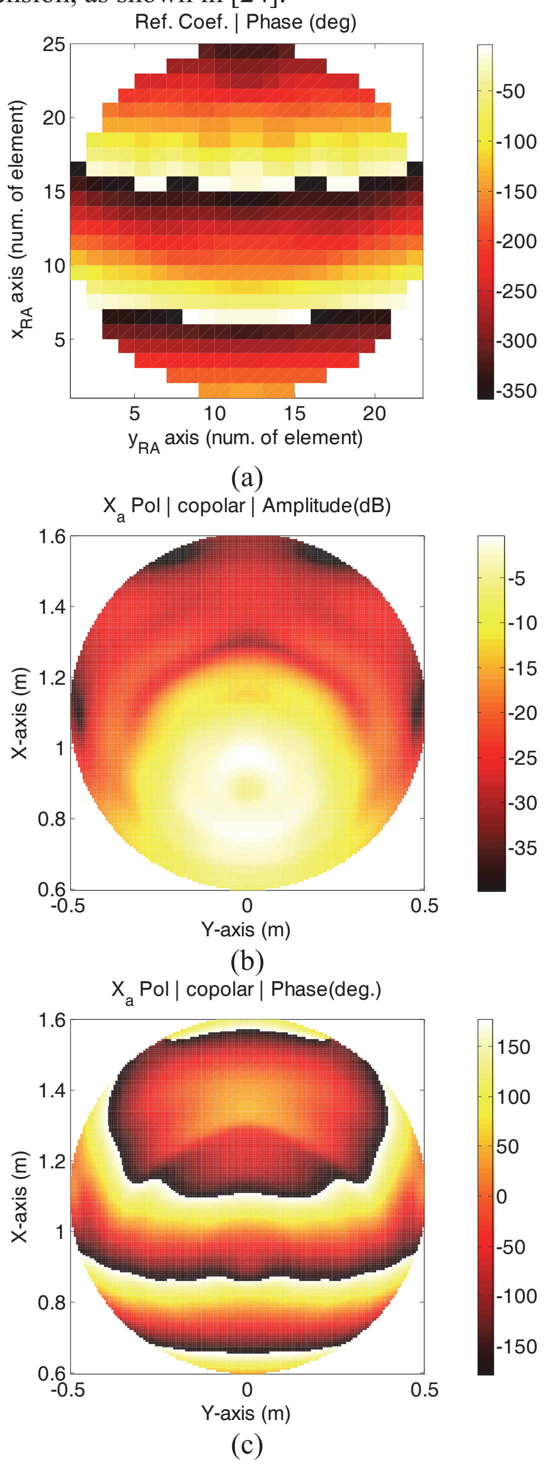


Fig. 5. Case of –6 degree deflected beam in elevation with continuous phase-shift control. (a) Required phase shift of the sub-reflectarray. (b) Amplitude and (c) phase of the electric field at the antenna aperture.

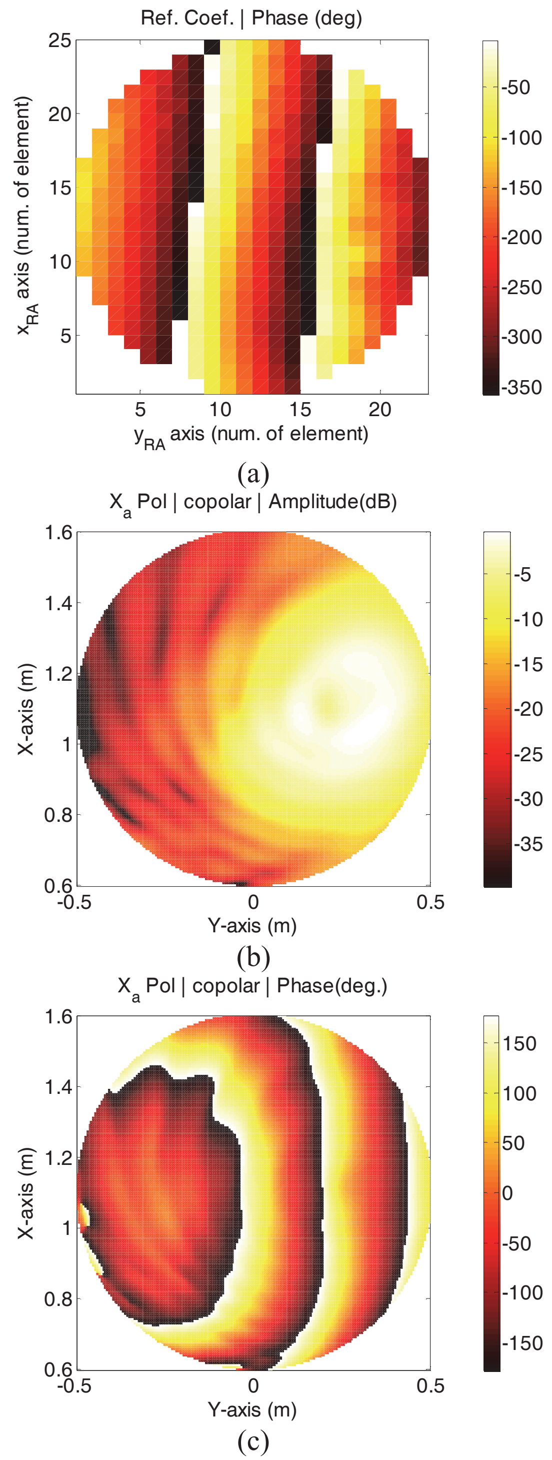


Fig. 6. Case of +6 degree deflected beam in azimuth with continuous phase-shift control. (a) Required phase shift of the sub-reflectarray. (b) Amplitude and (c) phase of the electric field at the antenna aperture.

The radiation patterns are computed for the two cases studied and they are compared with the case of the beam in the boresigth direction. The main cuts of the patterns, elevation and azimuth, are plotted in Fig. 7. A small reduction in gain with respect to the boresight beam is observed, 1dB in both cases, because of the reduction in the spillover efficiency. For the scanned beams, the spillover and aperture efficiencies are 0.7 and 0.29 respectively, being the total antenna efficiency 0.2. Note that the aperture efficiency is slightly lower than that obtained for the case of the beam in the boresight direction (0.31). In the case of the scanning in the azimuth plane, the cross-polar level is strongly increased (-14dB) since on the sub-reflectarray the beam is scanned out of the symmetry planes.

For a practical electronic control of the phase on the subreflectarray, the number of phase states is imposed by the number of control bits. Therefore, the phase distribution is quantized assuming n-bit control according to equation (4). In this case, 3-bit control is considered (eight phase states at each element). The discrete phase control implies a reduction in the degrees of freedom and some distortion in the radiation patterns is produced. The phase distribution for the case of +6degree deflected beam in azimuth and considering 8-level phase quantification is shown in Fig. 8. The corresponding radiation patterns obtained with these phase distributions are plotted in Fig. 9, showing a small distortion with respect to those assuming a continuous phase control, see Fig. 7.

The results have been obtained considering vertical polarization (X_a polarization), being the performance very similar for the orthogonal polarization.

IV. CONCLUSION

A dual-reflector antenna based on a reconfigurable reflectarray as subreflector has been presented as a solution for beam scanning applications that require high gain and narrow beams. In this antenna, the beam gain is obtained by a passive large parabolic reflector while the beam deflecting control is achieved by a small planar subreflectarray. The subreflectarray elements are based on printed patches aperture-coupled to delay lines where electronic control elements can be implemented. A synthesis technique, which is based on the analysis of the antenna in receive mode, has been applied to determine the phase-shift distribution on the subreflectarray surface for producing a given scanning angle. To validate the synthesis technique, an antenna has been defined at 11.7GHz and the beam scanning capabilities in both elevation and azimuth planes have been evaluated.

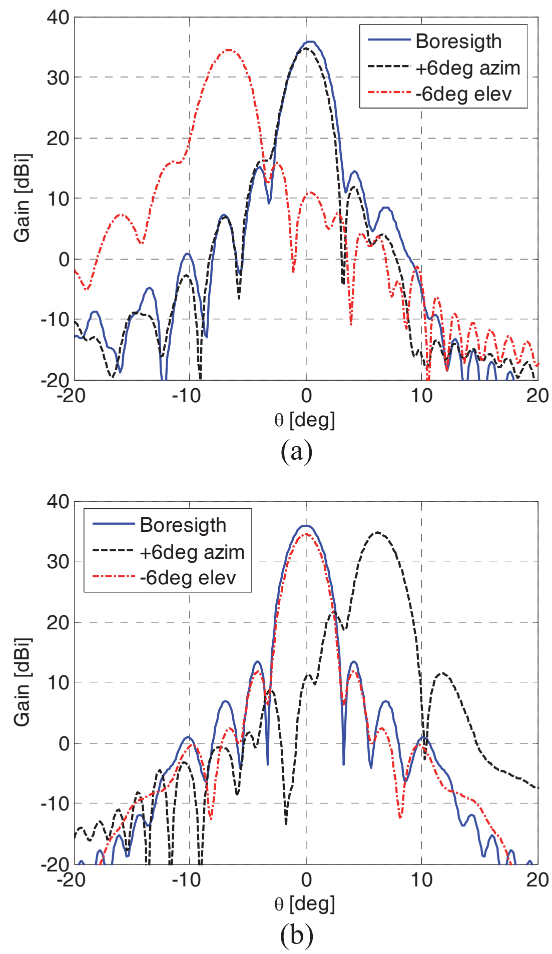


Fig. 7. Radiation patterns of different cases of scanned beams assuming continuous phase-shift control. Main cuts in (a) elevation and (b) azimuth.

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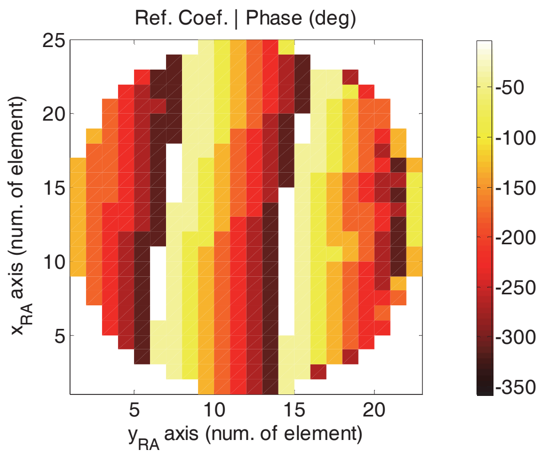
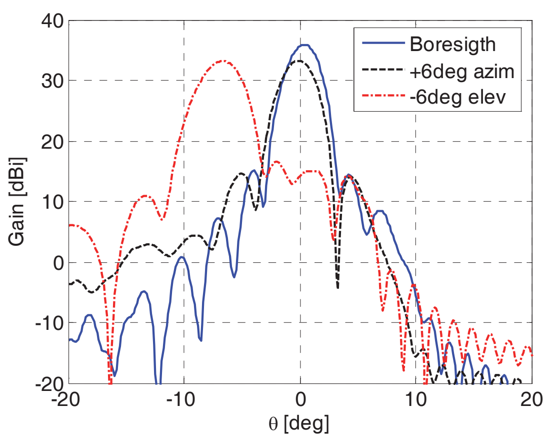
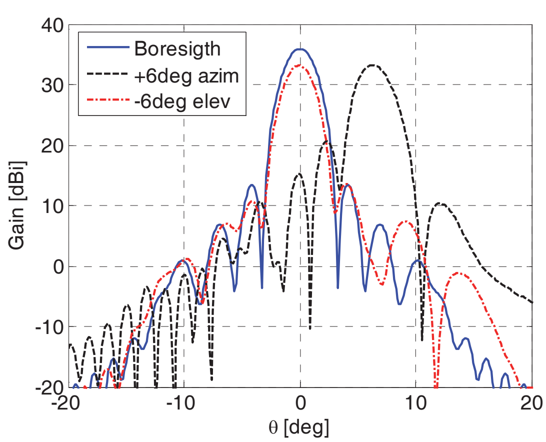


Fig. 8. Case of 3-bit phase-shift control. Required phase-shift for a beam deflected +6 degree in azimuth.



(a)



(b)

Fig. 9. Radiation patterns of different cases of scanned beams assuming 3-bit discrete phase-shift control. Main cuts in (a) elevation and (b) azimuth.

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