The Ultrawideband Leaky Lens Antenna

Simona Bruni, Member, IEEE, Andrea Neto, Member, IEEE, and Filippo Marliani

Abstract—A novel directive and nondispersive antenna is presented: the ultrawideband (UWB) leaky lens. It is based on the broad band Cherenkov radiation occurring at a slot printed between different infinite homogeneous dielectrics. The first part of the paper presents the antenna concept and the UWB design. The issues that are specifically addressed include the impedance matching, the radiation pattern and the phase purity. Subsequently the hardware demonstrator and the results of the measurement campaign are shown. The results fully validate the antenna concept. The antenna presents decade bandwidth impedance matching, directive and frequency independent patterns in the $H$-plane over two octaves, negligible cross-polarization levels, weak amplitude dispersivity and, according to the authors, the most stable phase center for a directive antenna over an ultra wide frequency range.

Index Terms—Cross polarization, pulse excited antennas, transfer functions, ultrawideband (UWB) antennas, ultrawideband radiation.

I. INTRODUCTION

RECENTLY, the United States Federal Communication Commission opened the spectrum from 3.1 to 10.6 GHz, permitting the use of a new unlicensed radio transmission technology, the ultrawideband (UWB) systems. UWB systems entail temporarily short pulses, below the nanosecond, and occupy a wide band with the only restriction to limit the radiated emissions levels. In addition the European Commission is following the same road and has been allocating the bands from 3.4 to 9 GHz for UWB systems.

Antennas act like filters and are critical elements in the signal flow for UWB systems. Antennas operating in UWB scenarios face the difficulty to maintain invariable performances, i.e., phase linearity, radiation efficiency and impedance match over large bandwidth. One typical problem is the “ringing effect,” i.e., the sub-nanosecond pulse is spread in the time domain and the wave form is altered resulting in the consequent deterioration of the time resolution of the system. Hence, the ideal impulse response shall show low amplitude and negligible phase distortion. This latter, being the most critical requisite, demands the conception of antennas with fixed phase centers and nondispersive feeding networks. Hence, base station antennas with high directivity may be used for high speed data networks.

Most, if not all, existing antennas can meet only a subset of the requisites listed above. They suffer from various types of limitations. One of the most successful examples, the Vivaldi antenna, typically trades off the bandwidth with high cross polarization levels (that can also be viewed as losses of efficiency) and the phase center stability, [1], [2], even more so when the aim is high directivity as in the long tapered slot (LTS) configurations[3]. In [4] these problems seem to be solved with a brilliantly arranged log periodic antenna that also achieves a moderate 11 dB directivity. However, a log periodic arrangement of dipoles intrinsically gives rise to phase dispersivity.

In [5], a novel UWB leaky lens antenna has been introduced. This antenna is weakly dispersive in amplitude and phase, polarization pure, reasonably efficient and directive. It is an evolution of the one presented in [6], where the first prototype of a leaky lens antenna had been presented. Leaky lenses are based on the almost frequency independent Cherenkov radiation mechanism occurring on a slot printed at the interface between two different dielectrics [7], [8]. If the two media are free space and a dense dielectric, radiation occurs primarily in the dielectric. This type of leaky radiation does not suffer from the typical dispersion effects that characterize all previously proposed leaky wave radiation mechanisms.

In this paper, the UWB leaky lens antenna and the physics behind its design are described in depth. The first part of the paper includes the antenna concept and the guidelines for the UWB design. The issues that are specifically addressed include the directivity, the impedance matching and the phase purity. The tradeoffs are described based on the discussion of physical considerations, analytical approximations as well as full wave numerical simulations. The hardware demonstrator and the results of the intense measurement campaign are subsequently shown. The results validate the antenna concept.

II. ANTENNA CONCEPT

The leaky wave radiation that occurs at an infinite slot etched on a ground plane that separates two different infinite dielectrics and is fed at a central point is intrinsically an UWB phenomenon. As explained in [8], in the canonical problem the radiated power is split in two conical beams. Each of the two beams forms an angle $\gamma_{AV}$ with the slot. This angle sets the shadow boundary line that defines the existence region for the leaky waves emerging from the slot, see Fig. 1(a). In order to achieve an unidirectional beam the two halves of the ground plane need to be tilted [Fig. 1(b)] as described in the following.
is the dimension transverse to the slot. Then, for $x_{z} = \sqrt{4/\pi}$, $x_{z} = \sqrt{4/\pi}$ is the wave-plane. Also, formed by a leaky wave where can be written can be approximated as follows:

$$\beta = \sqrt{\left(\frac{\omega_{s}}{k_{0}}\right)^{2} - 1}$$

where $\gamma_{1w}$ is the wave-plane, $\gamma_{1w}$ can be written

$$\gamma_{1w} = \cos^{-1}\left(\frac{\beta}{k_{0}}\right)$$

if the metallic plane is bent in correspondence of the feed, as shown in Fig. 1(b), the two shadow boundaries associated to the leaky waves coalesce and the resulting beam is unique at least in the $x, z$ plane.

In the bent structure, the diffracted fields, denominated space waves in [8], have a lower impact and will not be considered in the design. We will not give analytical proof of this operative hypothesis that instead will be validated by simulations and measurements. In the planar case, [8], the space waves emerging from the source, asymptotically play the role of providing continuity to the shape of the pattern radiated by the slot. Beyond the shadow boundary angles, see Fig. 1(a), the field would be zero if it was not for the space waves; thus a nonphysical discontinuity would exist. In Fig. 1(b) the shadow boundaries associated to each of the two leaky waves coalesce at broadside. When one leaky wave ceases to exist, it is replaced by the specular one, with respect to $x = 0$. Thus no discontinuity is to be found in the field in absence of space waves. For this reason there are essentially no space waves. This picture is valid in the $x, z$ plane, but it is more involved in the 3D case as will be discussed in Section IV-A.

A. Lens Description

A practical implementation of this concept is depicted in Fig. 3. The lens geometry can be thought of as the union of two dielectric lenses. Each lens is obtained as a union of an infinite

![Fig. 1. Side view ($H$-plane cut) of the leaky waves existence zones: (a) in a planar structure (b) in a bent structure.](image)

![Fig. 2. Normalized propagation constant (a) and attenuation constant (b) of the leaky wave as a function of the frequency when $\epsilon_{r1} = 1, \epsilon_{r2} = 3.27$ and $w_{s} = 0.25$ mm.](image)

To estimate the main beam angles one can approximate the magnetic current distribution in each cross section of the slot as

$$m_{y} = \frac{2}{w_{s} \pi} \frac{1}{\sqrt{1 - \left(\frac{2u}{w_{s}}\right)^{2}}}$$

where $y$ is the dimension transverse to the slot, [7]. Then, for slots sufficiently narrow, $w_{s} < \lambda_{2}/2\pi$ where $\lambda_{2}$ is the wavelength in the denser medium, the leaky wave propagation constant, $k_{x}^{\text{LW}}$ can be approximated as follows:

$$k_{x}^{\text{LW}} \approx \beta - \frac{D(\beta)}{D'(\beta)}$$

where

$$D(k_{x}) = \frac{1}{2\epsilon_{0}k_{0}} \sum_{i=1}^{2} \left(\alpha_{2i} - k_{0}^{2}\right) J_{0} \left(\frac{w_{s}}{4\sqrt{k_{x}^{2} - k_{0}^{2}}}\right) H_{0}^{2} \left(\frac{w_{s}}{4\sqrt{k_{x}^{2} - k_{0}^{2}}}\right).$$

Here $k_{0}, \epsilon_{0}$ are the propagation constant and characteristic impedance of free space and $k_{i}, \epsilon_{ri}$ for $i = 1, 2$, are the propagation constants in the two media of dielectric constant $\epsilon_{ri}$. Also, $\beta = \sqrt{(k_{x}^{2} + k_{0}^{2})/2}$ and the derivative of $D(k_{x})$ can be written in analytical form (see [7]).

The real and imaginary parts of the propagation constant as a function of the frequency are shown in Fig. 2(a) and (b), for $\epsilon_{r1} = 1, \epsilon_{r2} = \epsilon_{r} = 3.27$ and $w_{s} = 0.25$ mm.
set of cross section planes of truncated elliptical shape and decreasing dimension. In the lower focus, each of the ellipses contains a slot etched on a ground plane. The eccentricity of each ellipse is \( \epsilon = 1/2 \), with \( \epsilon_r \) the dielectric constant of the lens. In correspondence with the junction of the two slots and of the two original lenses, the shape of each one of the lenses is altered in order to host a ground plane of width \( g \) (see the side view in the inset of Fig. 3). This ground plane hosts the coplanar waveguide (CPW) transmission line that will feed the unique slot.

For the sake of convenience the origin of the reference system is in the center of the CPW ground plane.

In Fig. 3, a schematic and simplified representation of the radiation inside the lens is given for each of the elliptical cross sections. One can imagine that each of the leaky wave rays: a) emanating from one point in the slot as explained in [8], lies in the elliptical cross sections. Resorting to the description introduced in [9] they can then be imagined to undergo the following path: they are b) transmitted after the first interface (focusing effect in the far field); c) reflected at the first interface; d) transmitted at the second interface (unfocused); and e) reflected at the second interface and refocused toward the slot. The first transmitted rays, which carry most of the power due to small reflection at the lens-air interface, are all focused in the \( z \) direction. Asymptotically, only doubly reflected rays, which already lost power in the first two transmissions, return to the focus (slot line). This guarantees that the present lens simulates well the ideally infinite dielectric configuration discussed in [7], characterized by very low frequency dispersivity. The reader should notice that, although simplified and therefore approximate, as will be explained in Section IV-A, the ray description of Fig. 3 has constituted an essential support for the design of the UWB leaky lens antenna.

III. SLOT DIMENSIONING

The width and the length of the slot, combined with the dielectric constant, constitute the parameters to be fixed in the design of the antenna illustrated in Section II. In order to prove the validity of the antenna concept it has been decided to limit the maximum dimension of the lens \( L_{\text{lens}} \) to 16 cm. Consequently each half slot was taken to be 13 cm. As a general rule the dielectric constant should be relatively low so that reflections from the dielectric air interface would be minimized. A reasonable choice appeared to be \( \epsilon_r = 3.27 \), which is commercially available and suited to be machined. Finally, the slot’s width remained to be selected.

The attenuation constant of the leaky wave propagating along the slot is significantly influenced by the width of the slot and the frequency, as shown in Fig. 2. In order to minimize the truncation effects, it is good engineering practice to size the slot (length and width) in such a way that the power at the edges is attenuated to one tenth with respect to the feed level (end-point requirement). The end points constitute a source of space waves which contributions appear relevant especially at lower frequencies. In our design, we have relaxed the requirements on the pattern purity and aimed at a good match over a large frequency band, one decade. Consequently, we have sized the width of the slot to guarantee the end-point requirement only above 7.5 GHz but a good 50 ohm match between 4 and 40 GHz, as explained in the following section.

A. Matching

As anticipated, the apparent degree of freedom in the slot’s width disappears when the impedance matching of the slot is accounted for. In the design of an antenna aimed at operating over a decade bandwidth, it is not necessary to concentrate on second order details as one would typically do when dealing with narrow band antennas. Consequently, full wave simulations are out of the scope. Instead, we have developed a simplified and analytical formula for the input impedance of the leaky slot line that was used in the optimization of the CPW to slot transition.

The bulk of the current that is fed to the long slot flows parallel to the slot and supports two leaky waves that propagate in opposite directions (outgoing from the source). Accordingly, the input impedance of the long slot can be approximated as the parallel of two equal characteristic impedances each associated to one of the leaky waves, see Fig. 4. Overall, the input impedance of the center fed slot can be expressed as

\[
Z_{\text{in}} = \frac{Z_0}{2},
\]

The characteristic impedance of a leaky transmission line can be evaluated resorting to the method discussed in [10]. In the
and this implies that the $\varepsilon_r$-plane pattern weakly dependent on $\varepsilon_r$ after a refraction at the di-plane with respect to Fig. 8 (note $\varepsilon_r$-plane PLANE RADIATION PATTERN, which has differences between the results of the two commercial tools are to 40 GHz.

be expressed as present case the evaluation turns out to be analytical and it can be expressed as

$$Z_0 \approx \frac{-2j}{D'(k_z^{SW})}, \quad (5)$$

$D'(k_z^{SW})$ is proportional to $k_z^{SW}$ and this implies that the characteristic impedance has both a real and imaginary part, while nonleaky TEM lines would have purely real characteristic impedances.

The characteristic impedance curve as function of the width in terms of the wavelength is shown in Fig. 5. For a slot width $w_s = 0.25 \text{ mm}$, Fig. 6 shows the approximation of the input impedance of the slot for the frequencies from 4 to 40 GHz. This result demonstrates that a CPW with characteristic impedance has both a real and imaginary part, while nonleaky TEM lines would have purely real characteristic impedances.

In order to validate this design procedure the antenna described has been analyzed using two different commercial tools. The 3D tool HFSS has been used up to a frequency of 10 GHz, while for higher frequencies a completely planar structure has been analyzed using Ansoft Designer. The reflection coefficient referred to the 50 Ohm line is shown in Fig. 7. It is apparent that the matching is good for frequencies higher than 4 GHz. The most significant differences between the results of the two commercial tools are associated to the double reflections occurring at the dielectric air interface [9], which are not included in the Designer simulations, while they induce oscillations (period less than 1 GHz) in the reflections coefficient as a function of the frequency in the HFSS results. The same oscillatory behavior had been observed in [6]. In addition to the two numerical simulations the prediction based on our analytical expression, (4), is reported. It is clear that our approximation matches fairly well the results of Designer. Small differences arise for the lowest frequency ranges (in which case the finite length of the slot also introduces an oscillation with frequency) and for the higher frequency ranges, because designer simulations have been performed accounting for the CPW slot transition; the small piece of CPW lines introduces the very wide oscillations (15 GHz period).

IV. $\Pi$-PLANE RADIATION PATTERN

The shape of the lens described in Section II-A was chosen because it guarantees an $\Pi$-plane pattern weakly dependent on the frequency. Since this is a very unusual characteristic for directive antennas, this section will provide a physical mathematical demonstration of this property. A model for the $E$-plane pattern will not be given because in this plane the pattern is frequency dependent as similar dielectric antennas of elliptical cross section, [11].

A. Aperture Integration

With reference to the smaller elliptical cross sections drawn in Fig. 8, the lens edges can be identified via a subtended angle, $\theta_{sub} = \cos^{-1}(e) = \cos^{-1}(1/\sqrt{\varepsilon_r}) = 56^\circ$. All rays characterized by angles smaller than $\theta_{sub}$, after a refraction at the dielectric air interface are focused in broadside. The secondary radiation pattern, the one outside the lens, can be evaluated resorting to the standard aperture integration technique described [13, Ch. 3 and 12]. The procedure involves the evaluation of the Fourier transform (FT) of the electric and magnetic currents in the tetragonal aperture representing the projection of the lens surface on the $z = h$ plane with respect to Fig. 8 (note $L_x = L_{tense}$. In first approximation, it is assumed that the field outside such tetragon is equal to zero.

An accurate estimate of the equivalent currents on the aperture plane would involve the ray tracing of all rays propagating
planes, which renders \(-\)plane patterns with measured results in Section VII. However, it can be anticipated that, since the central portion of the lens is less excited than our simplified picture imagines (see Fig. 10 and the grey area), the amplitude of the \(H\)-plane patterns are overestimated by our simplified model, especially at high frequencies.

**B. \(H\)-Plane Pattern Analytical Formula**

The approximate procedure (see Appendix) to evaluate the far field in the \(H\)-plane eventually leads to

\[
E_\phi(r, \theta, \phi = 0) \approx A(r)(I_1 + I_2)
\]

where

\[
A(r) = I_f jk_0 e^{-jk_0 r} \frac{1 + \cos \theta}{4\pi r} e^{j \pi/4} e^{-jk_0 h} \sqrt{\frac{k_0^2}{2\pi h}} T
\]

where \(I_f\) is an unknown multiplication constant that accounts for the amplitude overestimation. \(I_1\) and \(I_2\) can be expressed analytically as follows:

\[
I_1 = L_y \left( \frac{e^{UL_x/2} - 1}{U} + \frac{e^{UL_x/2} - 1}{U^*} \right)
\]

\[
I_2 = -2L_y \left\{ \left( \frac{e^{UL_x/2} - 1}{U} - \frac{e^{UL_x/2} - 1}{U^*} \right) \right\}
\]

where \(U = -\delta_{LM}/\sin \gamma_{LM} + jk_0 \sin(\theta)\) and the asterisk denotes complex conjugation.

The approximations adopted in deriving (6) imply that its validity is restricted to the main beam region, where the radiation mechanism described is dominant. Fig. 11(a) shows the beam widths obtained using (6) for a test case characterized by \(L_y = 0.108\) \(\text{m}\), \(L_x = 0.16\) \(\text{m}\), \(\epsilon_r = 3.27\), \(\mu_r = 0.25\) mm. The curves pertinent to two different frequencies, 16 and 32 GHz, are reported after having being normalized in amplitude. One can notice that the patterns are essentially superimposed on this octave. The reason for this high frequency behavior is not obvious and some clarifications are in order. The attenuation constant for the present slot as a function of the frequency was shown in Fig. 2(b). It is apparent that, unlike the pointing angle, in the lens. The tracking of all direct, reflected and multiple reflected rays can be quite cumbersome. Fig. 9 shows a realistic picture of the direct rays incrementally radiating from the slot, inside the lens. These rays, at difference with the simplified ones in Fig. 3, constitute conical beam whose centers lay on the slot. In order to estimate the field on the integration aperture with higher precision, the relevant algorithm should resort to an accurate ray tracing (direct rays and rays with multiple reflections) that would become numerically cumbersome. In this paper it was decided that, in order to estimate the equivalent currents, only the rays directly emerging from the slots and refracted into the aperture plane, would be considered. Moreover, it was also decided to adopt the simplified ray picture of Fig. 3, with all rays lying in the \(x = \text{constant}\) planes, which renders the tracing of such rays analytical. In fact, the simplified ray picture implies constant phase on the aperture plane for every \(x\) coordinate as in the right hand side of Fig. 10, at difference with the realistic ray picture shown in the left of the same figure.
it is not a very slowly varying function. Its amplitude grows by a factor 1.25 over an octave, suggesting the beam width at 32 GHz should be 25% larger than the one at 16 GHz. If the slot’s dispersivity was the only important effect present, only the first of the two terms in (6), $I_1$, would represent the radiated field and the results would be those reported in Fig. 11(b).

However, the lens is tapered in such a way that the elliptical cross sections are smaller as they are closer to the ends of the slot. The rays emerging from different points undergo transmissions through larger or smaller elliptical cross sections. For lower frequencies the fields emerge from the entire slot and consequently the radiation is heavily affected by the lens tapering, represented by the $I_2$ term in (6), which causes broader beams. For higher frequencies the fields are predominantly radiated by the central part of the slot and thus they are not heavily affected by the tapering of the lens. Thus, in practice, the lens tapering tends to compensate for the dispersivity in the attenuation constant of the slot as a function of the frequency.

V. ANTENNA MANUFACTURING

A prototype of the UWB leaky lens antenna has been manufactured at the workshop of TNO. The lens was obtained as the union of two half lenses of material TMM-3 from Rogers ($\varepsilon_r = 3.27$). The dielectric was machined mechanically to assume the characteristic elliptical cone shape and the slots were etched on the outer flat slab, which is covered in metal. The shaping of the two half lenses is done in such a way to leave some space in the region of the common ground plane between the two lenses (insertion region). The CPW feeding line, which is machined on a separate small piece of the same dielectric, is then slid in the insertion region. In order to achieve 50 Ohm characteristic impedance and for the structure to be only weakly dispersive at the maximum operating frequency of 40 GHz, the slots composing the CPW had to be very small. The CPW lines printed between two infinite dielectrics also define a leaky type of propagation and its characteristic impedance was discussed in the Appendix of [15]. In the presence of a CPW printed between air and a dielectric of $\varepsilon_r = 3.27$, $a = 0.5 \mu m$ and $w_{CPW} = 0.05 \mu m$ are appropriate. This was also the minimal slot’s width that could be machined by TNO. Finally, the CPW line is connected to a SK coaxial connector. The final dimension of the UWB leaky lens are given in Table I and the antenna prototype is shown in Fig. 12.

VI. REFLECTION COEFFICIENT MEASUREMENTS

A measurement campaign was performed to evaluate the performances of the UWB leaky lens. The first parameter that was measured was the reflection coefficient, which is shown in Fig. 13 in the band from 1 to 50 GHz.

The reflection coefficient is below 8.5 dB over the entire design band (4–40 GHz). The measured results match very well with the expectations that were based on the calculations shown in Fig. 7, except for the fact that in Fig. 13 the actual length of the CPW transmission line was included. The only significant difference between calculations and measurements is that the small period oscillations due to the double reflections inside the lens, [9], are now visible over the entire frequency band. The main observation from Fig. 13 is that the amplitude of the oscillations is much stronger for lower frequencies than for higher frequencies.

The reason for this frequency dependence is that the realistic ray paths highly resemble the simplified ones in Fig. 3 at low frequency and thus double reflections are coherent. At high frequency, however, the various doubly reflected rays to the feed differ significantly in phase and therefore do not add coherently. The most important consequence of the double reflections can
TABLE I
GEOMETRICAL DIMENSIONS OF THE UWB LEAKY LENS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Actual Dimension</th>
</tr>
</thead>
<tbody>
<tr>
<td>$w_x$</td>
<td>0.25 mm</td>
</tr>
<tr>
<td>$L_e$</td>
<td>130 mm</td>
</tr>
<tr>
<td>$\gamma_{le}$</td>
<td>35°</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>$90^\circ - \gamma_{le} = 55^\circ$</td>
</tr>
<tr>
<td>$l_{Lens}$</td>
<td>160 mm</td>
</tr>
<tr>
<td>$g$</td>
<td>3 mm</td>
</tr>
<tr>
<td>$l_{GND}$</td>
<td>85 mm</td>
</tr>
<tr>
<td>$w_{GDL}$</td>
<td>0.05 mm</td>
</tr>
<tr>
<td>$a$</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>$c_{L}$</td>
<td>3.27</td>
</tr>
<tr>
<td>$b$</td>
<td>$\approx 100$ mm</td>
</tr>
<tr>
<td>$\theta_{slb}$</td>
<td>56°</td>
</tr>
<tr>
<td>$l_x$</td>
<td>160 mm</td>
</tr>
<tr>
<td>$l_y$</td>
<td>108 mm</td>
</tr>
</tbody>
</table>

Fig. 13. Measured and calculated reflection coefficient of the UWB leaky lens as a function of the frequency.

be found in the radiated fields. The radiation of a broad band time pulse is followed by radiation associated to the double reflected rays, which is roughly delayed for each different cross section by $4 S_m(x)/c_d$, with $c_d$ the phase velocity in the dielectric and $S_m(x)$ defined in Fig. 8 as the major semi-axis of the elliptical cross section at $x$. The coherence of these doubly reflected pulses renders the ringing radiation significant only for the lower frequencies. The bigger the lens is in terms of the wavelength, the smaller is the ringing.

VII. RADIATION PATTERNS MEASUREMENTS

The measurements of the radiation patterns were performed in the Compact Antenna Test Range (CATR) of the European Space Agency Research and Technology Center (ESTEC) in Noordwijk, the Netherlands. Both co- and cross-polar components in the $E$, $H$ and $D$ (diagonal) planes were measured. The results are shown in the following sections.

A. $E$-Plane

The co-polar radiation patterns in the $E$-plane are shown in Fig. 14(a) and (b).

The patterns are mostly symmetric with respect to broadside. While for low frequencies the patterns are very large, the beam width decreases linearly with frequency, which is congruent with a diffraction limited pattern. The measured side-lobes however are higher than one could wish for. The side lobes become higher as the frequency increases, as high as $\approx 4$ dB at 40 GHz. With reference to the smaller elliptical cross sections drawn in Fig. 8, $\theta_{slb} = 56^\circ$ defines the edges of the lens. $\theta_{slb}$ corresponds to the critical incidence angle for the dielectric air interface, at which point the reflection coefficient is equal to one. In this situation the prediction of a simplified geometrical optics [9] algorithm would imply no transmission through the lens. An analysis that included noticeable diffracted contributions emerging from the edge of the lens [12] would explain
the higher side lobes, as the frequency increases. This diffraction was not a major problem in integrated lens designs that involved directive feeds such as twin slots [11]. However, the feed is now a single slot on focus and thus nondirective in each (z, y) cross section. Thus, even if the $-3\text{dBi}$ angles in the $E$-plane decrease linearly with frequency the directivities in the $E$-plane are in general much lower than imagined by the simplified model (right hand-side of Fig. 10).

Once the focus is on the elliptical cross sections it is worth noting that from simple geometrical considerations it is evident that rays impinging on the dielectric air interface for angles larger than $56^\circ$ certainly do not contribute to the main beam as they are not focused broadside. This implies that 37% of all the power launched by the slot is not radiated in the main beam but scattered in arbitrary directions. This power can be considered lost and thus, using a terminology typical of reflector antennas, one can estimate the spill over efficiency, [13], of this lens at 63%. Any other slot configuration, for instance a design involving two parallel long slots, would reduce the spill over and the side lobes, but the enhancement would have to be traded off with the bandwidth of the antenna.

The cross polar radiation patterns in the $E$-plane are shown in Fig. 14(c). These patterns are at least 25 dB below the co-polar ones.

B. $H$-Plane

The amplitudes and phases of the radiation patterns in the $H$-plane are shown in Figs. 15 and 16. The patterns in the $H$-plane are essentially symmetric with respect to broadside. Again, at low frequencies they are much wider than at higher frequencies. This is mostly due to the fact that the field levels at the end of the slot are still very high for low frequencies. As the frequencies increase, the behavior of the antenna becomes asymptotic, as described by the two asymptotic contributions in Section IV-B. The much desired frequency independent pattern can be observed for frequencies higher than 10 GHz. In this plane the side lobes are 15 dB lower than the main lobes. This is congruent with the fact that the outer part of the lens (towards the end points of the slot) is weakly excited. The cross polarized radiation patterns in Fig. 15(c) are still low ($-20\text{dBi}$ with respect to the co-polar levels) even though not as low as in the $E$-plane.

The upper part of Fig. 16 shows a detail of the measured co-polar radiation patterns in the $H$-plane at three frequencies covering 2 octaves (8–16–32 GHz). These patterns are compared with the patterns predicted by the simplified model given in Section IV-B and already shown in Fig. 11(a). It is apparent that the main beams predicted analytically are congruent with the measured ones. In fact, the $-3\text{dBi}$ beam widths differ in all cases by less than 1 degree.

The lower part of Fig. 16(b) presents the measured phases of the radiation patterns in the main beam, normalized to zero at broadside. The phase center of an antenna at a given frequency $f_j$, is the geometrical point in space from which all waves appear to emanate with a spherical spreading. It is apparent that the antenna presents the same phase center at all frequencies and thus it would be well suited to excite a reflector over two octaves of BW. The phase center is localized at the origin of the $(x_0, y_0, z_0)$ reference system defined to evaluate the radiation patterns in Fig. 8.

Another example of directive antenna in which the Fourier Transform of the aperture distribution is completely real, is a uniformly excited reflector. However, in reflector design, phase errors typically occur because realistic feeds present phase centers that move with frequency. In the present UWB leaky lens the entire path from feeding transmission line to radiating aperture has been considered and obvious sources of phase distortion could not be identified (except for weak double reflections inside the lens). According to (6) the only variation of the phase
of the field as a function of the frequency is $e^{-j(k_0 r + k_2 h)} = e^{-j2\sqrt{\mu_0 \sigma_0 (r + \sqrt{\sigma_0})}}$. A linear variation of the phase as a function of the frequency corresponds to a pure time delay. The antenna can therefore be described as nondispersive in phase. Within the limits in which double reflections are neglected, the UWB leaky lens radiation described by (6) should also be capable of preserving time domain pulses. This specific point is the object of an ongoing investigation that will include time domain measurements.

C. D-Plane

The amplitudes and phases of the radiation patterns in the D-plane are shown in Fig. 17(a)–(c), presenting the co and cross polar components. The co-polar patterns are an average of the $E_r$ and $H_r$ plane and are thus well behaved. The cross polar patterns are higher than in the main plane leading to a peak with respect to the co-polar components as typical value. These values are still excellent when compared with those associated to other UWB directive antennas [3] and [14].

D. Directivity

The beam widths of the antenna on the three main planes as well as the measured directivities are detailed in Table II for different frequencies. It is worth noting that the directivity estimated from the 3 dB beam-width only would be significantly higher than that calculated from measurements, accounting for the patterns for all fields of view, thus revealing that the UWB leaky lens antenna loses considerable power in the secondary lobes. The measured directivity in the direction of maximum radiation, $\theta = 0$, is compared (Fig. 18) with the measured absolute gain. This latter could only be obtained accurately until 16 GHz. The overall result is that the directivity essentially saturates around 13 GHz. The measured directivities are $19 \pm 1 \text{dB}$ from 13 to 40 GHz. The reason for this has been traced to the actual field distribution in the radiating aperture. From the realistic ray picture in Fig. 10 it is clear that the excitation of the central cross section ($x_{0c} = 0$) of the lens is widely nonuniform especially at the highest frequencies. Direct leaky waves from the slot do not reach the zone of the radiating aperture depicted in grey. At low frequency, the presence of space waves emerging directly from the feed point compensates the absence of direct waves.
leaky rays, rendering the aperture distribution roughly uniform (at least in amplitude). However, even if the ray picture is frequency independent, the extension of the region where the space wave compensation is important becomes smaller as the frequency increases. Thus in the higher frequency regime (above 20 GHz), an increasingly significant portion of the central part of the lens is under-illuminated. For this reason the antenna is probably less dispersive in amplitude than all known broad band directive antennas. Directivity and gain are very similar up to 16 GHz, indicating negligible dielectric losses, at least in the lower frequency regime.

VIII. CONCLUSIONS

The measurements of the hardware presented in this paper show that the UWB leaky lens antenna has the following:

- Good impedance matching over a decade bandwidth;
- Directive radiation patterns that are frequency independent in the H-plane over two octaves;
- The most stable phase center for a directive antenna over two octaves;
- The smallest amplitude dispersivity for a directive antenna;
- Negligible (with respect to other UWB nondispersive antennas) cross polarization levels.

These characteristics alone render this antenna one of the best performing UWB antennas. In particular, they render this antenna the most suited to excite reflector antennas characterized by moderate to high F/D over and an ultra wide frequency band. The wave mechanisms described in this paper also indicate very weak phase dispersivity. To this regard a dedicated study of the time domain pulse preservation of UWB leaky lenses is the subject of ongoing study.

APPENDIX

To evaluate the Fourier transform (FT) of the electric and magnetic currents in the radiation aperture of Fig. 8, it is convenient to introduce an aperture reference system $(x_a, y_a, z_a)$ centered at $z = h$ and parallel to the original one $(x, y, z)$. The domain of integration is then a tetragon of major lengths $L_x$ along $\hat{i}_x$ and $L_y$ along $\hat{i}_y$, see Fig. 8.

Assuming negligible cross polarization, which is verified by measurements in Section VII and makes the procedure simpler, the electric fields oriented along $\hat{i}_{y_a}$ as a function of the observation angle $\theta_a$ in the $H$-plane ($\phi_a = 0$) can be expressed as

$$E_{y_a}(r, \theta_a, \phi_a = 0) = jk_0 \frac{e^{-jk_0r}}{4\pi r} \left\{ 1 + \cos(\theta_a) \right\},$$

$$\int_{-L_x/2}^{L_x/2} \int_{-L_y(x_a)/2}^{L_y(x_a)/2} E_{y_a}^o(x'_a, y'_a, z'_a = 0) e^{jk_0x'_a x''_a} dx'_a dy'_a$$

where $E_{y_a}^o$ is the electric field component along $\hat{i}_{y_a}$ after the transmission through the dielectric air interface, $L_y(x'_a) = L_y(1 - 2|x'_a|/L_x)$ defines the domain of the projection of the lens onto the radiating aperture along $y_a$ and $k_{x_a} = k_0 \sin(\theta_a)$. The distribution of the electric field, $E_{y_a}^o$, on the aperture plane, for each $x'_a$, will be assumed to be constant with respect to $y'_a$:

$$E_{y_a}(x'_a, y'_a, z'_a) \approx E_{y_a}^o(x'_a, y'_a = 0) \approx E_{y_a}^o(x'_a, y'_a = 0).$$

This is a brutally simplifying hypothesis. However due to the symmetry of the geometry with respect to the $x_a, z_a$ plane, the pattern in the $H$-plane is essentially insensitive to the distribution along $y_a$. Using this approximation, the inner integral in (9) can be evaluated analytically leading to

$$E_{y_a}(r, \theta_a, \phi_a = 0) \approx jk_0 \frac{e^{-jk_0r}}{4\pi r} \left\{ 1 + \cos(\theta_a) \right\},$$

$$\int_{-L_x/2}^{L_x/2} E_{y_a}^o(x'_a, y'_a = 0, z'_a = 0) L_y \left[ 1 - \frac{2|x'_a|}{L_x} \right] e^{jk_0x'_a x''_a} dx'_a.$$

In order to actually obtain the radiated field from (10), the distribution $E_{y_a}^o(x'_a) = E_{y_a}^o(x'_a, y'_a = 0, z'_a = 0) = 0$ should be evaluated. This task can be rendered simpler if one uses the two following simplifying hypotheses:

- Reflections from the end point of the slots will be neglected. In the present design this hypothesis limits the validity of the model to frequencies higher than 8 GHz.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>$BW_E$ (%)</th>
<th>$BW_H$ (%)</th>
<th>$BW_D$ (%)</th>
<th>Directivity (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>73</td>
<td>34</td>
<td>46.5</td>
<td>7.5</td>
</tr>
<tr>
<td>5</td>
<td>29.5</td>
<td>27.5</td>
<td>28</td>
<td>12.4</td>
</tr>
<tr>
<td>6.4</td>
<td>24</td>
<td>22.2</td>
<td>22.2</td>
<td>13.56</td>
</tr>
<tr>
<td>8</td>
<td>21.6</td>
<td>20</td>
<td>20.2</td>
<td>15.2</td>
</tr>
<tr>
<td>10</td>
<td>17.8</td>
<td>17.5</td>
<td>17</td>
<td>16.41</td>
</tr>
<tr>
<td>12.8</td>
<td>13.4</td>
<td>14.5</td>
<td>13.4</td>
<td>17.68</td>
</tr>
<tr>
<td>16</td>
<td>12.3</td>
<td>14.4</td>
<td>12.9</td>
<td>18.76</td>
</tr>
<tr>
<td>20</td>
<td>8.6</td>
<td>12.9</td>
<td>10.06</td>
<td>19.97</td>
</tr>
<tr>
<td>25.6</td>
<td>7.5</td>
<td>14.2</td>
<td>8.9</td>
<td>18.97</td>
</tr>
<tr>
<td>32</td>
<td>6.9</td>
<td>14.2</td>
<td>8.6</td>
<td>19.02</td>
</tr>
<tr>
<td>40</td>
<td>7.1</td>
<td>19</td>
<td>10</td>
<td>17.82</td>
</tr>
</tbody>
</table>
Reflections at the dielectric air interface will be neglected, so that the field distribution in ($x_1, y_1, z_1 = 0$), except for a transmission coefficient, $T$, is assumed to be the same as if each slot was printed between air and an infinite dielectric.

With reference to Fig. 19, the leaky wave emanating from the slot located for $x > 0$ is more conveniently expressed in the local reference system ($x_L, y_L, z_L$), which has $x_L$ axis parallel to the slot. In [8], the $x_L$-oriented electric vector potential, $\Psi$ was derived. It was radiated by the slot when this was fed by a unitary current source at the center of a planar canonical structure. The potential would present space wave and leaky wave components. As in the present case the space waves are considered to be negligible, the focus will be on the leaky waves only, $\Psi^{\text{LM}}$.

The electric field can be derived by applying $E = -\nabla \times \Psi^{\text{LM}}$, and the $y$ component reduces to

$$E_{yL} = -\frac{\partial \Psi^{\text{LM}}(x_L, z_L)}{\partial z_L}.$$

The partial derivative can be performed analytically and provides two terms, one that decays as $\zeta^{-3/2}$, which can be asymptotically neglected, and a second dominant one $\zeta^{-1/2}$. The expression of the $E_{yL}$ as a function of the position in the aperture plane can then be obtained analytically by means of simple, but tedious algebraic manipulations leading to

$$E_{yL}(x_L, z_L = 0) \approx I_f e^{j\pi/4} D(\zeta^{2/3}) e^{-j k_2 h / \sin(\gamma_L)}$$

$$\left\{ \frac{1}{2\pi h\sin(\gamma_L)} \right\} e^{-j k_2 h / \sin(\gamma_L)} e^{-j \delta_L / \sin(\gamma_L)} (2)$$

where $I_f$ is the feed electric current. The distribution presents constant phase and has an amplitude that is essentially exponentially attenuated. The presence of a cylindrical spreading at the denominator can be neglected in order to render the evaluation of (10) more easy

$$e^{-j k_2 h / \sin(\gamma_L)} \approx e^{-j k_2 h / \sin(\gamma_L)} = \sqrt{h\sin(\gamma_L)}.$$

This latter simplification is partially justified, at least at high frequencies, by the fact that when $x_L$ at the denominator becomes large, the numerator is already strongly dumped, so that the contribution of the spreading is rarely noticeable.

Finally, one should include a transmission coefficient to account for the dielectric air interface. This transmission coefficient can be assumed to be uniform and equal to the value assumed for normal ray incidence in first approximation: $T = 1 + \Gamma$ and for $\varepsilon_r = 3.27, \Gamma \approx 0.3$.

An estimation of the $H$-plane far field can then be obtained substituting (12) and (13) in (10)

$$E_{\phi_L}(r, \theta, \phi, \Phi = 0) \approx A(r)(I_1 + I_2)$$

where

$$A(r) = I_f \int \frac{e^{-j k_0 r / \sin(\gamma_L)}}{4 \pi r_L} \frac{e^{-j k_2 h / \sin(\gamma_L)}}{D(\zeta^{2/3})} \sqrt{\frac{k_2}{2\pi h}} \cdot T$$

and

$$I_1 = I_L \int_{-L/2}^{L/2} \frac{e^{-j (\delta_L / \sin(\gamma_L)) x_L e^{j k_0 x_L} d x_L}}{I_{x_L}^{2/2}}$$

$$I_2 = -\frac{2 I_{x_L}}{I_{x_L}} \int_{-L/2}^{L/2} \frac{e^{-j (\delta_L / \sin(\gamma_L)) x_L e^{j k_0 x_L} d x_L}}{I_{x_L}^{2/2}}$$

The two integrals $I_1$ and $I_2$ can be performed analytically and can be expressed as follows:

$$I_1 = I_L \left\{ \frac{U_{L2} - 1}{U_{L2} - 1} + \frac{U_{L2} - 1}{U_{L2} - 1} \right\}$$

$$I_2 = -\frac{2 I_{x_L}}{I_{x_L}} \left\{ \left[ \frac{e^{U_{L2} / 2} - 1}{U_{L2} - 1} \right] + \frac{e^{U_{L2} / 2} - 1}{U_{L2} - 1} \right\}$$

where $U = \delta_L / \sin(\gamma_L) + j k_0 x_L$. Note that the final expressions in (6)–(7) are obtained from (12)–(15) after noting that in the far field region $r_L \approx r, \theta_L = \theta$ and $\phi_L = \phi$.

ACKNOWLEDGMENT

The authors would like to thank M. Paquet for performing the measurements of the radiation patterns, M. Bruin for manufacturing the lens with great care, and F. Nennie for his assistance during the entire project.

REFERENCES


Simona Bruni (M’06) received the Laurea degree and the Ph.D. degree in telecommunications engineering from the University of Siena, Siena, Italy, in 2002 and 2006, respectively.

Since the beginning of 2007 she is working as an Antenna Engineer at Calearo, Vicenza, Italy. Her research interests are in the area of applied electromagnetics, focused on numerical and asymptotic methods, and design of broad-band antennas.

Dr. Bruni’s Ph.D. degree work was financed and hosted by the Defence, Security and Safety Institute of The Netherlands Organization for Applied Scientific Research (TNO), The Hague, The Netherlands.

Andrea Neto (M’00) received the Laurea degree (summa cum laude) in electronic engineering from the University of Florence, Italy, in 1994 and the Ph.D. degree in electromagnetics from the University of Siena, Italy, in 2000.

Part of his Ph.D. was developed at the European Space Agency Research and Technology Center, Noordwijk, The Netherlands, where he worked for over two years in the Antenna Section. From 2000 to 2001, he was a Postdoctoral Researcher at the California Institute of Technology, Pasadena, working for the Sub-millimeter Wave Advanced Technology Group (S.W.A.T.) Group of the Jet Propulsion Laboratory. Since 2002, has been a Senior Antenna Scientist at the Defence, Security and Safety Institute of the Netherlands Organization for Applied Scientific Research (TNO), The Hague, The Netherlands. His research interests are in the analysis and design of antennas, with emphasis on arrays, dielectric lens antennas, wideband antennas and EBG structures.

Filippo Marliani was born in Florence, Italy. He received the Laurea degree (cum laude) in electronic engineering from the University of Florence, Italy, in February 1998.

In 1998, he collaborated with the Earth Observation Section of the Institute of Applied Physics “Nello Carrara” (IFAC), National Research Council (CNR), in Florence. From November 1998 to October 1999, he was with the Electromagnetics Group, Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, MA. Since November 1999, he has been with the Electromagnetics and Space Environments Division of the European Space Agency, Noordwijk, The Netherlands, where in January 2007 he was appointed Head of the Electromagnetic Compatibility Section. His research interests concern theoretical and numerical electromagnetics for RFC/EMC applications and measurement procedures for unit and system level testing of space systems.