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On the Use of Fly's Eye Lenses with Leaky-Wave Feeds for Wideband Communications

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Abstract—In order to fulfill the exponential increase in the demand of high-speed wireless links, future XG wireless networks will be developed at higher carrier signal frequencies, reaching the hundreds of gigahertz. In this contribution, a leaky-wave-fed HDPE lens antenna working at G-band which can be useful for future XG communications is presented. The proposed lens design enables the generation of up to 40 beams, with gains higher than 30 dB. Analytical tools have been applied to optimize the lens aperture efficiency, validating the results via full-wave simulations. The reached aperture efficiency for the broadside beam is higher than 80% over a 44% relative bandwidth. The measurement results for a fabricated prototype show excellent agreement with the simulated performance.

Index Terms—Leaky-wave antennas, lens antennas, multibeam antennas, wideband communications.

I. INTRODUCTION

NOWADAYS, wireless communication systems are experimenting a revolution, driven by new data-demanding applications. Future applications will need wireless links with capacities orders of magnitude larger than currently achieved in back-haul, front-haul, and access networks. Exploiting small RF bandwidths with advanced signal processing techniques has reached a saturation point, and therefore, other system alternatives need to be explored. One of the most reasonable approaches is to increase the RF carrier frequency, where larger bandwidths can be exploited to boost the maximum achieved throughput [1]. However, the wave-propagation spreading factor proportionally increases to the square of the frequency, and therefore, moving to higher frequency bands requires the use of high-gain antennas (>30 dB) to fulfill the link budget. Furthermore, transmitted power is limited at high frequencies, as upconverting chains are required, which

exhibit low efficiencies. This fact magnifies the importance of achieving highly efficient front-ends and antennas.

In this article, we propose the use of wideband Fly's eye lens array architectures to generate multiple simultaneous fixed high-gain beams, in order to give coverage in future front-haul point-to-multipoint scenarios or access networks in dense small cells [Fig. 1(a)]. We consider multibeam systems with a single transceiver per beam or lens feeder. Possible application examples for the point-to-multipoint scenario are fixed wireless access connections, where a base station provides several buildings with internet broadband access, substituting fiber optics. Here, different beams in the base station point simultaneously at multiple static receivers situated in different buildings. An example for the second scenario, dense small cells, could be a football stadium, auditorium, or any crowded event. In this case, overlapping beams should simultaneously cover the whole field of view where the users are distributed [2].

The use of lens antennas allows to efficiently achieve larger gains >30 dBs, over larger bandwidths (>20%) in comparison with other more resonant antenna solutions, such as Fabry–Perot [3], metasurfaces [4], or transmitarrays [5]. Moreover, the lens-steering capability enables the generation of simultaneous multiple beams pointing in different directions with planar Fly's eye lens array panels, as displayed in Fig. 1(b). This facilitates the integration with high-frequency front-ends, which can be in this way mounted on a single printed circuit board (PCB) per panel. A high-gain beam with stable pointing angle over the whole bandwidth is achieved contrary to other frequency dispersive leaky-wave concepts (see [6], [7]). In order to arrive to a more compact antenna array, each lens can host several transceivers or beams. The number of beams per lens will be mainly limited by the size of the active circuitry in terms of wavelength. Maximizing the lens aperture efficiency allows to further reduce the lens array size.

Table I presents a summary of the performances for the most relevant designs of integrated lens antennas found in the literature. Materials with high dielectric permittivity, ϵ_r , such as silicon ($\epsilon_r = 11.9$) or alumina ($\epsilon_r = 9$) are chosen in many cases, presenting very low ohmic losses. The main disadvantage of these designs lays on the high reflection in the lens–air interface. There are several designs reported reaching high lens aperture efficiencies over a relative bandwidth of 15%–20% [8], [9]. Double-slot antennas [8] or waveguide

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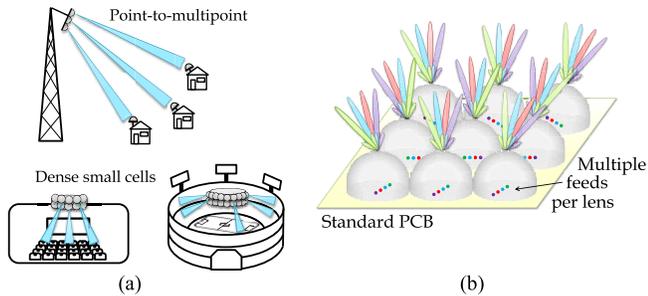


Fig. 1. (a) Scenarios for fixed multibeam antenna architectures. (b) Planar lens array panel generating multiple fixed simultaneous beams.

apertures [9] are usually used as lens feeders. Larger bandwidth designs are presented using double bow-tie antennas, in [10], and nonresonant leaky-wave antennas, in [11]. However, in both cases, the feed pattern which illuminates the lens surface does not allow for reaching high aperture efficiencies and higher sidelobe levels are reported. In [12] and [13], a resonant leaky-wave antenna (LWA) or Fabry–Perot antenna was proposed as a promising solution to act as a silicon lens feeder, due to its high directivity, symmetric pattern, compactness, low profile, and compatibility with silicon fabrication processes but presented narrow bandwidth. Designs using lenses in quartz ($\epsilon_r = 3.8$) and MACOR™ material ($\epsilon_r = 5.6$) have been reported in [14]–[16]. In [14], the beam quality is degraded due to the effects of multiple reflections. In [15], the feed pattern asymmetry results in lower lens aperture efficiency, presenting high reflection losses. In [16], the use of a shell with lower ϵ_r improves the reflection efficiency and optimizes the lens-steering capabilities. Unfortunately, in both cases, the dielectric loss is very high, which leads to a considerable power loss (3.7–5.6 dB reported in [16]). Low ϵ_r plastic materials ($\epsilon_r \leq 2.5$) represent a good candidate to fabricate lens antenna arrays, due to the availability of moderate loss materials (HDPE, Rexolite, Topas), with light weight and enabling cost-effective manufacturing through processes such as injection molding or 3-D printing. Nevertheless, the efficient illumination of these lenses is more challenging. The larger lens eccentricity, e , needed for low ϵ_r lenses implies that directive feeds are needed to avoid the lens critical angle. Most of the reported designs for lenses with $\epsilon_r \leq 2.5$ present reflection issues which decrease the antenna efficiency and degrade the radiation pattern quality [17]–[19].

In this article, we extend the LWA feed architecture presented in [13] for low ϵ_r lens materials. The goal is to achieve high gain with high aperture efficiency over a wide band, required for future wireless communication applications. A new methodology for the analysis and design of elliptical lenses illuminated by resonant LWAs is introduced, based on an analysis of the lens antenna in reception. This approach allows maximizing the aperture efficiency of the lens antenna by performing a field match between the geometrical optics frequency-independent field and the LWA field, avoiding more time-consuming parametric optimizations based on full wave (FW) or physical optics (PO). Design curves for the maximum aperture efficiency and bandwidth as a function of the lens ϵ_r

TABLE I
SIMULATED PERFORMANCES OF INTEGRATED LENS ANTENNAS*

Ref.	f_0 (GHz)	BW (%)	Dir. f_0 (dB)	Min η_{ap} (dB)	Max L_{diel} (dB)	Max. SLL (dB)
[8]	246	-	29.9	-1.1	-	-16
[9]	60	16	19.9	-2.3	-	-16
[10]	400	1:3	27	-2.6	0.2	-8
[11]	50	1:5	19	-3.3	1.5	-12.5
[13]	550	18	22.2	<i>-1</i>	-	-15
[14]	77	10	24	<i>-1.5</i>	0.5	-15
[17]	30	12.5	25.9	<i>-1.9</i>	0.5	-14
[18]	73.5	7	37.3	-2.9	1	-
[19]	290	30	31.5	-4	-	-8
This work	180	44	34	-1	0.5	-15

(*) Values in italic have been extracted from the articles.

are in this way derived. The steering properties of elliptical lenses are also maximized here, for the first time, by introducing an optimal truncation in the lens geometry that trade-offs reflection and taper efficiencies. Most of the previous works in Table I considered nontruncated lenses. We present the maximum number of overlapping beams that may be achieved per planar lens array panel [Fig. 1(b)] as a function of different lens materials. Finally, we report an experimental validation of the performances for a low-density elliptical lens made with HDPE, leading to state-of-the-art performances. The simulation results show an aperture efficiency higher than 80%, matching better than -10 dB and ohmic losses lower than 0.5 dB over 44% relative bandwidth. The measurements are in very good agreement with the simulations.

This article is organized as follows. Section II describes the resonant LWA feeder concept. Section III describes the lens aperture efficiency optimization methodology. Section IV describes bandwidth properties of resonant LWAs in terms of aperture efficiency, as a function of the lens ϵ_r . Section V shows the scanning performances in truncated elliptical lenses with different ϵ_r . Sections VI and VII show the design and experimental results of a prototype fabricated in low ϵ_r material.

II. LENS FEEDER: RESONANT LEAKY-WAVE ANTENNA

The proposed feeder geometry is shown in Fig. 2. It consists of a resonant (Fabry–Perot) air cavity between a ground plane and a lens dielectric medium. The excitation of the air cavity is performed by means of a squared waveguide TE₀₁ mode illuminating a double-slot opening in the ground plane [Fig. 2(b) and (c)], similar approach to the one used in [20] and [21]. The double slot is used to further shape the pattern, as will be explained in following subsections, in order to maximize the aperture efficiency. It also plays an important role in the impedance matching to the waveguide. The squared waveguide size is fixed to $wg = 0.8\lambda_0$ at the central frequency [Fig. 2(b)], to avoid the excitation of higher order modes.

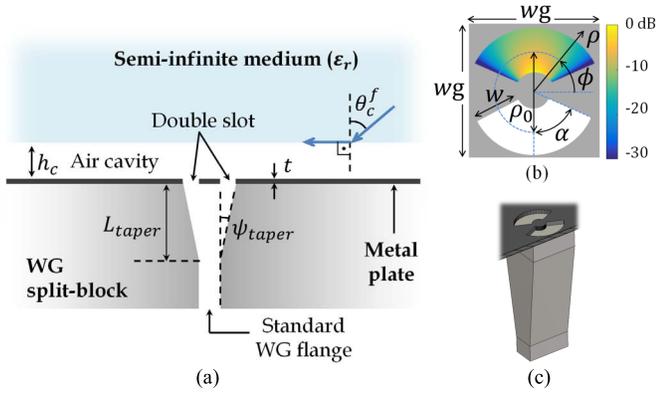


Fig. 2. (a) LWA geometry. A semi-infinite dielectric slab is located at a distance $h_c \sim \lambda_0/2$ to create a Fabry–Perot cavity illuminated by a tapered WG ($L_{taper} = 4.4\lambda_0$, $\psi_{taper} = 3^\circ$). θ_c^f is the feed critical angle, after which the cavity presents total reflection. Blue arrows: incident and transmitted rays in the dielectric–air interface. (b) Main parameters of the double-slot iris and approximated magnetic current distribution from (8). (c) 3-D view of the tapered waveguide and double slot in the ground plane.

The waveguide is then tapered to a standard flange as shown in Fig. 2(a) and (c). This tapering has minor impact on the radiation and impedance matching.

The physical phenomenon exploited in standard resonant LWAs or Fabry–Perot antennas is the excitation of a pair of nearly degenerated TM_1/TE_1 leaky-wave modes inside a resonant dielectric cavity [22], such as air, placed between a ground plane and a quarter wavelength dielectric substrate with higher ϵ_r [Fig. 2(a)]. These modes propagate radially by means of multiple reflections along the cavity, leaking at the same time energy into the infinite dielectric medium. This effect increases the antenna effective area and thus, its directivity but at the cost of bandwidth [23]. In [12], it was shown that an enhancement in the *bandwidth* \times *directivity* performance of a resonant LWA can be achieved by using a resonant air cavity below a semi-infinite dense medium (the lens), as displayed in Fig. 2(a). In this configuration, the LW modes present propagation phase constants $\beta_{LW} = k_0\sqrt{\epsilon_r} \sin\theta_{LW} < k_0$, which implies that LW modes radiate always in elevation angles smaller than the feed critical angle [θ_c^f in Fig. 2(a)], and very low power is radiated at $\theta > \sin^{-1}(1/\sqrt{\epsilon_r})$. Furthermore, in order to achieve a broadband design, it is convenient to choose a medium with low ϵ_r to reduce the frequency dispersion of the TM_1/TE_1 leaky-wave modes [24].

III. LENS APERTURE EFFICIENCY OPTIMIZATION

The proposed lens geometry is shown in Fig. 3(a). The lens is an elliptical surface characterized by an eccentricity $e = 1/\sqrt{\epsilon_r}$ and truncated at an angle θ_{edge} defined from the bottom lens focus. The lens truncation helps to achieve high aperture efficiency when compared to more classical full hemispherical lens designs [8]. The LW ground plane is placed at a distance Δz above the lens focal plane, since the phase center of LWAs is below the ground plane [25].

In order to evaluate the lens antenna aperture efficiency, η_{ap} , we propose an analysis in reception as detailed in the Appendix. The aperture efficiency can in this way be expressed

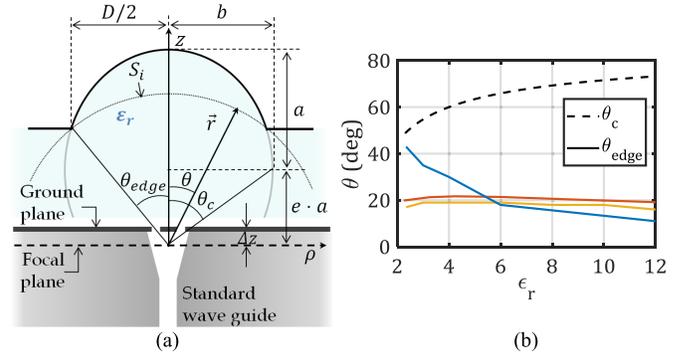


Fig. 3. (a) Elliptical lens parameters are related as $a^2 = b^2 + (e \cdot a)^2$, and the focal distance is defined as $F = a + e \cdot a$. (b) Dashed line: lens critical angle, θ_c , as shown in (a). Solid line: comparison of optimum lens edge truncation angles, θ_{edge} , as defined in (a) where the blue line is θ_{edge} for maximum η_{il} with the LWA as feeder, the red line is θ_{edge} for maximum steering angle applying the geometric criteria, and the yellow line is θ_{edge} for maximum steering angle with a cosine primary pattern extracted with PO.

as a field reaction between a frequency-independent analytical field and the field radiated by the feeder in an infinite dielectric medium. This approach in reception allows us to maximize the lens aperture efficiency by optimizing the lens feed geometry, its phase center position, Δz , and the lens θ_{edge} , without performing the secondary pattern computation, which would be more time-consuming, as in more standard PO approaches [8].

The aperture efficiency for this focusing system can be evaluated dividing the power P_L received by a matched load by the power captured by the lens aperture from the incident plane wave, $P_{in} = 1/(2\zeta_0)|E_0^{PW}|^2 A_{lens}$, being $A_{lens} = \pi D^2/4$ the projected area for a lens with diameter D [Fig. 3(a)] and E_0^{PW} the plane wave amplitude, as follows:

$$\eta_{ap} = \frac{P_L}{P_{in}} = \eta_{il}\eta_{ref}. \quad (1)$$

η_{ap} accounts for the illumination efficiency and the reflection efficiency, and is equivalent to the ratio between the lens directivity and that of a uniform circular aperture. The efficiency η_{il} represents the antenna efficiency assuming the use of a perfect matching layer in the whole frequency band. It accounts for taper and spillover efficiencies (power reaching the focal plane but not intercepted by the feed). η_{ref} is the power reflected in the air–lens interface.

The power received by the load, P_L , as shown in the Appendix, can be expressed as a reaction integral between two terms: 1) the inward incident field inside the lens, $\vec{E}_{GO}^{S_i}$, caused by the broadside plane wave impinging on the lens, which is frequency independent, and 2) the field radiated in an infinite dielectric medium by the feed antenna considered in transmission, \vec{E}_a^{GP} , as

$$P_L = \frac{\int_0^{2\pi} \int_0^{\theta_{edge}} \vec{E}_{GO}^{S_i}(\theta, \phi) \cdot \vec{E}_a^{GP}(\theta, \phi) e^{jk\Delta z \cos\theta} R^2 \sin\theta d\theta d\phi}{4\zeta_d^2 P_{rad}} \quad (2)$$

where P_{rad} is the total power radiated by the feed antenna in the infinite dielectric medium and ζ_d the wave impedance in

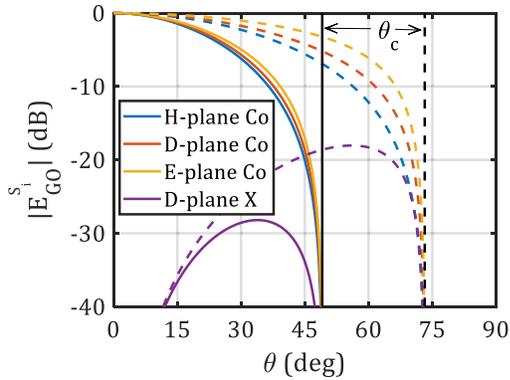


Fig. 4. Normalized $|E_{GO}^{S_i}|$ versus lens elevation angle θ over S_i , until lens critical angle. Solid line: $\epsilon_r = 2.3$, dashed: $\epsilon_r = 12$. The lens critical angles, θ_c , are marked with black lines. Results shown with no matching layer.

the lens medium. The field reaction in (2) is evaluated over the truncated spherical surface S_i [Fig. 3(a)] with radius R , centered at the lens focus and limited by the lens truncation angle, θ_{edge} . The term $e^{jk\Delta z \cos\theta}$ assumes that the lens is in the far-field of the feed antenna. $\vec{E}_{GO}^{S_i}$ can be approximated in S_i as a superposition of local plane waves, whose amplitude can be analytically derived applying geometrical optics (GO) [26].

The aperture efficiency is maximized when the inward incident field, $\vec{E}_{GO}^{S_i}$, is a conjugate match (in terms of amplitude, phase, and polarization) with the field radiated by the feed antenna, $\vec{E}_a^{GP}(2)$. Moreover, in order to achieve high reflection efficiency, η_{ref} , only the lens region above the lens critical angle, θ_c , should be illuminated [top half of the ellipse, see Fig. 3(a)]. θ_c exclusively depends on the dielectric ϵ_r and increases for higher values of ϵ_r as $\theta_c = \tan^{-1} \sqrt{\epsilon_r - 1}$ [Fig. 3(b)]. Thus, lenses with lower ϵ_r need to be illuminated with more directive beams than denser lenses.

Fig. 4 shows the amplitude of the incident field, $\vec{E}_{GO}^{S_i}$, for nontruncated elliptical lenses in materials with $\epsilon_r = 2.3$ and $\epsilon_r = 12$, plotted until θ_c . The phase of this field is constant over the chosen sphere S_i , where the integral in (2) is evaluated. The amplitude taper is mainly caused by the Fresnel transmission coefficients in the air–lens interface [26], which decay when approaching θ_c . Here, lenses without matching layer have been considered. If a matching layer was introduced, the corresponding transmission coefficients would result in a more symmetric incident field, $\vec{E}_{GO}^{S_i}$, over ϕ . A smaller contribution to this taper is originated by the power spreading factor in the refracted rays inside the lens. The lens eccentricity, e , increases for lower ϵ_r , resulting in a $\vec{E}_{GO}^{S_i}$ which approaches the cosine pattern. Instead, for high ϵ_r , where the lens approaches an spherical form, $\vec{E}_{GO}^{S_i}$ is closer to a squared pattern. In any case, both fields present a very low taper at the edges, differing very much to those corresponding to a parabolic reflector, where $\vec{E}_{GO}^{S_i}$ presents a secant shape [27].

IV. BANDWIDTH PROPERTIES OF THE RESONANT LEAKY-WAVE LENS ANTENNA VERSUS ϵ_r

The analysis in reception described in Section III has been applied to quantify the LWA bandwidth in terms of pattern

stability, or equivalently illumination efficiency, η_{il} , when acting as an integrated elliptical lens feeder. The LWA radiation patterns have been calculated with an asymptotic evaluation of the spectral Green's function (SGF) in stratified media [23], convoluted with a simplified current approximation for the double slot (Appendix). This approach allows us to optimize the feed-lens system aperture efficiency exclusively applying analytical tools. The first step to extract the bandwidth properties for different ϵ_r is to calculate the optimum lens, LW cavity, and double-slot parameters which give the maximum η_{il} . Next, the relative bandwidth for these optimum cases is determined in terms of the lens η_{il} over frequency.

When evaluating the resonant LWA as lens feeder, the main effect on the bandwidth is associated with the dispersion of the main leaky-wave modes [24]. However, the maximum η_{il} highly depends on the spurious TM_0 contribution [28] (being θ_{TM_0} its radiation angle), which can radiate inside or outside the truncated lens rim, θ_{edge} [Fig. 3(a)]. The distance ρ_0 in the double slot [Fig. 2(b)] becomes a key parameter to optimize this maximum efficiency, as it is directly related to the position of a null in the Fourier transform (FT) of the double-slot equivalent currents, which can be used to further shape the pattern. The double slot, circular to achieve rotationally symmetric patterns, has been considered for this analysis with an angle α of 60° and width w of $0.15\lambda_0$, being those realistic values to achieve a good matching to the waveguide.

In order to clarify the impact of the TM_0 radiation and double-slot distance in η_{il} , we analyze more deeply two extreme cases: $\epsilon_r = 2.3$ and $\epsilon_r = 12$. This ϵ_r -range is representative for most common lens materials found in the literature. Fig. 5 shows the far-field patterns of the resonant LWA, \vec{E}_a , for these two cases with maximum η_{il} , with an elementary excitation (left) and with the double-slot iris (right). For $\epsilon_r = 12$, the feed critical angle [Fig. 2(a)] arrives at 16.7° , and so the radiation is confined in smaller angles. The lens should be, therefore, intuitively truncated at $\theta_{edge} \leq 17^\circ$, in order to obtain high η_{tap} . $\vec{E}_{GO}^{S_i}$ presents an almost constant amplitude distribution at S_i for $\theta < 17^\circ$ [Figs. 4 and 5 (top)], and, therefore, a squared primary pattern is required in this case to reach a high η_{tap} . A cavity larger than the standard half wavelength, $h_c = 0.57\lambda_0$, contributes to the desired pattern form, as in this case the TM_1/TE_1 modes radiate toward larger angles [Fig. 5 (top left)]. The TM_0 mode radiation would reduce in this case the η_{tap} , and so the best compromise is to keep it out of the lens rim ($\theta_{edge} < \theta_{TM_0}$), as in [12] and [21]. In this way, the double slot should ideally fully cancel the effect of the TM_0 , in order to reduce the spillover, by introducing a null in the FT of its equivalent currents at $\theta_{iris} \sim \theta_{TM_0}$, as shown in Fig. 5 (top right). The optimum double-slot ρ_0 is in this case around $0.64\lambda_0$. This is the best compromise between canceling completely the TM_0 (higher η_{so}) and keeping the pattern inside the lens rim as squared as possible (higher η_{tap}). The resulting optimum pattern for $\epsilon_r = 12$ is shown in Fig. 5 (top right), where θ_{iris} can be clearly distinguished in the E-plane at $\theta = 15^\circ$, lowering the TM_0 radiation peak in around 15 dB. The position of θ_{iris} has in case of high ϵ_r materials similar dispersion characteristic as

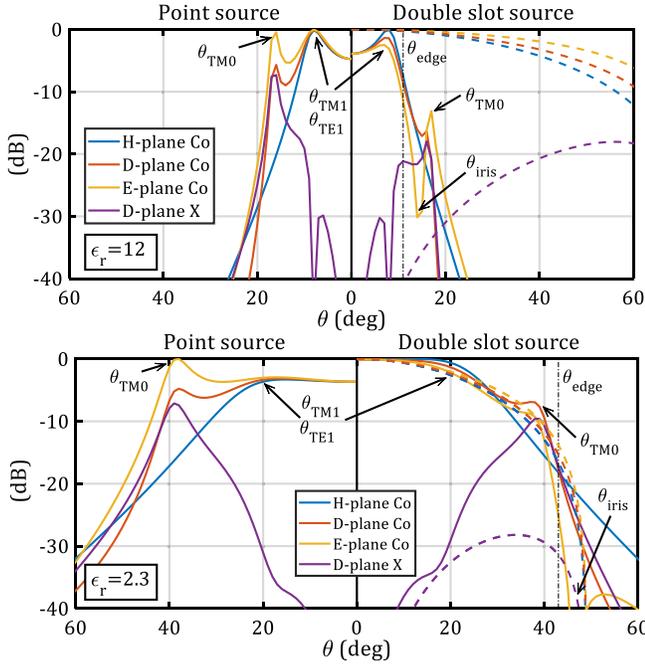


Fig. 5. LWA far-field patterns, \vec{E}_a , for $\epsilon_r = 12$ (top) and $\epsilon_r = 2.3$ (bottom). Left side: x -polarized elementary magnetic source. Right side: double-slot iris source. The shown patterns are symmetric with respect to θ . For reference, $\vec{E}_{GO}^{S_i}$ at S_i (dashed lines) is also plotted. All curves have been extracted using the geometric parameters (cavity, double slot, and lens) which maximize η_{il} . Those are $h_c = 0.56\lambda_0$, $\rho_0 = 0.5\lambda_0$, and $\theta_{edge} = 43^\circ$ for $\epsilon_r = 2.3$, and $h_c = 0.57\lambda_0$, $\rho_0 = 0.64\lambda_0$ and $\theta_{edge} = 11^\circ$ for $\epsilon_r = 12$.

the LWA TM_1/TE_1 and modes and, therefore, does not impact significantly in the relative bandwidth of the lens η_{il} .

For $\epsilon_r = 2.3$, the feed critical angle arrives at 41° . The lens should be, therefore, truncated at $\theta_{edge} \leq 41^\circ$, in order to obtain a high η_{tap} . For this angular range, $\vec{E}_{GO}^{S_i}$ presents an amplitude distribution at S_i which approaches a cosine with approximately -15 dB taper [Figs. 4 and 5 (bottom)]. In this case, it is more convenient to choose $\theta_{edge} > \theta_{TM0}$ and only taper down the TM_0 contribution by means of the double slot. The optimum ρ_0 is, therefore, $0.5\lambda_0$, for which the null in the FT of the double-slot equivalent currents $\theta_{iris} > \theta_{TM0}$ [Fig. 5 (bottom right)]. The resulting LWA pattern [Fig. 5 (bottom right)] presents in this way a very similar taper to the incident field, reaching high η_{tap} . Here, θ_{iris} appears at around 45° , and the TM_0 radiation is tapered down to -8 dB, being the high correlation between \vec{E}_a and $\vec{E}_{GO}^{S_i}$ evident. For low ϵ_r , the position for θ_{iris} , related to the optimum ρ_0 , appears in the region beyond the feed critical angle, where very low power is radiated by the LWA. In this way, the narrow-band characteristic of the slot θ_{iris} has a small impact on the primary patterns. Their dispersion with frequency is consequently mostly related to the LW modes dispersion, which are very broadband for low ϵ_r . This fact provides these lenses with a very large η_{il} bandwidth.

Fig. 6 shows the maximum reachable lens η_{il} as a function of the double-slot distance, ρ_0 for several lens ϵ_r . We can observe that the overall maximum η_{il} appears for $\epsilon_r = 2.3$ thanks to the very good field correlation explained previously,

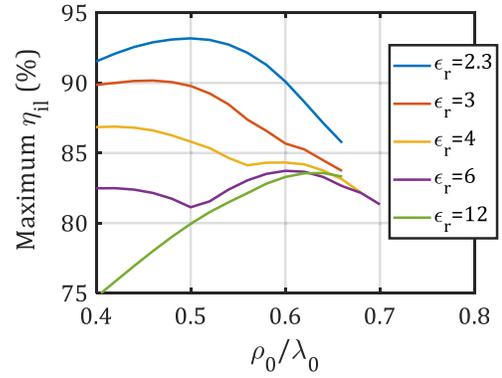


Fig. 6. Maximum η_{il} versus double-slot distance ρ_0 normalized to λ_0 . The rest of geometric parameters are optimized for every double-slot distance.

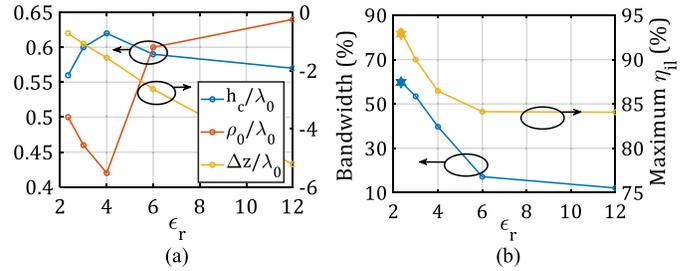


Fig. 7. (a) Optimum cavity height, double-slot distance, and LWA phase center position for maximum η_{il} versus lens ϵ_r . (b) LWA relative bandwidth for lens $\eta_{il} > 80\%$ as a function of ϵ_r , for a design with the maximum η_{il} , and maximum η_{il} achieved versus ϵ_r . The stars represent the simulated values reached with the HDPE prototype.

achieved by keeping the TM_0 inside the lens rim and a double-slot distance of $\rho_0 \sim 0.5\lambda_0$. As we move to higher ϵ_r , a second relative maximum appears for $\rho_0 \sim 0.6\lambda_0$, becoming an absolute maximum for $\epsilon_r = 12$, associated with the TM_0 radiating outside the lens rim. As the ϵ_r increases from 2.3 to 4, the taper in $\vec{E}_{GO}^{S_i}$ decreases, demanding a lower value of amplitude taper to keep a high η_{tap} . The inflection point between these two design strategies appears at around $\epsilon_r = 6$, where a good compromise between η_{tap} and η_{so} can be reached for both $\theta_{edge} < \theta_{TM0}$ (optimum $\rho_0 = 0.6$) and $\theta_{edge} > \theta_{TM0}$ (optimum $\rho_0 = 0.4$). If a wideband design is pursued, the design with $\theta_{edge} > \theta_{TM0}$ and $\rho_0 = 0.4$ should be chosen, for which the frequency dispersion of θ_{iris} has a lower impact on the η_{il} bandwidth.

Following this design approach, the LWA lens antenna has been optimized as a function of ϵ_r . The obtained parameters are summarized in Figs. 3(b) and 7(a). The relative frequency bandwidth for which $\eta_{il} > 80\%$ is achieved, as well as the maximum reached η_{il} are reported in Fig. 7(b). Relative bandwidths higher than 40% can be achieved with $\epsilon_r < 4$, whereas 15% bandwidth is reached with this criterion in case of using a lens with $\epsilon_r \sim 12$.

V. ELLIPTICAL LENS SCANNING PROPERTIES VERSUS ϵ_r

The lens-steering capability defines the maximum number of beams, overlapping at -3 dB, which can be generated with a single Fly's eye lens planar panel [Fig. 1(b)]. In [29],

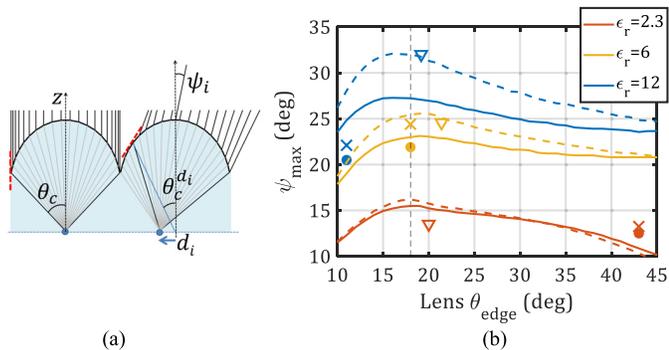


Fig. 8. (a) Elliptical lens ray-tracing example for on- and off-focus feeds, where the lens critical angles are shown for both cases: θ_c and $\theta_c^{d_i}$, respectively. (b) Maximum steering angle ψ_{max} achieved in free space versus lens θ_{edge} . Triangles: geometric approximation. Solid line: PO analysis with cosine illumination for H-plane steering; dashed line: the same for E-plane. Circles: PO analysis with LW feed optimized for broadside for H-plane steering; crosses: the same for E-plane.

the steering properties of extended hemispherical lenses were studied for double-slot feeders. Here, instead, we investigate the steering capabilities of truncated elliptical lenses as function of the lens truncation angle, θ_{edge} [Fig. 3(a)], using ideal feeders adapted to each θ_{edge} . The scope is to derive the optimum θ_{edge} that maximizes the steering angle (accepting 3 dB scan loss) and aperture efficiencies by making a tradeoff between the different loss mechanisms involved (taper, reflection, and spillover). The considered lens materials are the same as in Section IV. Finally, the steering capabilities for lenses with different ϵ_r fed by the LWA are evaluated, completing the performance analysis for such feeders.

Displacing the feed a distance d_i along the lens focal plane generates a beam-steering angle ψ_i in the lens far-field [Fig. 8(a)]. ψ_i can be geometrically approximated by tracing the central ray as $\psi_i = \sin^{-1}(\sqrt{\epsilon_r}(d_i/F))$, being F the lens focal distance [Fig. 3(a)]. It is clear that lower ϵ_r materials need higher feed displacements to achieve a certain steering angle. In order to avoid significant spillover loss, the maximum feed displacement should be $d_{max} = D/4$. On one hand, for a certain d_i , larger θ_{edge} (smaller F/D) leads to higher steering angles, ψ_i . On the other hand, θ_c appears at a smaller angle when the lens is illuminated off-focus [$\theta_c^{d_i}$ in Fig. 8(a)], and so the lens should be truncated at θ_{edge} smaller than $\theta_c^{d_i}$ (larger F/D) to avoid significant reflection loss. In this way, the optimum θ_{edge} which maximizes the steering angle should fulfill $\theta_{edge} = \theta_c^{d_{max}}$, for $d_{max} = D/4$. This criterion is independent of the lens D . Fig. 3(b) shows the optimum θ_{edge} for different ϵ_r , extracted with the described approach. For lenses with ϵ_r between 2 and 12, the optimum θ_{edge} stays around 20° in all cases. Fig. 8 shows the corresponding ψ_{max} calculated geometrically (triangles).

The calculated optimum θ_{edge} has been validated with a single-plane Physical Optics tool, extracting ψ_{max} for which η_{ap} decays 3 dB with respect to broadside. Lenses with $D = 18\lambda_0$ are considered, illuminated by cosine patterns with -11 dB taper at the lens edge. The results are shown in Fig. 8(b), being the optimum θ_{edge} near 18° for all studied cases, validating the geometric approximation [Fig. 3(b)].

The maximum steering angles extracted here for $\epsilon_r = 12$ are 27° and 30° for the H- and E-planes, respectively, for truncated lenses with $\theta_{edge} = 18^\circ$ and adapted ideal cosine feeds. These steering angles are higher than in the case of the silicon lenses considered in [29], where a maximum scan angle of 20° was reported for $\theta_{edge} \sim 60^\circ$ [off-scale in Fig. 8(b)].

When using the LWA as a feeder, in case of $\epsilon_r = 6$, the optimum lens θ_{edge} values which maximize η_{il} at broadside and ψ_{max} for steering are nearly coincident [see Fig. 3(b)]. In case of $\epsilon_r < 6$, the optimum θ_{edge} for the lens broadside beam is larger than the optimum θ_{edge} for maximum steering. The maximum number of beams is limited, in this case, by the increment in the reflection loss due to the larger total-reflection area. In case of $\epsilon_r > 6$, the optimum θ_{edge} for the broadside beam is smaller than the optimum θ_{edge} for steering. Here, the maximum steering angle reached is limited by the increasing spillover loss, due to the larger F/D ratio.

The achieved steering angles with the LWA patterns optimized for broadside [Fig. 5 (right)] have been calculated using an *ad hoc* PO tool considering the whole lens surface [30]. The results are plotted in Fig. 8(b) for $\epsilon_r = 2.3, 6$, and 12 (circles). The PO simulated results for the LWA have been validated via FW simulations (no multiple reflections) carried out with the software EMPIRE-XPU [31]. The results are in good agreement, as shown in Fig. 9.

Coming back to the scope of this analysis, the number of beams achieved per plane considering lenses with LW feeders and $D = 18\lambda_0$ can be evaluated from the maximum steering angle and the -3 dB beamwidth ($\sim 3.5^\circ$) as follows: 7 beams ($\psi_{max} = 12.5^\circ$) with $\epsilon_r = 2.3$, 12 beams ($\psi_{max} = 21.5^\circ$) with $\epsilon_r = 6$, and 11 beams ($\psi_{max} = 19.75^\circ$) with $\epsilon_r = 12$. In a two-dimensional circular planar array of lenses, those steering angles correspond to 40, 110, and 90 beams for $\epsilon_r = 2.3, 6$, and 12 , respectively. In the multifixed beam scenarios introduced in Section I, each beam corresponds to one feed antenna and transceiver.

VI. BROADBAND PROTOTYPE

From the analyses in previous sections, the optimal lens material for this application seems to be $\epsilon_r = 2.3$, since it presents the highest efficiency and widest bandwidth. In addition, due to the lens low ϵ_r , there is no need to use a matching layer, which added to the lower material cost results in the most cost-effective approach. The only compromise made when choosing a lens with low ϵ_r is the reduction in the potential field of view achievable with a Fly's eye array [Fig. 1(b)], as discussed in Section V. In this section, a design covering the whole G-band (140–220 GHz) is presented, using an HDPE lens ($\epsilon_r = 2.3$, $\tan\delta = 0.00033$). This material has been chosen after characterizing several plastics in the frequency range of interest, as it presented the lowest measured dielectric loss. As the final step, a fine tuning of the double-slot geometry is carried out in order to achieve a good impedance matching to the waveguide over the whole frequency band. The analytic primary patterns calculation and lens aperture efficiency computation are validated for the final optimized model by means of FW simulations.

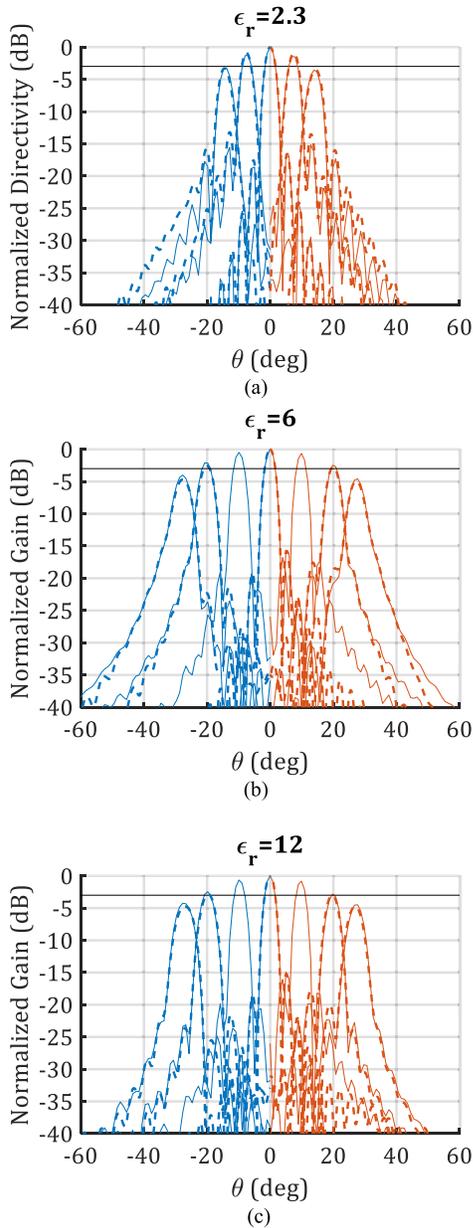


Fig. 9. Lens far-field magnitude simulated with a 2-D PO (solid line) and FW without multiple reflections (dashed line), normalized to the broadside beam. The lenses ($D = 18\lambda_0$) and LWA feeds are optimized for maximum η_{il} at broadside (Section IV). (a) $\epsilon_r = 2.3$ for $d = 0, 0.08, \text{ and } 0.16 D$. (b) $\epsilon_r = 6$ for $d = 0, 0.11, 0.23, \text{ and } 0.30 D$. (c) $\epsilon_r = 12$ for $d = 0, 0.13, 0.25, \text{ and } 0.34 D$. Blue: E-plane. Red: H-plane. A black line marks the 3 dB scan loss criterion for maximum steering angle.

A. Feed radiation patterns and impedance matching

The final geometrical values of the optimized double-slot iris are set to $a = 65.5^\circ$, $t = 0.08\lambda_0$, $\rho_0 = 0.54\lambda_0$, $w = 0.3\lambda_0$, and the cavity $h = 0.56\lambda_0$ (Fig. 2), being λ_0 to the wavelength at $f_0 = 180$ GHz. The slots have been widened to reach a good impedance matching; ρ_0 has been slightly increased with respect to the optimum. The impact of those small variations in the η_{il} is negligible, as will be shown in the next subsection. Fig. 10 shows the radiation pattern magnitude and phase at f_0 , while Fig. 11(a) shows

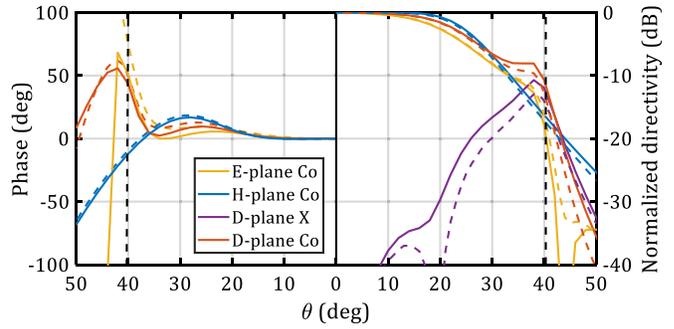


Fig. 10. Feed radiation patterns and phase for the fabricated prototype at f_0 . Solid lines: SGF solution, dashed lines: FW simulations. The prototype lens θ_{edge} is marked with dashed black lines.

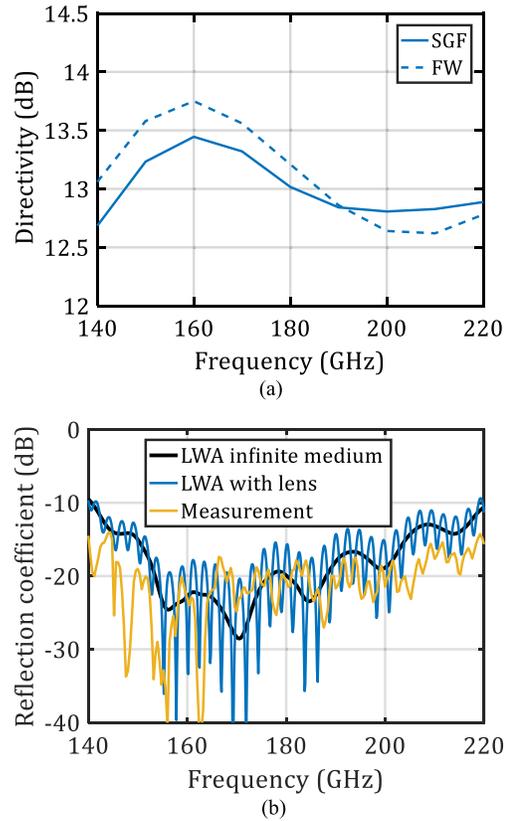


Fig. 11. (a) Maximum directivity over frequency for the prototype LWA radiating in the semi-infinite medium and (b) LWA reflection coefficient.

the maximum directivity over frequency. In both cases, there is a good agreement between analytically computed results and FW simulations, validating the approximate expression of the iris current distribution introduced in (8) (Appendix). The maximum variation in the lens LWA feed directivity over the entire frequency band is around 1 dB. Fig. 11(b) shows the FW simulation results for the LWA reflection coefficient radiating in a semi-infinite medium and inside the lens (including multiple reflections), in both cases lower than -10 dB over the whole operational bandwidth. The FW simulation model is done with perfect electric conductor boundary conditions and includes a taper from a standard flange to a squared waveguide [Fig. 2(a) and (c)].

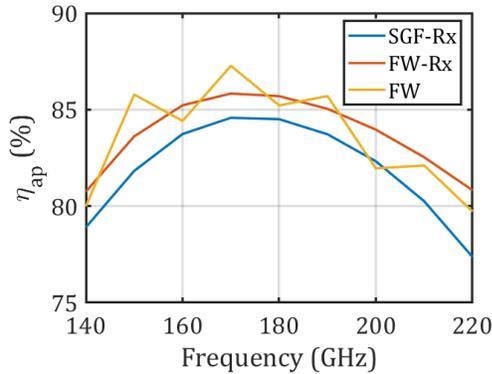


Fig. 12. Lens η_{ap} over frequency obtained with the reception analysis using the SGF primary pattern with the double-slot current approximation (SGF-Rx) and the FW simulated primary pattern (FW-Rx), together with an FW simulation with multiple reflections, which introduce a small ripple.

B. Lens Radiation Performance

In this section, the simulation results for a *G*-band HDPE lens with $D = 3$ cm are shown. After tuning the double-slot geometry in order to achieve a good matching over the whole bandwidth, a fine readjustment of the feed Δz and lens θ_{edge} has been performed to maximize the lens η_{ap} , applying the reception analysis described in Section III. The final value for θ_{edge} is 40.3° , slightly smaller than the optimum displayed in Section IV due to the higher taper in the E-plane, caused by the larger ρ_0 . The final Δz has been set to -0.95 mm.

The resulting η_{ap} is displayed in Fig. 12. Here, the analysis in reception (Rx) has been applied on one side to the primary patterns calculated with the SGF convoluted with the current approximation for the double slot (SGF-Rx) and on the other side to the FW primary patterns (FW-Rx), validating (8) over the whole frequency band. FW simulations with the lens, including multiple reflections, have been performed in order to validate the lens η_{ap} optimization and analysis in reception (Fig. 12) and to obtain the lens far-field radiation patterns (Fig. 13). The good agreement between the η_{ap} obtained through the analysis in reception and FW simulation proves that the multiple reflections and possible spillover do not have any significant impact on the lens antenna far-fields, due to the lens low ϵ_r . The FW simulation results show η_{ap} higher than 80% in the whole frequency band. The achieved bandwidth is also marked in Fig. 7, reaching almost identical values with respect to the ideal case.

In Fig. 13, the normalized simulated co-polarized patterns for the broadside and two steered beams are shown. The highest lens cross-polar level at broadside normalized to the copolar maximum directivity is -20 dB at 140 GHz. The co-polar gain over frequency is displayed in Fig. 14, corresponding the lens η_{ap} drop in the steered beams to the expected from the analysis in Section V. The maximum estimated ohmic loss in the lens dielectric material is about 0.5 dB at the highest frequency (220 GHz). The ohmic losses in the 1 cm-long waveguide are between 0.8 and 1.5 dB in the whole frequency band, for an estimated effective metal conductivity of $3.6e5$ S/m. The conductor loss could be

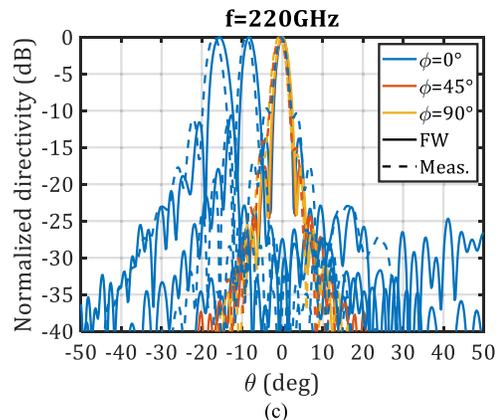
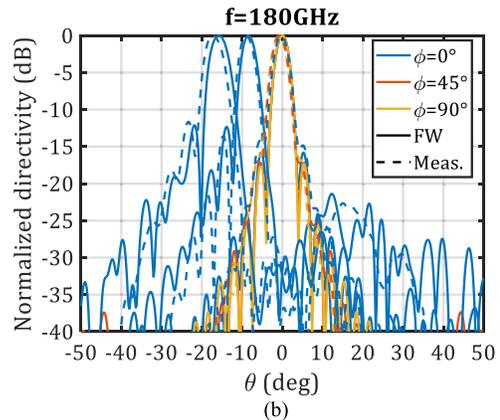
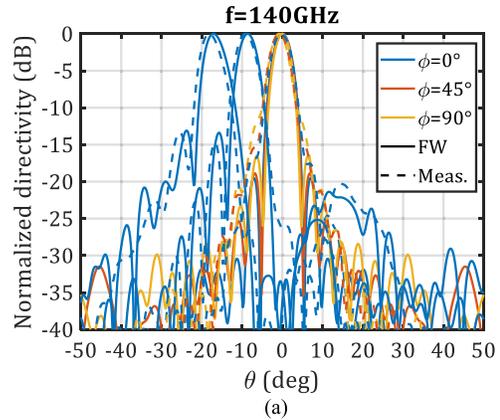


Fig. 13. FW simulated (multiple reflections) and measured lens patterns normalized to each beam maximum, at (a) 140 GHz, (b) 180 GHz, and (c) 220 GHz. The steered beams correspond to feed displacements, $d = 2.8$ and 5.6 mm, for which the measured steering angle is 9.25° and 18.25° for 140 GHz, 9° and 17.25° for 180 GHz, and 9° and 17° for 220 GHz, respectively.

significantly reduced in an integrated design by shortening the waveguide and improving its roughness. Table II summarizes the estimated loss contributions for a single lens.

VII. FABRICATION AND MEASUREMENTS

Two identical prototypes have been fabricated consisting of three main parts: an aluminum split-block waveguide tapered to a standard waveguide flange, a thin metal plate containing

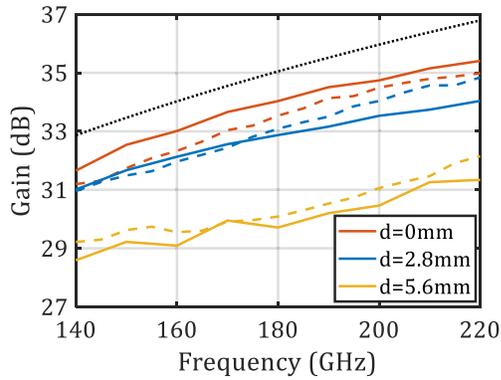


Fig. 14. Gain over frequency for broadside and steered beams. Solid line: FW simulations. Dashed line: measurements. These results do not include loss in the waveguide. Dotted black curve: maximum directivity for a circular aperture with $D = 3$ cm.

TABLE II
LOSSES IN THE FABRICATED PROTOTYPE

Freq.	Waveguide	Mismatch	Dielectric
140GHz	1.48dB	0.31dB	0.24dB
180GHz	0.88dB	0.02dB	0.32dB
220 GHz	0.86dB	0.14dB	0.40dB

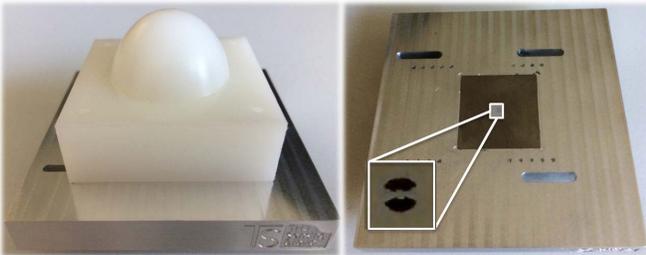


Fig. 15. Lens antenna prototype. The left-hand side shows the HDPE lens antenna, placed on the ground plane. On the right-hand side, the double-slot iris is shown. Positioning holes have been drilled at a constant distance of 1.4 mm, to displace the lens and measure the broadside and steered beams.

the double-slot iris, and the HDPE lens with the air cavity. Fig. 15 shows a picture of one of the fabricated prototypes. The waveguide and lens have been manufactured through standard milling techniques, whereas the double-slot iris has been realized by means of laser cut. The measured reflection coefficient is in accordance with the simulations [Fig. 11(b)]. The matching level shift between the lossless FW simulations and measurements is related to the waveguide ohmic losses.

Applying the analysis in reception explained in the Appendix, the near-field coupling between two identical lenses, η_c , facing each other is calculated [Fig. 16(a)]. This parameter, which can be measured with a simple setup, can be used to determine the single-antenna radiation efficiency and, therefore, gain, and to evaluate the accuracy of the estimated

losses in the lens and waveguide. Fig. 16(c) shows the lens-to-lens near-field coupling calculated semianalytically, FW simulated, and measured, with the reference plane at the standard waveguide flanges. In the semi-analytic calculation, mismatch loss, losses in the waveguide, and dielectric loss in the lens (Table II) have been added to the near-field coupling calculated with the analysis in reception. The transmission between two lenses in the FW simulation does not include multiple reflections. The difference between both curves shows the limitation in the accuracy of the GO approach used for the analytic calculation of the transmitted fields after the lens, \vec{E}_{GO}^{Sc} . In the lens-to-lens near-field coupling measurements, time gating has been applied to eliminate the contribution of multiple reflections in the lens-air interfaces. The time gating has been performed applying a window which considers only the first transmitted rays from one lens to the other as shown in Fig. 16(b). All additional transmitted rays coming from multiple reflections inside the lenses or between them are filtered in time domain, and finally, the resulting frequency response is extracted from the filtered results [plotted in Fig. 16(c)]. The time-gated measurement results show an agreement with the FW simulations within ± 1 dB in the worst case, which is translated in ± 0.5 dB for a single lens. This error includes the impact of tolerances in the measurement, fabrication, and material properties estimation, such as lens $\tan \delta$ and waveguide-effective conductivity.

An FW simulation has been performed eliminating the lenses and keeping the LWA feeds in the same position (the space between feeders is filled with dielectric). The transmission reached in this configuration is lower than -30 dB in the whole band. This value corresponds to the directivity of both feeds (13 ± 0.5 dB) minus the propagation loss in the dielectric material (55–59 dB). The presence of the lenses improves the coupling in 27–31 dB.

In order to obtain the lens far-field patterns, near-field measurements have been performed over a planar surface on top of the lens antenna, by means of a probe (i.e., an open-ended WR5 waveguide) placed on an automatic positioner. A squared surface with an area of 50×50 mm² has been measured at 8.5 cm distance from the lens, taking sampling points at 680 μ m distance (half wavelength at the highest frequency, 220 GHz). Around the measuring probe, a hollow metallic diffraction cone has been installed, in order to minimize the effect of multiple reflections between the lens and measurement equipment. A standard probe correction has been then applied to remove the small impact of the probe in the measured far-field, with the same approach presented in [32]. Fig. 13 compares the measured and FW simulated lens far-field patterns for broadside and two steered beams, showing an excellent agreement over the whole frequency band. This can be also seen in the results for the maximum gain over frequency shown in Fig. 14. The maximum antenna gain at broadside has been estimated by adding the dielectric loss, validated with the lens-to-lens coupling method, to the directivity extracted from the near-field measurements. In case of the steered beams, the gain has been evaluated by normalizing the radiation intensity with the power radiated

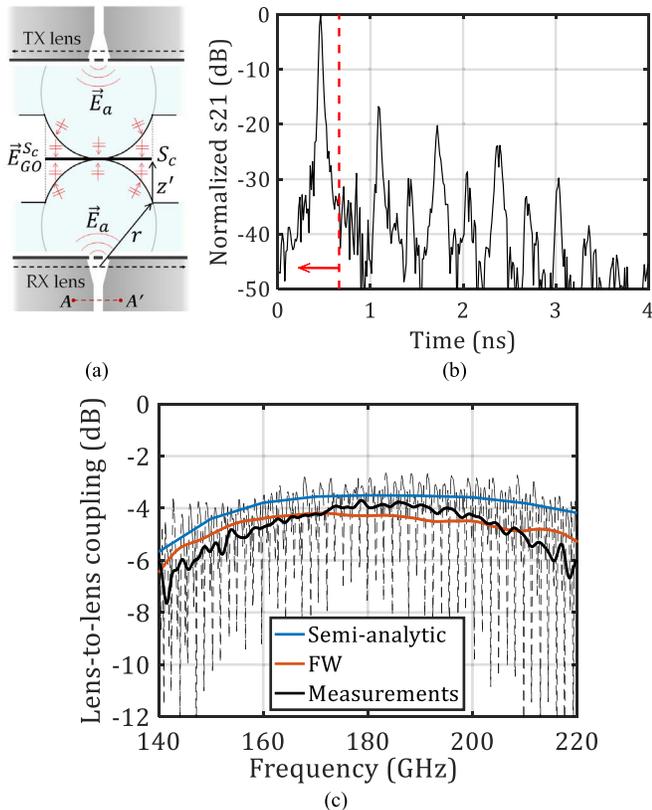


Fig. 16. (a) Analysis in reception of lens-to-lens near-field coupling. (b) Time-domain representation of the measured transmission between the two lenses. The red dashed line marks the applied time window. (c) Lens-to-lens near-field coupling efficiency where the black lines correspond to the measured S21: raw (dashed) and time gated (solid).

and dielectric loss of the broadside case. The losses in the waveguide are not included in this plot. These results confirm that 7 beams in a linear array or 40 beams in a two-dimensional array configuration can be achieved with an HDPE lens planar array, with a maximum scan loss of 3 dB. The maximum difference between the FW simulated and measured maximum directivities is 0.7 dB in the whole bandwidth and for all the measured beams, which is related to the measurement and fabrication tolerances. Several measurements were performed changing the measurement plane height. Measurement repeatability showed changes in the estimated directivity of around 0.5 dB.

The transmission between two lenses placed at different distances (50 cm, 1 m and 2 m) has been as well measured in order to validate the antenna gain estimation. Fig. 17 shows the raw and time-gated measurement results compared to the transmission calculated with Friis equation. The measured results converge to the estimation with Friis equation when the link distance approaches 2 m, where both lenses can be considered to be at far-field distance. The link measurement with 2 m distance presents higher ripple, since it was performed with a single lens placed in front of a metal plate at 1 m distance, due to the lack of cables reaching this length. This measurement in reflection has been as well performed placing the metal plate at 50 cm distance, showing the same

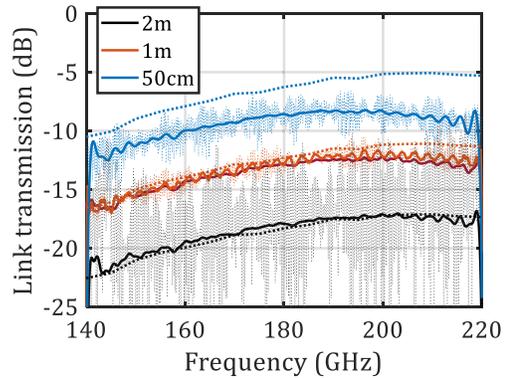


Fig. 17. Measured and estimated transmissions between two lenses placed at large distances. Thin dashed lines: raw transmission measurements. Solid lines: measurements after time gating. Dotted lines: results estimated with Friis equation. Results include both dielectric loss and conductor loss. For 1 m distance, light red: measurement in transmission; dark red: measurement in reflection.

results as the measured transmission between two lenses at 1 m distance (Fig. 17).

VIII. CONCLUSION

An analytical design methodology based on an analysis in reception is proposed to optimize the truncated elliptical lenses fed with resonant LWAs. With this aim, the aperture efficiency is expressed as a reaction integral between the lens inward spherical GO field and the field radiated by the LW feed inside the lens. Both fields can be analytically evaluated, simplifying the optimization of the lens antenna with respect to other full-wave or PO techniques. Applying this analysis in reception, design rules are derived for the use of the LWA as a feeder for elliptical lenses with different ϵ_r . Optimum geometric parameters for the feeder and lens are extracted, showing the maximum illumination efficiency and bandwidth reached for each lens ϵ_r . The steering capabilities for truncated elliptical lenses are also displayed for different ϵ_r . Following the proposed design rules, the best compromise between illumination efficiency and steering capabilities can be selected depending on the application.

A broadband integrated lens antenna is presented for future wireless communications at high frequencies. The proposed antenna is based on the resonant leaky-wave feed coupled to an HDPE elliptical lens. This design represents a trade-off with high efficiency, very large bandwidth, and low cost but limited number of off-focus beams. The lens achieves 34 dB gain at 180 GHz, reaching an aperture efficiency higher than 80% and ohmic losses lower than 0.5 dB over 44% relative bandwidth for the broadside beam. This design enables to generate up to 40 beams overlapping at -3 dB with a planar two-dimensional Fly's eye array of lenses, keeping the gain higher than 30 dB at 180 GHz. The reflection coefficient is lower than -10 dB over the whole frequency band. The antenna has been fabricated and measured, showing excellent agreement with the simulated performances.

APPENDIX

A. Lens Analysis in Reception

In [33], an analysis in reception for the optimization of reflector feeds was proposed. The power received by the feed can then be estimated using a Thevenin equivalent circuit when the reflector is illuminated with a plane wave. The open circuit voltage at the defined terminals can be expressed as a reaction field integral between the considered problem in reception and an equivalent problem in transmission. The generalization of this procedure to any geometry via the use of the equivalence theorem was described in [34]. In this appendix, a similar methodology is applied to analyze the proposed lens antenna geometry [see Fig. 3(a)].

B. Thevenin Equivalent Circuit

An equivalent Thevenin circuit for the problem in Fig. 3(a) can be derived by evaluating the open circuit induced by the incident plane wave onto the antenna terminals, AA' , inside the waveguide, the antenna input impedance Z_a , and the load impedance Z_L [Fig. 18(b)]. The open circuit field inside a single mode waveguide can be expressed as $\vec{E}_{oc} = V_{oc}\vec{e}_0$, where \vec{e}_0 is the electric-field-associated eigenvector of the considered waveguide mode (TE₀₁ in this case). The open circuit voltage, V_{oc} , can be evaluated as

$$V_{oc} = \iint_{S_{AA'}} \vec{E}_{oc} \cdot \vec{e}_0 dS. \quad (3)$$

The power dissipated in a matched load ($Z_a^* = Z_L$) can be calculated as $P_L = |V_{oc}|^2 / (8R_a)$, being $R_a = \text{Re}(Z_a)$ the antenna radiation resistance.

C. Reaction Integral

The value of the open-circuit voltage, V_{oc} in (3), necessary to evaluate the received power, P_L , can be expressed as a reaction integral between the open-circuit field in the antenna in reception and the equivalent currents characterizing the same antenna in transmission. Let us consider the antenna in transmission fed by an electric current, \vec{J}_{Ix} , with arbitrary amplitude, I_{Ix} . This current is related to the magnetic field by $\vec{J}_{Ix} = -\hat{z} \times \vec{H} = -\hat{z} \times I_{Ix} \vec{h}_0$, where \vec{h}_0 is the magnetic-field-associated eigenvector of the considered waveguide mode. Since $\vec{e}_0 = \hat{z} \times \vec{h}_0$, the integral in (3) can be then expressed as

$$V_{oc} I_{Ix} = - \iint_{S_{AA'}} \vec{E}_{oc} \cdot \vec{J}_{Ix} dS. \quad (4)$$

The antenna radiation resistance, R_a , can at the same time be related to the power radiated by the antenna, P_{rad} , excited by the same current amplitude, I_{Ix} , as $P_{rad} = (|I_{Ix}|^2 / 2) R_a$. Thus, P_L can be expressed using the reaction integral in (4) as

$$P_L = \frac{|V_{oc} I_{Ix}|^2}{16 P_{rad}}. \quad (5)$$

Equation (5) was reported in [34]. Note that the actual amplitude of the equivalent current in transmission, I_{Ix} , cancels out in this expression. Once we have expressed the

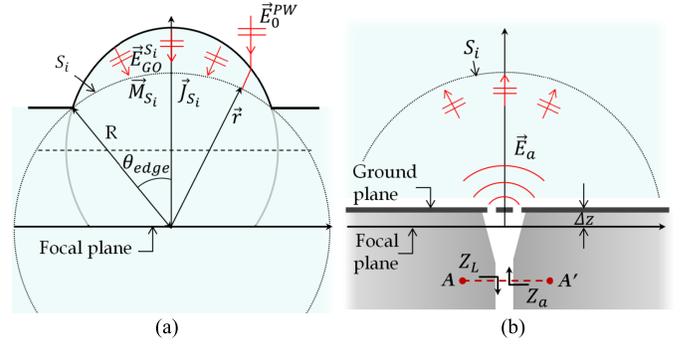


Fig. 18. (a) Boundary conditions to calculate the incident field, $\vec{E}_{GO}^{S_i}$, from the ideal plane wave impinging the lens, \vec{E}_0^{PW} , and associated induction currents in S_i , \vec{M}_{S_i} , and \vec{J}_{S_i} . (b) Boundary conditions to calculate the fields radiated in transmission mode, \vec{E}_a .

received power as a reaction integral between two different problems (currents in transmission and open circuit field in reception), we can use the reciprocity theorem to change this reaction integral to any surface, by interchanging fields and currents between both problems.

D. Aperture Efficiency Calculation

In order to carry out the lens optimization, the most convenient surface to perform the reaction integral is the spherical surface S_i , shown in Fig. 3(a) and Fig. 18, defined as close as possible to the lens surface, since here all fields involved in the reaction integral can be analytically evaluated. In the equivalent problem in reception, the field induced by the incident plane wave at the open-circuited antenna terminals, \vec{E}_{oc} , can be evaluated by means of a set of surface equivalent currents \vec{M}_{S_i} and \vec{J}_{S_i} at S_i . In the equivalent problem in transmission, the fields radiated by the feed antenna excited by \vec{J}_{Ix} , $[\vec{E}_a, \vec{H}_a]$, should be evaluated at S_i in the presence of the lens. We can, therefore, apply the reciprocity theorem and reformulate (4) as

$$V_{oc} I_{Ix} = \iint_{S_i} (\vec{H}_a \cdot \vec{M}_{S_i} - \vec{E}_a \cdot \vec{J}_{S_i}) dS. \quad (6)$$

This expression can be further simplified. On one side, the equivalent currents \vec{M}_{S_i} and \vec{J}_{S_i} used in (6) can be approximated using a GO field propagation $[\vec{E}_{GO}^{S_i}, \vec{H}_{GO}^{S_i}]$ evaluated in the absence of the antenna feed [Fig. 18(a)], as $\vec{M}_{S_i} = \vec{E}_{GO}^{S_i} \times \hat{n}$ and $\vec{J}_{S_i} = \hat{n} \times \vec{H}_{GO}^{S_i}$ being $\hat{n} = -\hat{r}$. The analytical expression of these fields in an elliptical lens was given in [26]. The diffracted fields are neglected, since S_i is taken very close to the lens surface. On the other side, neglecting the impact of the multiple reflections at the lens–air interface, the radiated fields $[\vec{E}_a, \vec{H}_a]$ used in (6) can be approximated as the ones radiated into an infinite dielectric medium [Fig. 18(b)]. If the reaction surface S_i is chosen to be in the far-field of the feed antenna, a small displacement in the Z -direction of the feeding antenna with respect to the lens focal plane, Δz , [Fig. 18(b)] can be accounted with a variable phase term intrinsic in \vec{E}_a ,

as follows:

$$\vec{E}_a(\theta, \phi) = \vec{E}_a^{GP}(\theta, \phi)e^{jk\Delta z\cos\theta} \quad (7)$$

where $\vec{E}_a^{GP}(\theta, \phi)$ is the far-field of the antenna calculated with phase reference placed at ground plane. This field can be evaluated resorting to an asymptotic evaluation of the spectral Green's function for stratified media [23], with the following approximation of the double-slot iris current:

$$\begin{aligned} \vec{m}_s(\rho, \phi) \\ \approx \frac{1}{\rho} \cos\left(\frac{\pi}{2\alpha}\left(\phi \pm \frac{\pi}{2}\right)\right) \text{rect}\left(\phi \pm \frac{\pi}{2}, 2\alpha\right) \text{rect}(\rho - \rho_0/2, w)\hat{x} \end{aligned} \quad (8)$$

where α , ρ_0 , and w are the parameters describing the iris geometry [Fig. 2(b)]. Since both the GO incident field, $\vec{E}_{GO}^{S_i}$, and the far-field radiated by the antenna, \vec{E}_a , are spherical waves, (6) can be simplified into

$$V_{oc}I_{tx} = \frac{2}{\zeta_d} \int_0^{2\pi} \int_0^{\theta_{edge}} \vec{E}_{GO}^{S_i}(\theta, \phi) \cdot \vec{E}_a^{GP}(\theta, \phi)e^{jk\Delta z\cos\theta} dS \quad (9)$$

where ζ_d is the wave impedance in the lens medium.

E. Lens-to-Lens Near-Field Coupling Efficiency

In order to determine the near-field coupling efficiency η_c between two lenses [Fig. 16(a)], one of them is considered in transmission mode, whereas the other is analyzed in reception. In this way, the field impinging the receiving lens corresponds to the field radiated by the lens considered in transmission. η_c is then calculated as the ratio between the power dissipated at the receiving lens load, P_L , and the power radiated by the transmitting lens feed, P_{rad} . Following similar steps than in the previous sections, the term $V_{oc}I_{tx}$ can be determined as the reaction integral between the fields transmitted by both lenses, $[\vec{E}_{GO}^{S_c}, \vec{H}_{GO}^{S_c}]$, in the plane between the two lenses [S_c in Fig. 16(a)]. This analysis neglects the impact of multiple reflections, inside and between both lenses. The fields transmitted at S_c can be approximated using a GO approach, where the rays are assumed to be propagating in the z -direction with a uniform phase front (valid only when the feed's phase center is at the ellipse's focus). For such case

$$\vec{E}_{GO}^{S_c}(\rho, \phi) = [\tau_{\parallel}(\rho)(\vec{E}_a(\vec{r}) \cdot \hat{\theta})\hat{\rho} + \tau_{\perp}(\rho)(\vec{E}_a(\vec{r}) \cdot \hat{\phi})\hat{\phi}]e^{-jk_0z'} \quad (10)$$

being \vec{r} the vector defining the lens surface, z' the distance between the lens surface and the integration plane [Fig. 16(a)], and τ_{\parallel} and τ_{\perp} the parallel and perpendicular Fresnel transmission coefficients as defined in [8], respectively. Since the propagation of the transmitted field from the lens surface up to S_c assumes all transmitted rays to be propagating perpendicularly to S_c , the near-field coupling efficiency, η_c , can be expressed only as a function of the electric fields as

$$\eta_c = \frac{P_L}{P_{rad}} = \frac{\left| \frac{2}{\zeta_0} \int_0^{2\pi} \int_0^{D/2} \vec{E}_{GO}^{S_c}(\rho, \phi) \cdot \vec{E}_{GO}^{S_c}(\rho, \phi) \rho d\rho d\phi \right|^2}{16P_{rad}^2} \quad (11)$$

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