A Parallel Plate Ultra-Wideband Multibeam Microwave Lens Antenna

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Abstract—An ultra-wideband multibeam microwave lens antenna operating from 8 GHz to 18 GHz is proposed. The antenna consists of four excitation ports connected to a parallel plate waveguide filled with a cylindrical dielectric slab, which serves as a lens in order to modify the cylindrical wavefront launched by the excitation ports. The output of the lens is a plane wave guided to a radiation aperture with a linear tapered flare. Four distinct fan-beams covering 40° in the azimuth plane with an elevation beamwidth of 30° and a minimum gain of 15 dBi have been achieved. The main advantages of our design include its relative simplicity, ease of fabrication, having a low profile, not requiring an antenna feed, and high-power handling capability. The design procedure is presented together with the optimization procedures that have been applied to all parts of the antenna system to achieve the desired performance. The proposed structure has been simulated with CST Microwave Studio software. There is an excellent agreement between the simulation and measurement results.

Index Terms—Lens, UWB antennas, waveguide components, multibeam antennas.

I. INTRODUCTION

Lenses are well-known structures used for modifying the shape of a wavefront by introducing an engineered delay in the path of the wave. For dielectric lenses, the lens shape as well as its dielectric constant can be controlled to provide the desired wavefront. In most microwave applications, the main purpose of using a lens structure is to increase the gain of the antenna by transforming a spherical wavefront to a plane wavefront [1].

Making a multibeam antenna system is another application, where a lens structure can become very useful. Multibeam ultra-wideband antennas, with selective directional beams, high gain and low grating lobes, are useful to many communication and radar systems including automotive radars [2]. Pencil beams as well as fan beams can be created using lens structures, where fan beams are more suitable for some Doppler radars as well as direction finding systems [3].

The Luneburg lens is one of the most well known dielectric lenses used in microwave lens antennas to create multibeam patterns. This lens has a spherical shape with a gradual variation of its relative permittivity from 2 at its core to 1 on its surface. By placing several feeds around the lens, a high-gain multibeam antenna with independent beams is created. High costs, bulkiness, and manufacturing problems are the main drawbacks of the Luneburg lens [4]. In contrast, the Rotman lens has low cost and it can be made ultra-wideband [5]; however, it gets very lossy due to the sidewall reflections [6]. Gradient refractive index (GRIN) metamaterials, which possess gradient permittivity and/or permeability are also used as lens for manipulating the wavefront and transform a spherical wave to a plane wave [7]. Moreover, using a GRIN lens makes it possible to manipulate the amplitude and phase of the aperture field distribution simultaneously.

Several variations of Luneburg lenses with fan-beam scanning capability have been designed by using parallel plate techniques [8]. In these techniques, one is interested in the propagation of the $TE_{10}$ mode of a parallel plate waveguide therefore a cylindrical Luneburg lens is designed by varying the plate spacing so that the resulting index of refraction mimics the Luneburgs law [9]. Typically, a planar linear tapered slot antenna requires an antenna feed at the focal point of the lens.

In the present work, we are also going to adopt a parallel plate structure with a "Luneburg inspired" cylindrical lens, but with two main differences: (1) instead of having an index of refraction that varies with the radial distance from the axis of the cylindrical lens, we consider a material with a uniform dielectric constant whose value is optimized to maximize gain performance while simplifying manufacturing process; (2) the replacement of the antenna feed with a coaxial connector pin to reduce the overall size and increase the bandwidth compared to [9].

As it is shown in Fig. 1a, the structure is fed by a simple coaxial cable, which creates a cylindrical wavefront (see Fig. 1b), that passes through a parallel plate waveguide filled with a single slab of dielectric with a permittivity of 2.1 (teflon).

The dielectric plays the role of a lens and modifies the cylindrical wavefront to create a planar wavefront at its output, which is shown in Fig. 2. This planar wavefront is then guided towards the aperture of the antenna and due to the uniform phase distribution on the planar wavefront a directive fan beam is created. The beam steering is simply done by exciting the corresponding input. The main advantage of the proposed structure is its simplicity, where a single slab dielectric is used as the lens structure with no need to use a layered dielectric structure and no geometrical modifications or are necessary to be applied to the parallel plate waveguide. Moreover,
II. ANTENNA SYSTEM DESIGN

As can be seen in Fig. 1, the proposed system consists of three main parts: the feed, the cylindrical lens, and the radiation aperture. The excited cylindrical wavefront, created by coaxial feeds, is guided towards the lens and modified to generate a plane wavefront at its output, where this plane wavefront is guided towards the radiation aperture. We are going to give a step by step discussion of each part.

A. Lens Design

The proposed cylindrical lens sandwiched between two parallel conducting plates is illustrated in Fig. 1. The lens, which is made out of a low-cost Polytetrafluoroethylene (PTFE) teflon material ($\epsilon_r = 2.1$ and $\tan \delta = 0.0002$), has a cylindrical cross section to give the desired step-variation of the index of refraction. The cylindrical cross section is easy to fabricate and flexible enough to obtain the desired performance. The lens is fed by four SMA connectors placed at an appropriate distance away from the edge of the cylindrical lens to obtain better matching with the lens. This distance increases the degrees of freedom to optimize the lens’ return loss value.

The mode of propagation in the proposed parallel plate structure is determined by the direction of the electric field due to the connector pin. Referring to Fig. 1, there is a sectional parallel waveguide defined by the 4 feed pins and the center of the cylindrical lens (see the region limited by OAB). Within this region, waves travel along the radial direction with the equiphase surfaces represented by curved surfaces of constant radius. In the previous work [9], the $TE_{10}$ mode is the dominant mode because the excitation ports are miniaturized microstrip antennas and microstrip antennas excite $TE_{10}$ modes inside a waveguide. However, in this paper, connector pins excite the $E_z$ field component in the waveguide. Therefore, the radial variations of this field are represented by Hankel functions and in the sectional-plate geometry the propagation mode is $TM^{2}$ with $E_z$ and $H_\phi$ components [16].

In Fig. 2, the transformations of a cylindrical wavefront to a planar one for different permittivity values are compared. One can observe that when the permittivity is lower (for $\epsilon_r = 1.5$) the shape of the wavefront is convex. Then, by increasing the permittivity one obtains a sufficiently flat wavefront (for $\epsilon_r = 2.1$). Further increases of the permittivity cause the wavefront to become concave (for $\epsilon_r = 4$) and, finally, even larger values of the permittivity cause the wavefront to differ significantly from the desired planar wavefront.

As the permittivity increases a smaller disk is needed to create a planar wavefront. However, when the wave is launched from the excitation ports, a portion of the wave leaks from the sides of the lens (assuming all the other dimensions are kept fixed), hence the sidelobe level (SLL) will increase. Therefore, there should be a compromise between the size of the system, the SLL, and the gain. As it turns out, the relative permittivity of 2.1 is the optimal value that results

![Fig. 1: The multibeam ultra-wideband antenna made of three main parts: (a) the feed structure consisting of 4 coaxial cable connectors, and the "Luneburg inspired" cylindrical lens. (b) A view of the inside parts of the proposed antenna (radiation aperture). The dimensions are given in Table I.](image-url)
Fig. 2: Simulation results showing the magnitude of the electric field of the propagating wave for lenses with different permittivities at 13 GHz. (a) $\epsilon_r = 1.5$; (b) $\epsilon_r = 2.1$; (c) $\epsilon_r = 4$; and, (d) $\epsilon_r = 10$.

in dimensions compatible with our design requirements. The planar wavefront is essential for generating a focused beam, which, in turn, depends on the location of the feed. In order to achieve the best illumination, the phase center of the SMA feed connectors must be located at the focal point of the lens and the radius of the lens should be more than $\lambda_g/4$ where $\lambda_g$ is the wavelength inside the lens at the lower frequency of the design bandwidth [9].

It is interesting to note that in a Luneburg lens antenna designers use a dielectric profile where $\epsilon_r$ varies from the center of lens to the lens surface. The variation of $\epsilon_r$ is related to the radius of lens. However, lens fabrication process is very difficult. In this paper we succeeded to obtain excellent result only by using single material (PTFE) with a uniform value of the dielectric permittivity $\epsilon_r = 2.1$, which makes the fabrication easier compared to other designs based on the use of the Luneburg lens.

B. Aperture Design

The radiation aperture of the lens is the final step in the antenna system design. A radiation aperture is chosen among different types of the antennas because: (1) it has a widebandwidth behavior; (2) it does not require a matching circuit, such as in the case of microstrip antennas; (3) it can handle higher levels of RF power compared to microstrip antennas. In addition, the lens should have a wide fan beam for each

Fig. 3: Parameter definitions for a narrow horn antenna to determine the radiation beamwidths.

Fig. 4: Simulation results of the realized gain at $f = 13$ GHz for lenses with different permittivities. (a) $\epsilon = 1.5$; (b) $\epsilon = 2.1$; (c) $\epsilon = 4$; and, (d) $\epsilon = 10$.

Fig. 5: Simulation results for (a) realized gain for lossless and lossy dielectric lens. (b) frequency response of the realized gain and radiation efficiency.
Fig. 6: Simulated radiation pattern for excitation of (a) port 1; (b) port 2; (c) port 3; and, (d) port 4.

excitation port so that a narrow aperture can be used for this structure. Accordingly, we can approximate the radiation aperture as a narrow horn antenna and calculate the initial dimensions of the aperture using the approximate formulas for the azimuth and elevation plane radiation beamwidth for a narrow horn antenna [17]. Referring to Fig. 3, the azimuth and elevation plane radiation beamwidth are

\[
\psi_{\text{Az}}^{3dB} = \frac{53 \lambda}{b}, \\
\psi_{\text{El}}^{3dB} = \frac{68 \lambda}{a},
\]

where \( b \) corresponds to the distance \( L_5 \) in Fig. 1b, \( a = r\alpha \) is the effective horizontal width for the lens aperture, where \( \alpha = 60\pi/180 \) and \( r = \lambda_g/4 \) is the radius of the lens. These expressions provide the initial values for the design parameters in Fig. 1. Then, the final dimensions are obtained using the Particle Swarm Optimization feature of the CST software. Table II reports the optimized values and shows that there is good agreement between the initial approximations obtained with formulas (1) and (2), and the simulation results for port 1.

Fig. 4 compares the radiation patterns of lenses with different relative permittivities at 13 GHz, because this is the center frequency of operation. Similar to the results shown in Fig. 2, the gain improves going from \( \epsilon_r = 1.5 \) to \( \epsilon_r = 2.1 \), then it becomes worse for larger values of the dielectric relative permittivity. We note also that \( \epsilon_r = 2.1 \) is achieved with teflon that has low values of \( \tan\delta \) at high frequencies (more than 10 GHz). To better understand the effect of dielectric losses, the realized gain of the antenna is simulated versus the frequency in Fig. 5a. While it is clear that there is an advantage at having a dielectric with low \( \tan\delta \), the effect of \( \tan\delta \) in the realized gain is more significant at higher frequencies by about 1 dB. Fig. 5b shows the realized gain of the antenna and its efficiency. One can observe that there is a variation of about 5 dB within the operation bandwidth and that the efficiency is maximum around 14 GHz. Since the behavior of both the gain and the efficiency are not linear with the frequency, this explains the reason for designing the antenna and its feed at the center frequency.

It should also be mentioned that the dispersion, i.e. frequency dependent permittivity, affects also the shape of the wavefront at different frequencies. Therefore, low dispersion and low loss materials are more suitable for lens applications. Since the input ports are placed next to each other, the coupling between them is inevitable. As can be seen in Fig. 6, by exciting each input port, a distinct beam is created in the desired direction.

C. Feed SMA Design

Four coaxial connector pins are used to excite the lens because connector pins have the advantage of having small size compared to the use of microstrip antennas and the transition from the coaxial pin to the waveguide has wider bandwidth than a microstrip antenna. Referring to Fig. 1, the four pins are placed in air between the dielectric lens and the reflector wall AB. The connector pin and the coaxial probe have an inductive behavior, which limits the whole frequency bandwidth. Hence, this inductive behavior can be compensated by using a capacitive loading. Referring to Fig. 1, the capacitive loading is introduced by adding a disk to the end of the pin, so as to create a capacitance between the floor of the waveguide and the disk itself. The disk acts as an impedance

<table>
<thead>
<tr>
<th>Plane</th>
<th>8 GHz</th>
<th>10 GHz</th>
<th>13 GHz</th>
<th>18 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \phi_{\text{cal}} )</td>
<td>19.7°</td>
<td>16.6°</td>
<td>15.1°</td>
<td>9.1°</td>
</tr>
<tr>
<td>( \phi_{\text{cal}} )</td>
<td>61°</td>
<td>46.5°</td>
<td>37.1°</td>
<td>31.1°</td>
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Table II: Comparison of the simulation and calculated radiation beamwidth for port 1
where the design parameters $R_{disk}$ and $h_c$ can be tuned to achieve the desired matching. The distance $L_6$ of the connector pin from the cavity wall is initially set to be a quarter wavelength at the central frequency (13 GHz) so that this results in an open circuit seen by the coaxial probe. Moreover, the approximate value of the capacitance introduced by the disk is

$$C = \epsilon_0 \frac{\pi R_{disk}^2}{d}$$  \hspace{1cm} (3)$$

where $R_{disk}$ is defined in Fig. 1 and $d$ is the distance between the disk and the floor of the waveguide [18]. The location of the feed and the geometrical parameters that define them were then fine-tuned using CST Microwave Studio software with the goal to maximize the bandwidth and minimize the reflections.

In order to create a strong coupling between the coaxial probe and the cavity the distance $d$ is chosen to be very small. To further justify this design, we present in Fig. 7b the cross section of the overall antenna system and its equivalent transmission line model as it is seen from the feeding port. The elements that appear in Fig. 7b are: $Z_d$ is the characteristic impedance of the transmission line representing the dielectric lens ($\theta_d$ is its electrical length); $Z_w$ is the characteristic impedance of transmission line representing the free space between the dielectric lens and the disk ($\theta_w$ is its electrical length); $Z_{p1}$ is the characteristic impedance of the transmission line representing the connector pin ($\theta_{p1}$ is its electrical length); $Z_{p2}$ is the characteristic impedance of the transmission line representing the disk ($\theta_{p2}$ is its electrical length); $Z_0$ is the characteristic impedance of the transmission line representing the free space between the pin plus the disk and the reflector wall AB ($\theta_h$ is its electrical length).

The stored reactive energy in proximity of the coaxial pin and the top wall causes a reactive impedance $-jX_B$, with [18]

$$X_B = \frac{\int_0^\infty \pi r_0^2 \nu^2}{cL}$$ \hspace{1cm} (4)$$

where $r_0$ is the radius of the coaxial feed and $L = L_7$ is the distance between the coaxial feed connectors, shown in Fig. 1.

The connector pin impedance $Z_p$ is given by the expression for an unbalanced cylindrical stripline [19], [20]

$$Z_p = \frac{60}{\sqrt{\epsilon_{ef}} \ln \frac{2h}{a}}$$ \hspace{1cm} (5)$$

Table III: Measured and simulated gain and beamwidth of port 1 at 4 frequencies

<table>
<thead>
<tr>
<th>Port 1 Simulated</th>
<th>Measured</th>
</tr>
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<tbody>
<tr>
<td>$f$</td>
<td>$\phi$</td>
</tr>
<tr>
<td>8 GHz</td>
<td>19.7°</td>
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<tr>
<td>10 GHz</td>
<td>16.6°</td>
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<tr>
<td>15 GHz</td>
<td>15.1°</td>
</tr>
<tr>
<td>18 GHz</td>
<td>9.1°</td>
</tr>
</tbody>
</table>

where $\epsilon_{ef}$ is the effective dielectric for the coaxial transmission line when the left region (PTFE) is filled with dielectric and the right region (from the connector pin to the reflector wall) is air. The distance between the transmission line and ground is $h = d$ and for $Z_{p1}$, $a = r_0$ and for $Z_{p2}$, $a = R_{disk}$ is the radius of the disk. The radial waveguide impedance can be easily given by [16]

$$Z_p^\rho(TM^2) = \frac{E_z}{-H_\phi} = -j\beta_p \frac{H_\nu^{(2)}(\beta_p \rho)}{\omega\epsilon_h H_\nu^{(2)'}(\beta_p \rho)}$$ \hspace{1cm} (6)$$

where $H_\nu^{(2)}(\beta_p \rho)$ and $H_\nu^{(2)'}(\beta_p \rho)$ are the second kind Hankel function and its derivative, respectively, $\alpha = 2\pi - AOBL$ is the central angle of the lens from the radiation part of the antenna, $\rho$ is the radius in the cylindrical coordinate system and $\beta_p$ is the propagation constant in $\rho$ direction given by

$$\beta_p = \sqrt{\beta_0^2 - \beta_z^2} = \sqrt{\omega^2 \epsilon_r \epsilon_0 \mu_0 - \frac{n\pi}{h}} n = 1, 2, ...$$ \hspace{1cm} (7)$$

The above formulas provide good initial approximation for the design. Then full wave CST Microwave Studio software and particle swarm optimization, determine the dimensions given in Table I.

III. FABRICATION AND MEASUREMENT

In this section, radiation pattern measurements as well as S-parameter measurements are discussed for the fabricated antenna. The measured mutual coupling between the ports are shown in Fig. 8. The measured VSWR of all ports are shown in Fig. 9. As can be seen the bandwidth for VSWR<3 is 10 GHz. At each measurement step, one of the input ports is excited and all the other ports are terminated to a matched load. The coupling between ports is less than

Fig. 8: Measurement results showing $S_{11}$ and coupling between the ports.

Fig. 9: Measured VSWR of the antenna.
Fig. 10: Measured radiation gain of the antenna at (a) 8 GHz; (b) 13 GHz (c) and 18 GHz.

12 dB throughout the frequency band of operation. Due to the structure symmetry, other ports have the same coupling. Table III compares results of the simulations and measurements for beamwidth and gain at different frequencies, which show a good agreement between simulations and measurements. The overlap between neighboring radiation patterns are also shown in Fig. 10.

IV. CONCLUSION

A compact ultra-wideband multibeam microwave lens antenna for operation in the frequency range from 8 GHz to 18 GHz has been proposed. A parallel plate waveguide structure with 4 input ports with a dielectric lens to manipulate the wavefront and a linearly tapered radiation aperture has been designed an optimized using particle swarm optimization to achieve minimum return loss and desired radiation pattern over the frequency band of operation. Four distinct radiation beams can be generated by the proposed structure, which covers a scan angle of 40° in the azimuth plane. The beamwidth in elevation plane is about 30°. The main advantage of the proposed structure lies in its simple design and high power handling capability. This design can be easily modified by, for example, increasing the number of connectors to obtain more beams with an overall higher gain value. For the excitation a simple coax input is used without any need for complicated antennas such as linear tapered slot line. A single dielectric slab is used to manipulate the curvature of the wavefront and obtain a planar wavefront, without any need for multilayer dielectric structures such as Luneburg lens. The manipulated wavefront is then guided towards a radiation aperture resembling a parallel plate horn antenna. This antenna has been comprehensively simulated and optimized and the final design has been fabricated and measured. A good agreement between measurements and simulations has been achieved.

REFERENCES


