A Millimeter-Wave 2D Beam Steering Antenna Using Extended Hemispherical Dielectric Lens Antenna Subarrays

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ABSTRACT Recent work has proposed that millimeter-wave beam steering antennas consisting of lens antenna subarrays (LASs) reduce hardware complexity. This manuscript extends the concept to 2D beam steering with extended hemispherical dielectric lenses (EHDLs). To accomplish this, we introduce a design process to maximize scan range and side lobe level (SLL) performance. The design process first employs the solution of the geometric disk covering problem to identify the initial positions of the feed antennas such that the subarray size, $M$, is minimized. This process is followed by systematic 3D full-wave simulation-based parametric sweeps of lens geometry and feed antenna positions to maximize scan range and minimize SLL. Finally, we demonstrate this process with a 38 GHz antenna consisting of $L = 7$ LASs and $M = 17$ feed antennas per LAS. The resulting antenna has a $\pm 36^\circ$ field of view, $-9.5$ dBC SLL, $5^\circ$ half-power beamwidth, and $\sim 20$ dBi maximum realized gain. Compared to the existing literature on subarray-based beam-steering antennas, this antenna performs with a more extensive scan range while offering a comparable SLL performance.

INDEX TERMS Beam steering, lens antennas, millimeter-wave, phased arrays, three-dimensional (3-D) printing, subarray.

I. INTRODUCTION Modern mm-wave wireless communication systems have long accepted that access points and user equipment must rapidly adapt to a changing environment while being resilient to path and blockage losses. A common approach to addressing these needs is the utilization of high-gain phased arrays that can provide beam steering by equipping a phase shifter per antenna element [1]. We can further improve the data rate of such a system by utilizing a hybrid multiple-input-multiple-output (MIMO) architecture. These are formed by linearly scaling up the phased array electronics with the number of desired signal chains and corresponding analog-to-digital/digital-to-analog converters [2], [3]. However, the drive to include many antenna elements and multiple signal chains makes hybrid MIMO architectures very complex to implement, high in operating cost, and hungry in DC power consumption [4]. For large arrays, DC power consumption of control components such as phase shifters (through their loss compensating or variable gain amplifiers) becomes a significant issue that motivates researchers to investigate alternative architectures [5]. Researchers have also recognized phase shifter-associated hardware complexity and cost as a challenge due to many control and bias signals that need routing per the included phase shifters. There have been recent improvements in reducing this complexity as one can incorporate entire transmit/receive modules on a single BiCMOS chip [6], yet, difficulty in integration of large antenna arrays continues.

A well-recognized technique to address the hardware complexity of large antenna arrays is the utilization of subarrays [7], [8]. In traditional subarray design, one groups a large
$N$ element antenna array into $L$ subarrays, each consisting of $M$ elements. Each subarray is then fed with a phase shifter, reducing the total number of phase shifters by a factor of $L$. This approach is not without its performance drawbacks. The array factor ($AF$) of the $N$ element antenna array becomes $AF = AF_p \ast AF_s$ where $AF_p$ and $AF_s$ stand for the primary (subarray) and secondary (array of subarrays) $AF$s. Since $AF_s$ exhibits grating lobes, $AF$ suffers from high side lobe level (SLL) and reduced scan range problems. Literature presents scan range and SLL enhancement with concepts such as overlapped/interleaved subarrays [9] and randomly distributed subarray elements [10], [11]. While these improvements show it is possible to reduce the SLL down to $\approx 15$ dB, they achieve scan ranges on the order of $\pm 10^\circ$ to $\pm 20^\circ$.

Alternative techniques proposed for increasing the scan range rely on vector summation with variable gain amplifiers [12] or directly altering the phase or amplitude of certain subarray elements [13], [14], [15]. However, the resulting scan range remains below that of traditional phased array antennas. In a departure from the standard techniques, we have recently shown that a lens antenna subarray (LAS)-based mm-wave antenna can provide an extensive scan range with reduced hardware complexity [16]. Fig.1 depicts the generic LAS-based antenna architecture for a single RF chain. Each subarray incorporates a microwave lens and a single-pole $M$ throw (SPMT) switch network to introduce electronic beam steering in $AF_p$.

To maintain a low antenna profile and perform with similar spectrum efficiency to traditional phased array systems, one must design LAS-based systems with electrically small lenses [17], [18]. Since the performance of electrically small lenses may suffer from issues such as multiple internal reflections, smaller effective aperture, and high inter-lens couplings, the design of a LAS-based antenna with a wide scan range and low SLL becomes challenging. Reference [16] pursued the first LAS-based antenna design with extended hemicylindrical slab lenses to demonstrate the concept with 1D beam steering. It operated at 38 GHz with $-9.4$ dB SLL and with a $\pm 37.5^\circ$ 1D scan range. In [19], we extended the concept to 2D beam steering by utilizing extended hemispherical dielectric lenses (EHDL) and by introducing a new design process to achieve a simulated scan range of $\pm 40^\circ$, $-9$ dB SLL, $5^\circ$ half power beamwidth (HPBW) and $>20$ dBi maximum realized gain from a $L = 7$, $M = 17$ LAS-based antenna. This strategy is different from switched lens antenna designs [20] as it considers an array of multiple electrically small lenses while scanning in all azimuths.

Section II provides the details of the proposed design process that could not have been included in [19] due to the page limitations. Section III presents for the first time an experimental characterization of a 2D beam-steering LAS-based antenna design to confirm the validity of the proposed design process. Finally, we show that the presented design performs with the best scan range compared to other state-of-the-art subarray designs reported to date while maintaining a comparable SLL.

II. DESIGN PROCESS FOR 2D LAS-BASED ANTENNAS

The 1D LAS antenna design presented in [16] is based on maximizing the gain of the center feed antenna under a single lens. This approach was unsuitable for carrying out the presented 2D LAS design as it led to both a small scan range and high SLL. An essential need in 2D LAS design is to consider the beam steering and SLL performance when one positions the feed antenna out of the $x$ and $y$ axes to steer the beam towards different elevation and azimuth directions.

A. BEAMWIDTH CONTOURS

The possibility that side lobes appear out of the beam steering plane of the primary lobe presents a critical design challenge. For example, when the feed antenna is translated away from the center of an EHDL (20 mm diameter and 12 mm height) by 7 mm while making a $30^\circ$ angle with the $x$-axis, the maximum of the radiation pattern manifests in the $\phi = -150^\circ$ cut (see Fig.2a). However, this pattern does not accurately represent the SLL. As seen in the 3D radiation pattern plot in Fig.2b, the peaks of the side lobes are not necessarily in the same azimuth cut of the main beam. To capture the side lobe behavior in the design process, we extract beamwidth contours from the 3D radiation pattern and plot the contours in a polar plot where radial distance represents the elevation angle $\theta$ and angular position corresponds to the azimuth angle $\phi$. One can create contours for absolute gain values or relative values normalized to the maximum gain. The $-7$ dB contours in Fig.2c illustrate the gain contours of the 3D radiation pattern from Fig.2b that are $7$ dB below the maximum realized gain. As shown in this figure, there exist side lobes on the $30^\circ$ and $-150^\circ$ azimuth cuts along with the $-170^\circ$ azimuth cut. From these contour plots, one can evaluate side lobe magnitude and location as well as the direction and shape of the main beam. For example, if one desires an SLL less than $-7$ dB, then the contour plot should have only one contour corresponding to the main beam. The presented example fails with a total of four contours.

Fig.3 illustrates the proposed design process for the 2D LAS-based antenna. The procedure is iterative in adjusting the lens’ geometry and spacing based on specific design criteria such as scan range, SLL, and gain variations by utilizing the beamwidth contours extracted from full-wave simulated 3D radiation patterns.
B. LENS AND FEED ANTENNA

We selected the lens geometry to be an EHDL [21] as the geometry is simple and well suited for fused deposition manufacturing (FDM) based 3D printing. EHDL geometry is also the natural progression from the 1D extended hemicylindrical slab lens used in [16] since one can obtain an EHDL by rotating the cross-section of the 1D lens about its primary axis. Although the manufacturing capability allows for complex lens shapes, the lens surface is kept spherical for its enhanced off-axis steering ability [22]. Identifying and selecting the best material for realizing the lens is beyond the scope of this manuscript. As such, we selected acrylonitrile butadiene styrene (ABS) for its wide availability in 3D printing and well known dielectric properties ($\varepsilon_r = 2.4$, $\tan\delta = 0.006$) from recent research [23].

For the feeding antenna, we utilized an aperture-coupled patch antenna which we designed as described in [16] within the presence of a semi-infinite ABS material. This decision is a departure from most traditional lens antenna designs, where the feed antenna type and its characteristics are considered carefully in terms of efficiently illuminating the lens surface while minimizing the side lobes. Such feed antennas are physically large, which prevents dense packing on the focal plane. Therefore, these feeds cannot provide beams intersecting at their half-power beamwidth (HPBW) contours when packed next to each other. This strategy contrasts with our case, in which it is a design goal to pack many feeds on the focal plane. Additionally, the small size of our lens further complicates feed optimization as radiation characteristics are affected by the lens surface and internal reflections. Still, prior works ([24]) show that patch antennas form simple and effective feeds for small lenses, motivating their use in this work. We designed the patch as in [16], but to be well-matched when placed directly under the dielectric lens without any air gaps. The substrate material was chosen as RO4003C with $\varepsilon_r = 3.35$ and $\tan\delta = 0.0027$. The design exhibits a 7% $|S_{11}| < -10$ dB bandwidth which is well-suited for operation in the 37 to 40 GHz mm-wave band with geometry detailed in Fig.4.

The primary design parameters for the EHDL are diameter and extension length. We select the initial subarray size to equally subdivide the 2D planar aperture of a half-wavelength ($\lambda_0/2$)-spaced phased array. For example, if we were to replace an $N = 64$ element square $\lambda_0/2$ spaced phased antenna array aperture with a LAS-based antenna, a potential choice could be to pick $L = 4$ with $M = 16$ feed antennas to subdivide the array by four. The lens diameter can be approximated from

$$D \approx \frac{\lambda_0 \times \sqrt{M}}{2}$$

which at a center frequency of 38 GHz leads to 15.8 mm. Note that this is a diameter of $2\lambda_0$, which makes the lens
electrically small. For the extension length, the initial value can then be set by following the hyper-hemispherical lens equations provided in [21], in our case leading to an extension length of 7.8 mm. This length is one of the primary parameters that will be modified while following the design flowchart.

C. FEED ANTENNA POSITIONING BASED ON THE GEOMETRIC DISK COVERING PROBLEM

The next consideration is positioning the feed antennas on the focal plane of the EHDL. For this design, we first consider the HPBW contours of the feed antennas. As the shape of these contours is nearly circular, one can approximate its diameter from the full-wave simulated pattern of a feed antenna located at the center of the focal plane. Then, the design goal is to determine feed antenna positions so that the HPBW contours overlap to cover the lens’s field of view. Taking this idea, we can reformulate the feed positioning problem into the geometric disk covering problem (GDC). GDC solves for the position and radius of M circles that cover a unit circle. One can consider these M circles as the HPBW contours from each feed antenna, and the unit circle represents the entire field of view of the lens. For example, if M = 7, the GDC solution gives that the radius of each covering circle is 0.5 with centers positioned on a triangular grid, as shown in Fig. 5a. If the HPBW of the feed antennas is ±10°, the unit circle represents a scan coverage of ±20°. For this example scenario, Fig. 5 shows the scan range as the number of antennas is varied from M = 1 to 20. This plot can determine the number of feed antennas needed to achieve the desired scan range.

While some M values lead to trivial solutions, solving the disk covering problem for any given M is nondeterministic polynomial-time complete (i.e., NP-Complete) [25] and requires an iterative or approximate solution such as a black-box optimization from [26]. In this work, we used black-box optimization to solve the constrained non-linear optimization problem. The implementation follows [26] in that we first rasterize a unit circle with a high number of points. Then, we use the M circles to calculate the coverage ratio (via the number of points enclosed by at least one of the M circles divided by the total points within the circle). An adaptive particle swarm optimizer [27] is then used to maximize this coverage ratio while constraining the center positions to fall within the circle. The result of this optimization is the set of covering circle center positions.

Next, we must transform the solved circle centers into physical feed positions. The circle centers from the GDC solution are points in spherical space. The azimuth components of these points map directly to feed positions, but the elevation components do not. We use the relation derived in [28] to relate angle to translation for lens antennas via

\[ d = E \tan \gamma \]  

where \( d \) is the translation from the center axis, \( \gamma \) is the elevation angle, and \( E \) is the extension length. Plugging in our extension length and GDC elevation angles to this formula gives the radial component of the polar position of the feeds on the focal plane. However, there are several issues with this simple solution, such as changes in HPBW shape [24] for off-axis elements, as well as complexities from the electrically small lens. Hence, after obtaining an initial mapping from GDC to positions via this relation, we still carry out several full-wave simulations to adjust the feed antenna positions until the simulated scan coverage matches the prediction.

D. DESIGN FLOWCHART ITERATIONS

We carry out the presented LAS-based antenna design with the assistance of programs written in the Julia programming language [29]. Julia is a free, high-performance, open-source programming language designed specifically for scientific computing. Binaries are available on the Julia Language website. We have packaged many calculations for following the design flowchart into a general-purpose Julia library freely available on GitHub, Antennas.jl, that provides an easy-to-use interface for working with antenna and antenna array pattern data. Antennas.jl can ingest exported pattern data from HFSS via a comma-separated value (CSV) file. Then, using the Plots.jl library and Contour.jl library, the beamwidth contours can be plotted at the desired level.

The goal is to produce a 2D beam steering LAS-based antenna with at least ±37.5° scan range and SLL below —9 dB. Meeting this goal will produce an antenna that meets or exceeds the performance of the previous design presented in [16] while achieving beam steering in 2D. We first start with the initial lens geometry detailed in the previous section. Then, we extract the HPBW contour of the feed antenna placed in the center of the focal plane using full-wave simulations and Julia programs. Next, the GDC problem is solved for \( M = 16 \) to estimate the total scan range and relative feed positions. If the scan range is smaller than desired, one can increment the number of feed antennas and solve the GDC problem again. If the scan range from the GDC solution is promising, Eq. 2 is employed to determine the initial feed antenna positions. At this stage, achieving the scan range may cause some feed antennas to get very close or collide with each other (particularly the antennas within...
the vicinity of the center of the focal plane). This proximity issue arose in the design pursued in this manuscript. Hence, following the design flowchart, \( M \) was increased by one, and we repeated the design process from the GDC solution. There is an implicit exit condition in this design process within the first loop, such that incrementing \( M \) by one has an upper limit due to the fixed size of the lens from the initial design geometry. If the initial lens diameter is too small, there would not be enough area on the focal plane to pack the patch antennas. In this case, one would increase the lens diameter and restart the design process. The diameter increase should be on the order of a half-wavelength or larger to make a drastic change in initial conditions. On the other hand, the modify lens geometry task encountered in later stages of the design process is reserved for small geometry changes to make fine adjustments.

Using Eq. 2, we can compute the approximate patch positions of this design from the initial lens geometry of \( E = 7.8 \) mm. Our GDC solution had circle centers on three rings, 15\(^\circ\), 25\(^\circ\), and 30\(^\circ\). Therefore, the patches are arranged on the focal plane in three rings with distance from the center set by Eq. 2 and azimuth set directly by GDC. These feed positions must then also be checked for SLL and gain variation. The lens geometry primarily controls these metrics. Like HPBW, SLL is checked through the beamwidth contours by drawing contours at the \(-9\) dB level. An equal number of contours and feed antennas imply that the design has no higher sidelobes. If SLL and gain variation do not meet the desired specifications, one must modify the lens geometry and return to the GDC problem. For the design reported in this manuscript, we found extension length adjustment to be satisfactory to achieve an SLL level below \(-9\) dB without any adjustments in the lens diameter. Specifically, the total lens height providing the desired performance was determined to be 16.8 mm, implying an extension length of 8.9 mm. This is 1.1 mm taller than the \([21]\) solution. Again, using Eq. 2, we solve for the patch positions using this extension length. An extension length of 8.9 mm resulted in offsets from the center of 2.1 mm, 3.6 mm, and 4.5 mm. Adjustments are made to compensate for the small-lens distortions which result in the final feed antenna positions and their HPBW contours shown in Fig.6a and Fig.6b, respectively. The single-lens performance for the center excitation is depicted in Fig.7. The lens has a maximum gain of 12.6 dBi and directivity of 14.2 dBi, yielding a radiation efficiency of 70 % due to the conductor and material losses.

Once the design of the LAS is complete, we proceed to estimate the performance of the LAS within an array to form the LAS-based antenna. We consider an \( L = 7 \) LAS-based antenna for the presented design, where we arrange the array elements on a triangular grid. Ideally, \( L \) does not impact the scan range. However, mutual coupling among the lenses may impact the SLL and change the scan range achievable with desired SLL performance. Hence, one should verify the idealistic analysis based on array factor multiplication with full-wave simulations as in Section II-E.

An SPMT switch network will select a feed antenna under each LAS, which will be identical to all other LASs within the antenna. Feed antennas will be excited equally in amplitude. Consequently, the array factor (i.e., \( AF_s \) based on the description given in the introduction) will steer the main beam within the HPBW contour of the selected feed antenna by applying proper phase shifts. In creating the array, we assume there is a small gap between the lenses such that the lenses do not touch each other. This spacing is necessary to avoid the meshing issues encountered in the EM solver, which results in significantly longer simulation times for the entire LAS-based antenna. A gap of 0.1 mm (noted as \( G \) in Fig.4) was found not to impact the performance of the antenna while avoiding the aforementioned issue.

The Julia package Antennas.jl includes an array factor tool that implements the array factor analysis. First, we import the 3D pattern data exported from HFSS for each feed antenna element excitation. Then, we identify the direction of maximum radiation for each feed antenna. We approximate the HPBWs of all feed antennas to be identical to that of the center-fed antenna from previous simulations. For each feed antenna, we generate array factors to steer the \( L = 7 \) element array to the direction of the feed’s maximum radiation along with directions that trace out its HPBW contour. Subsequently, we multiply these array factors by the single lens radiation pattern generated by the excited feed antenna. The HPBW and SLL contours of these patterns can be plotted.
in the same fashion as the single lens to observe performance. One may need to adjust element spacing if the design is not performing well. However, we needed no such adjustment for the presented design as we achieved ±36° scan range in every azimuth cut and up to ±42° in some azimuth cuts with SLL better than −9 dB, as desired in design specifications. The HPBW contours for the \( L = 7 \) element LAS-based antenna are depicted in Fig.8.

**E. FULL-WAVE LAS-BASED ANTENNA SIMULATIONS**

The final step in designing the LAS-based antenna is to simulate a complete connectorized antenna model as seen in Fig.9a. As this step is computationally intensive, we performed this simulation once when we considered the performance estimated from the array factor analysis promising enough to proceed further with the fabrication and testing.

Simulating the complete model with all \( L \times M = 119 \) feed antennas excited was impossible with our available computational resources. Providing an experimental capability to individually excite all feed antennas was also impossible without integrated circuit incorporation due to the size of coaxial cables and connectors. Therefore, a feeding network that excites only a single feed under each subarray was designed and utilized for our practical measurements. Specifically, we designed the feed network to excite the center feed. Under this scenario, the feed lines are brought towards the edge of the board to provide excitation through 2.92 mm vertical-launch connectors (PE45451, Pasternack). These connectors are included in the full-wave model using the manufacturer-provided 3D connector model with the addition of material properties and wave port excitations. This approach is equivalent to leaving all other feed antennas within the subarray open-terminated. Single-lens simulations show that exciting a single feed antenna while others are open terminated does not alter the radiation pattern, justifying this measurement approach. Adjacent feed antennas within the subarray exhibit mutual couplings less than −15 dB, whereas non-adjacent feed antenna couplings are much below this value. This resulting feed network is shown in Fig.9b.

Since we fixed the feed network to excite a single antenna element under each lens, different feed network boards are needed to cover all beam steering possibilities (a full electronic beam steering implementation is beyond the scope of this manuscript). On the other hand, the cost of 3D printing lenses is significantly lower in contrast to PCB manufacturing. An alternative is to print shifted lens assemblies paired with the center-fed PCB such that the excited element is in a different relative position under the lens. More precisely, we can effectively shift the center feed to the position of the other feeds of interest. Simulation results show that the radiation pattern remains approximately the same in this scenario as if the LAS-based antenna had the off-axis feeds excited directly. Hence, we used this technique in this experimental design for cost reduction.

Ideal connector models and equal length delay lines employed in the simulation model emulate the condition when subarrays are excited in equal phase and amplitude, resulting in broadside radiation. The ports at the connectors can also be excited with relative phase delays to simulate the beam steering performance when subarrays are excited with equal amplitude but with the proper relative phase differences. The experimental setup also follows this model, where we individually measure radiation from each subarray and sum the patterns in software with the desired relative phase differences. Section III explains that the experimental setup also employs a calibration process before the pattern summation to remove uncertainties associated with measurements.
Full-wave simulation of the complete model provided beam steering results that are consistent with those obtained from the array factor analysis with ±36° coverage and SLL better than −10 dB. The maximum realized gain is 19.8 dBi (including connector loss) with a directivity of 22.8 dBi. Due to the similarity with array factor analysis (Fig. 8), we chose to omit the full-wave HPBW contour plot from this manuscript.

III. EXPERIMENTAL VERIFICATION

The lenses are 3D printed on a Prusa MK3S printer with standard natural-colored ABS filament with a 0.4 mm nozzle at 0.2 mm layer heights with 100% infill. The entire lens assembly is joined to a single piece by adding a 0.5 mm joining layer (noted as $H_{\text{Join}}$ in Fig. 4). This addition does not increase the extension height. Holes in the lens assembly design match the mounting holes of the vertical-launch K-connections on the PCB and act as an alignment mechanism between the PCB and lens. We utilize screws passing through these holes as the mechanism to bring lens assembly and feed board in complete contact. However, the joining layer used in the 3D printed lens assembly is thin and prone to bending. This likely causes some air gap above the patch antennas and manifests as irregularities in $S_{11}$ measurements. To hold the antenna upright, we 3D printed a test stand as well on the Prusa MK3S with black polyethylene terephthalate glycol (PETG). This complete setup is depicted in Fig. 10.

$S_{11}$ of the completed assembly was measured from 30 GHz to 40 GHz (Fig. 11). All excited feed antennas are well matched at 38 GHz. Despite the identical connectors, cables, and feed line lengths, variations exist among the $S_{11}$ responses. In addition to the potential air gap, another reason for this variation is due to the compression-based connectors. We observe that slight perturbations in their tightening may manifest in observable changes in the $S_{11}$ response. Despite being matched, such variations may impact the phase response, necessitating the calibration approach described below. We compensated the measured gain data for the simulated feed line losses and manufacturer-reported connector losses. Since the measured gain matches well with the simulation, we can conclude that the feed antennas are well-matched as designed.

The gain pattern from each feed antenna (i.e., subarray) was measured individually, recording amplitude and phase with a standard gain horn. In software, we combined the measured amplitude and phase data from each subarray to form the realized gain pattern of the entire LAS-based antenna. In addition to the non-idealities mentioned above (i.e., air gaps and compression of connectors), there is additional non-ideality associated with vibration and movement. At 38 GHz, 0.5 mm of movement is equivalent to $\pi/8$ of phase error and enough to significantly distort far-field measurements. Such slight movement readily occurs during the disconnection and connection of cables to each excited feed antenna. To address this issue, we performed a software calibration in which we added a phase offset to each measured lens pattern to align the phases in the direction of maximum radiation. After applying these phase offsets, we added the phase shifts from the analytical array factor formula to individual patterns measured from each lens to generate the radiation pattern.

Our far-field measurement capability limits us only to take 2D radiation pattern measurements. Hence, several azimuth cuts are needed to validate the design process, simulated radiation pattern, and realized gain performance. Four different physical configurations of the lenses are measured. These configurations correspond to the various feed antennas’ excitation and are indicated in Fig. 6. The considered feed antennas are the center element (with accompanying cross-polarization measurement) (A), the elements furthest from the lens center in the E and H planes (B and C), and the element located on the 45° diagonal (D). We used a different stand for each lens configuration to properly rotate the antenna to measure the azimuth cut that contains the direction of maximum radiation.

Fig. 12 depicts the broadside E-Plane radiation performance of the antenna along with its cross-polarization. The measured and simulated realized gains match closely up to ±50°. There is a ≈5 dB difference in the side lobe patterns corresponding to the direction of the horizon (90°). This discrepancy is generally seen in the other radiation pattern measurements as well. This difference may be due to the large ground plane introduced by the feeding network and the 3D printed test stand or the phase errors introduced via
Fig. 13. LAS-based antenna beam steering performance normalized to the broadside radiation. Solid lines: Beam is steered in elevation by ±5° away from the direction of maximum gain of the excited patch. Dashed lines: Beam is steered in elevation by ±5° away from the direction of the (solid line) maximum.

Fig. 14. Simulated and measured radiation patterns for feed antenna excitations B and C. The antennas provide beam steering in φ = 90° and φ = 0°, respectively, with maximum radiation towards identical elevation angle but with different SLL performance. Solid lines: measured, dashed: simulated.

vibrations in the rotating platform. Nevertheless, despite this SLL increase, the SLL level is still −15.4 dB and well below the −9 dB specification used in the design. The realized gain is 19.4 dBi and matches well to the simulated realized gain of 19.8 dBi. The measured broadside cross-polarization is 16.2 dB lower than the co-polarization. Due to several potential contributions, this measurement is larger than the simulated value of 42.7 dB. One contribution is the presence of the 3D-printed holder adding a potential scatterer. Another is the polarization misalignment between the reference horn and antenna under test. In addition, typical standard gain horn antennas can exhibit cross-polarization levels comparable to our measured value, which would limit the dynamic range.

The radiation patterns from the off-axis excitations are shown in Fig.13. The solid lines represent the radiation patterns steered in the direction of maximum radiation of the excited feed antenna. The dashed lines represent steering the solid line patterns by ±5° (i.e., within the HPBW of the single lens’ beam). The solid traces correspond well with simulated results. The direction of the maximums agrees as patch D has a maximum at 18° and patch B has a maximum at 27°. This measurement maintains the SLL performance at −10.4 dB. The dashed traces also agree well with simulated performance. Steering within the HPBW of the associated single-lens beam drops the traces in power by 3 dB as expected. The outer beam (from patch B) is steered to ±34° in Fig.13. This beam can be steered further to ±36° while not exceeding an SLL of −9.5 dB. These results match the expectation of a ±36° scan range from full-wave simulations.

To observe pattern differences in the E and H-plane scans, the two patches in the φ = 0° and φ = 90° positions were measured and compared in Fig.14. These patches correspond with patch B and C in Fig.6. Both patches steer to ±30° with varying sidelobe behaviors. The E-plane cut has a higher SLL of −10.5 dB while the H-plane has a lower SLL of −12.7 dB. However, as both SLLs are under −9 dB, both feed antennas are equally suited to be used within the LAS-based antenna context. The measured data matches reasonably well with the simulated data, with noteworthy features well represented.

The measurements presented demonstrate the challenging cases of the LAS-based antenna in steering far off-axis. While these measurements differ slightly from simulated results, the general behavior is consistent. The broadside behavior matches very well, while off-axis behavior approaches expected performance. Therefore, the measurements show success in demonstrating this antenna’s ability to steer to ±34° covering ±36° while maintaining a low SLL.

IV. CONCLUDING REMARKS

In this work, we introduced a 2D LAS-based antenna using EHDLS. To facilitate the design, we presented a systematic approach to achieving good performance in scan range and side lobe level. The concept is demonstrated with an antenna consisting of L = 7 LAS and M = 17 feed antennas per LAS. The antenna operates at 38 GHz with a realized gain of 19.8 dBi and worst case SLL of −9.5 dB while exhibiting a coverage of ±36°. The measured performance corresponds well with simulation-based predictions. Table 1 outlines the performance of the presented LAS-based antenna within the context of subarray-based antennas reported to date. The proposed design operates at the 38 GHz band, provides 2D beam-steering capability, and exhibits the highest beam-steering range with nearly −10 dB SLL performance. Additionally, this design remains low profile through small, efficient lenses, allowing for incorporation in large, spectrum-efficient phased arrays.

Table 1. Beam steering antennas with reduced phase shifter count.

| Ref | Tech.       | N   | M   | Freq (GHz) | # of PS | Scan Range | SLL (dB) | HPBW |
|-----|-------------|-----|-----|------------|---------|------------|----------|------|
| [10] | Random      | 30  | 1/A | 7.9        | 12      | ±14°       | -15°     | -4.1 |
| [9]  | Interlaced  | 16  | 5   | 7.9        | 10      | ±10°       | -19°     | -7   |
| [30] | Vector Sum  | 8   | 1   | 12.6       | 28      | ±8° VGAS   | ±18.5°   | -9   | 14.8 |
| [12] | 1D LAS (EHDLs) | 20  | 1   | 38        | 3       | ±37.5°     | ±4°      | 5.2  |
| [16] | 1D LAS (Metasurface) | 5   | 1   | 10        | 3       | ±30°       | -3°      | -7   |
| This Work | 2D LAS (EHDLs) | 17  | 7   | 38        | 7       | ±36°       | ±3°      | 5.5  |
REFERENCES

[1] M. Shafi, F. A. Molisch, J. P. Smith, T. Haustein, P. Zhu, P. D. Silva, F. Tufvesson, A. Benjebbour, and G. Wunder, “5G: A tutorial overview of standards, trials, challenges, deployment, and practice,” IEEE J. Sel. Areas Commun., vol. 35, no. 6, pp. 1201–1221, Jun. 2017.

[2] W. Roh, J. Y. Seol, J. Park, B. Lee, J. Lee, Y. Kim, J. Cho, K. Cheun, and F. Aryanfar, “Millimeter-wave beamforming as an enabling technology for 5G Tular communications: Theoretical feasibility and prototype results,” IEEE Commun. Mag., vol. 52, no. 2, pp. 106–113, Feb. 2014.

[3] S. Han, I. Chih-Lin, Z. Xu, and C. Rowell, “Large-scale antenna systems with hybrid analog and digital beamforming for millimeter wave 5G,” IEEE Commun. Mag., vol. 53, no. 1, pp. 186–194, Jan. 2015.

[4] R. L. G. Cavalcante, S. Staniczak, M. Schubert, A. Eisenblatter, and U. Tuerke, “Toward energy-efficient 5G wireless communications technologies: Tools for decoupling the scale of networks from the growth of operating power,” IEEE Signal Process. Mag., vol. 31, no. 6, pp. 24–34, Nov. 2014.

[5] R. Méndez-Rial, C. Rusu, N. González-Prelcic, A. Alkhateeb, and R. W. Heath, “Hybrid MIMO architectures for millimeter wave communications: Phase shifters or switches?” IEEE Access, vol. 4, pp. 247–267, 2016.

[6] C. Liu, Q. Li, Y. Li, X.-D. Deng, H. Tang, R. Wang, H. Liu, and Y.-Z. Xiong, “A Ka-band single-chip SiGe BiCMOS phased-array transmit/receive front-end,” IEEE Trans. Microw. Theory Techn., vol. 64, no. 11, pp. 3667–3677, Nov. 2016.

[7] J. Nemt, “Network approach for reducing the number of phase shifters in a limited scan phased array,” U.S. Patent 3 803 625, Apr. 9, 1974.

[8] J. R. Mailloux and R. P. Franchi, “Reducing the number of phase shifters in limited-scan arrays,” IEEE Trans. Antennas Propag., vol. 68, no. 1, pp. 70–80, Jan. 2020.

[9] F. Akbar and A. Mortazawi, “Scalable phased array architectures with a reduced number of tunable phase shifters,” IEEE Trans. Microw. Theory Techn., vol. 65, no. 9, pp. 3428–3434, Sep. 2017.

[10] W.-S. Lee, S.-T. Kang, K.-S. Oh, and J.-W. Yu, “Design methodology for phased subarray antennas with optimized element phase control,” in Proc. Eur. Microw. Conf., Oct. 2013, pp. 347–350.

[11] R. L. Haupt, “Reducing grating lobes due to subarray amplitude tapering,” IEEE Trans. Antennas Propag., vol. 64, no. 11, pp. 4648–4658, Nov. 2016.

[12] B. Rupakula, A. H. Aljuhani, and G. M. Rebeiz, “Limited scan-angle phased arrays using randomly grouped subarrays and reduced number of phase shifters,” IEEE Trans. Antennas Propag., vol. 68, no. 1, pp. 70–80, Jan. 2020.

[13] G. Mucmu, M. Karac, and J. Mendoza, “Mm-wave beam steering antenna with reduced hardware complexity using lens antenna subarrays,” IEEE Antennas Wireless Propag. Lett., vol. 17, no. 9, pp. 1603–1607, Sep. 2018.

[14] M. Karabacak, H. Arslan, and G. Mumcu, “Lens antenna subarrays in mmWave hybrid MIMO systems,” IEEE Access, vol. 8, pp. 216634–216644, 2020.

[15] M. Karabacak, G. Mumcu, and H. Arslan, “Hybrid MIMO architecture using lens arrays,” U.S. Patent 10 714 836, Jul. 14, 2020.

[16] K. A. Shila and G. Mumcu, “Mm-wave beam steering antenna based on extended hemispherical lens antenna subarrays,” in Proc. IEEE Int. Symp. Antennas Propag. North Amer. Radio Sci. Meeting, Montreal, QC, Canada, Jul. 2020, pp. 1517–1518.

[17] A. Artemenko, A. Mozharskiy, A. Malshev, R. Maslennikov, A. Sevastyanov, and V. Sosrin, “Experimental characterization of E-band two-dimensional electronically beam-steerable integrated lens antennas,” IEEE Antennas Wireless Propag. Lett., vol. 12, pp. 1188–1191, 2013.

[18] D. F. Filipovic, S. S. Gearhart, and G. M. Rebeiz, “Double-slot antennas on extended hemispherical and elliptical silicon dielectric lenses,” IEEE Trans. Microw. Theory Techn., vol. 41, no. 10, pp. 1738–1749, Oct. 1993.

[19] D. F. Filipovic, G. P. Gauthier, S. Raman, and G. M. Rebeiz, “Off-axis properties of silicon and quartz dielectric lens antennas,” IEEE Trans. Antennas Propag., vol. 45, no. 5, pp. 766–768, May 1997.

[20] M. Kacar, T. M. Weller, and G. Mumcu, “3D printed wideband multilayered dual-polarized stacked patch antenna with integrated MMIC switch,” IEEE Open J. Antennas Propag., vol. 2, pp. 38–48, 2021.

[21] G. Godi, R. Saulou, and D. Thouroude, “Performance of reduced size substrate lens antennas for millimeter-wave communications,” IEEE Trans. Antennas Propag., vol. 53, no. 4, pp. 1278–1286, Apr. 2005.

[22] R. J. Fowler, M. S. Paterson, and S. L. Tanimoto, “Optimal packing and covering in the plane are NP-complete,” Inf. Process. Lett., vol. 12, no. 3, pp. 133–137, 1981.

[23] C. T. Zahn, “Black box maximization of circular coverage,” J. Res. Nat. Bur. Standards B, vol. 66, pp. 181–216, Aug. 1962.

[24] Z.-H. Zhan, J. Zhang, Y. Li, and H.-S.-H. Chung, “Adaptive particle swarm optimization,” IEEE Trans. Syst., Man, Cybern., B, Cybern., vol. 39, no. 6, pp. 1362–1381, Dec. 2009.

[25] H. Frid, “Closed-form relation between the scan angle and feed position for extended hemispherical lenses based on ray tracing,” IEEE Antennas Wireless Propag. Lett., vol. 15, pp. 1963–1966, 2016.

[26] J. Beznason, A. Edelman, S. Karpinski, and V. B. Shah, “Julia: A fresh approach to numerical computing,” SIAM Rev., vol. 59, no. 1, pp. 65–98, 2017.

[27] M. Khalil, R. Tafazolli, T. A. Rahman, and M. R. Kamarudin, “Design of phased arrays of series-fed patch antennas with reduced number of the controllers for 28-GHz mm-wave applications,” IEEE Antennas Wireless Propag. Lett., vol. 15, pp. 1305–1308, 2016.

[28] R. Xu and Z. N. Chen, “A compact beamsteering metamaterial lens array antenna with low-cost phased array,” IEEE Antennas Propag. Lett., vol. 69, no. 4, pp. 1992–2002, Apr. 2021.

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