IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION, VOL. 70, NO. 5, MAY 2022

High-Gain Lens-Horn Antennas for Energy-Efficient 5G Millimeter-Wave Communication Infrastructure

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Abstract—Lower efficiencies of power amplifiers and increased path losses at higher frequencies are two reasons why utilizing millimeter-wave frequencies for future wireless communications is challenging. In this article, a high-gain multilens-horn antenna system is presented. The antenna system provides highly effective isotropic radiated power by using only a few high-gain antenna elements, where each element only needs a low transmit power of 26 dBm to reach users at far distances. Moreover, due to channel reciprocity, both down- and uplink benefit from an increased antenna gain. It is shown that a reasonable number of high-gain antennas can provide coverage in a 60° sector with an inter-site distance of 300 m. Based on radio-planning simulations, three horn shapes are identified to provide sufficient coverage. The final optimized dual-polarized multilens-horn design is fabricated and experimentally verified. With a low transmit power at the base station (BS) of 26 dBm, a signal strength of at least −65 dBm is achieved in 98.9% of the sector at 28 GHz under non-line-of-sight conditions for the vertical polarization. For the horizontal polarization, an area of 96.2% is covered.

Index Terms—5G mobile communication, base stations (BSs), directive antennas, millimeter wave communication, multiple-input-multiple-output (MIMO), single-input-single-output (SISO).

I. INTRODUCTION

The intention of exploiting higher frequencies for mobile communications is driven by the steadily growing demand for higher data rates and the rising number of mobile communications devices. Despite the potentials, utilizing frequencies above 6 GHz comes with new challenges. It is stated in [1] and [2] that the propagation characteristics at mmWave frequencies differ from the propagation characteristics at lower frequencies. For instance, radio propagation relies more on the line-of-sight (LOS) path and strong reflections and higher penetration losses are observed. These challenges make it more difficult to reach users at far distances, where it is likely to have non-LOS (NLOS) communications. Hence, to reach users at far distances, a high effective isotropic radiated power (EIRP) is needed, which is the transmit power supplied to the antenna, $P_{TX}$, times the antenna gain, $G$. Therefore, two variables can be tuned.

The transmit power can be increased by using high-power amplifiers (HPA). However, the design of high-efficiency linear HPAs becomes more challenging with increasing frequency [3]. This makes the approach of only increasing $P_{TX}$ less desirable. Therefore, it is necessary to increase the antenna gain as well.

This can be achieved, for example, if multiple antenna elements are combined and by using high gain antenna elements. The first case describes the multiple-input-multiple-output (MIMO) approach, which provides high flexibility if realized fully digital and shows a high potential for future wireless systems if realized on a larger scale, so-called massive MIMO. However, as described in [4], one of the challenges of fully digital massive MIMO systems is the large number of radio frequency (RF) chains and the hardware needed for those. Therefore, to lower the hardware complexity, hybrid massive MIMO systems that are realized with analog beamforming subarrays are considered in [5]. The use of subarrays can significantly reduce the number of analog-, digital-, up- and down-converters at the expense of the need for analog beamformers in each subarray.

Driven by the goal to use high gain antenna elements instead of active subarrays, we have studied in [6] the impact of the antenna element directivity in a fixed-beam scenario. Based on the outcome, it is observed that high-gain antenna elements used to partition the mobile communication cell can outperform low-gain antenna elements in single-input-single-output (SISO) and MIMO mode. The notion of fixed-beam sector partitioning can be seen as an extension of the cell-sectoring [7], which is currently used in mobile communications. For the sector partitioning, multiple beams are generated by high-gain antenna elements, where each is designed to provide a particular beam shape and antenna gain to cover a specific region of the cell sector. A similar approach is a switched-beam system, e.g. [8], where the antenna system can switch between predefined beams.

Based on the previous studies, this article proposes to replace subarrays with high-gain antenna elements and to extend the currently used cell-sectoring. To do so, we study the use of lens-horn antenna elements in an antenna system for millimeter-wave wireless communications using a system-level design and verification methodology. The following bullet points summarize the outcome of the study:

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1) a detailed design of an energy-efficient multilens-horn (MLH) antenna system with cell partitioning and fixed-beams for reduced intracell interference:
   a) using dual-polarized compact lens-horn antennas designed for the n257 band (26.5–29.5 GHz, [9]);
   b) with low transmission power requirement per antenna element of 26 dBm for energy-efficient mobile communications link;
2) experimental system-level verification with a complete base station (BS) antenna that covers a 60° sector.

As a whole, this work’s novel contribution is the investigation of the extension of the currently used cell-sectoring in mobile communications by using highly directional, fixed-beam antennas to utilize higher frequencies while reducing the required transmit power. The work shows that such an antenna system is feasible by giving an example of a high-gain lens-horn antenna system underlaid with system simulation and measurement results. Section II presents the considered scenario, including the system requirements. For a generic approach, the fixed-beam solution is designed for a unit sector, which can be used to provide coverage to a 360° multisite scenario shown in Fig. 1. The bottom part of Fig. 1 indicates that the element beams of the fixed-beam approach overlap, which is unavoidable if full coverage is desired. This beam overlap can be beneficial, as shown in the performance study [10], where we have put the developed antenna system in a multiuser scenario to the test. The study’s outcome shows that the high-gain antenna elements used for partitioning the area can achieve good performances in SISO and MIMO mode, and although the beam overlaps are rather small, we have observed a benefit of using digital beamforming for an improved user separation. Section III shows the initial coverage performance using Gaussian beams and presenting the final antenna design. In Section IV, the antenna simulation is presented together with the measurement for the vertical and horizontal polarization. The final radiation characteristics are fed back to a system simulator to show the achieved coverage for the vertical and horizontal polarization at various frequencies.

II. System-Level Requirements

A. Micro-Cell Scenario

The considered scenario parameters in this study are based on an urban micro cell and are listed in Table I and visualized in Fig. 1. There are six 60° sectors, each covered by a BS antenna system. BSs and UEs are located at a height of 10 and 1.5 m, respectively. As it is shown in the bottom part of Fig. 1, the goal is to provide a high-gain antenna system that is capable of providing simultaneously generated beams to cover the sector. Each beam is generated by a single antenna element, which has its own RF chain. The utilized frequency spectrum is the n257 band (26.5–29.5 GHz) with the center frequency at 28 GHz. To obtain an estimate of the received power levels under realistic conditions, the 3GPP TR 38.901 UMi street canyon path loss model is applied under NLOS conditions. The applied model can be found in [12, Table VII.4.1-1].

B. Link Budget Estimation

To establish a link between BS and UE at a specific data-rate, the received power must be equal or larger than the receiver sensitivity (RS). In [13], the RS limit is defined as

$$P_{\text{RX,sens}} = -174 \frac{\text{dBm}}{\text{Hz}} + 10 \log_{10} B + NF_{\text{sys}}$$

$$+ 10 \log_{10} \left( \frac{E_b}{N_0} \frac{R_{\text{net}}}{B} \right)$$

where $-174$ (dBm/Hz) is the thermal noise at 300 K, $NF_{\text{sys}}$ is the system noise figure, $B$ is the applied channel bandwidth, $E_b$ is the bit energy, $N_0$ is the noise power density within the channel bandwidth and $R_{\text{net}}$ is the net bit-rate. Note that the SNR per bit, $(E_b/N_0)$, depends on the applied modulation scheme and the accepted symbol error rate, where the relation is displayed in Fig. 2.

Table I

| Description                  | Parameters |
|------------------------------|------------|
| Inter-site distance (ISD)    | 300 m      |
| Sector opening               | 60°        |
| $h_{\text{BS}}$              | 10 m       |
| $h_{\text{UE}}$              | 1.5 m      |
| $f_c$                        | 28 GHz     |
| Channel bandwidth, $B$       | 400 MHz$^1$|
| Transmit power, $P_{\text{TX}}$ | 26 dBm$^2$ |

1Considered n257 band channel bandwidth: 50 MHz, 100 MHz, 200 MHz, and 400 MHz [9].

2Power limit is equal to the maximum total radiation power of the UE power class 2 defined in [9] for mmWave communications.
The theoretical symbol error rate versus SNR per bit based on [11].

**TABLE II**

| Net bit rate | Downlink | Uplink |
|--------------|----------|--------|
| $B$          | 0.8 Gbit/s | 0.4 Gbit/s |
| $R_{B}$      | 400 MHz   | 200 MHz |
| $N_{0}$      | 14 dB     | 14 dB  |
| $N_{sys}$    | 5 dB      | 9 dB   |
| $P_{RX,sens}$| $-65$ dBm | $-72$ dBm |

The applied values and the calculated $P_{RX,sens}$ are listed in Table II for downlink (DL) and uplink (UL). Note that the different data-rates are due to DL to UL ratio of 2:1 [14]. With these settings, the corresponding data-rates, summarized in Table II, for DL and UL are higher than the experienced data-rate (defined in [14]) requirements for scenarios such as urban macro and dense urban in [15].

For this study, we are assuming channel reciprocity and aim to provide a received power of $-65$ dBm for both DL and UL. For DL, this achieves the calculated data rate of 0.8 Gbit/s while for UL this power level results in a margin of 7 dB to cope, for example, with polarization mismatch. By enforcing this symmetry, the transmit power levels in DL and UL have to be equal. Hence, the maximum UE TX power of a mmWave UE power class 2, 26 dBm, as defined in [9], is used in this study as the maximum transmit power level. This, in turn, reduces the overall power consumption of the BS. Furthermore, to obtain a total link budget and calculate the required EIRP of the BS system, a UE antenna gain of 11 dBi is assumed. This UE antenna gain aligns with the current state-of-the-art UE chip-sets, like the Qualcomm QTM525 module [16]. Therefore, it is ensured that the received power level would be achieved in DL and UL due to channel reciprocity of TDD systems.

By taking all these considerations into account, the minimum EIRP within the half-power beamwidth (HPBW) and the peak gain are calculated to receive a sufficiently strong signal within the HPBW. The calculated values are listed in Table III. The calculation shows that the highest needed BS antenna gain is 31 dBi.

**C. Physical Constraints on the System**

By using an aperture antenna, e.g., a horn antenna, a large antenna gain can be realized by increasing its aperture. The dimension of a horn is directly linked to the antenna directivity by [17]

$$D_{dir, dB} = 10 \log_{10} \left( \frac{\epsilon_{ap} \lambda^2}{4 \pi A} \right) \quad (2)$$

where $\epsilon_{ap}$ is the aperture efficiency, $A$ is the physical aperture and $\lambda$ is the free-space wavelength.

With growing aperture size, the horn length has to grow to avoid large phase deviations over the electromagnetic field distribution in the aperture. This phase deviation is caused by the path difference between the feed to aperture center and the feed to aperture edge. Given by [17], the path difference is reduced sufficient enough for a horn element length of

$$l \approx \frac{D^2}{3 \lambda} \quad (3)$$

where $D$ is the diameter of the aperture. Hence, higher directivity comes at the cost of larger overall antenna dimensions. Based on the estimated link budget, derived in Section II-B, the largest aperture diameter is estimated to be $D \approx 0.18$ m assuming $\epsilon_{ap} = 1$. In this case, the horn length would amount to $l \approx 1$ m, resulting in a rather large overall system. Hence, the aim is to keep the aperture efficiency high and, therefore, $D$ at its smallest. Furthermore, a maximum horn length of $l = 0.2$ m has been chosen based on mechanical considerations. The resulting phase deviation over the aperture then has to be compensated by the antenna design.

**III. Antenna Design**

**A. Sector Coverage**

The goal is to provide sufficient coverage so that the area of interest is covered with $-65$ dBm received power, as discussed in Section II-B.

Based on this goal, the HPBW and orientation of each antenna are computed for the scenario, given in Section II-A, using Gaussian beams. In the next step, the orientation and beam shape of the Gaussian beams are optimized in a MATLAB environment to provide sufficient coverage. At this stage, we did not consider pattern variation due to polarization, while we do consider the polarization variations later on in the study. For the initial search, a single Gaussian beam is directed toward the center of the defined scenario. The beam
TABLE IV
BEAM CONFIGURATION AT 28 GHz FOR IDEAL GAUSSIAN BEAMS

| #  | Antenna gain [dBi] | Beamwidth (Hor./Vert.) [°] | Orientation (Az/θz) [°] | Lens-horn pos. (x,y) |
|----|-------------------|-----------------------------|-------------------------|---------------------|
| 1  | ~ 16.5            | (28.6,27.7)                 | (0.31)                  | (~14.1,1.4,−10.4)  |
| 2  | ~ 24.6            | (14.9,6.1)                  | (15.8)                  | (~0.6,−19.2,−1.0)  |
| 3  | ~ 24.6            | (14.9,6.1)                  | (−15.8)                 | (~0.6,19.2,−1.0)   |
| 4  | ~ 31.8            | (5.5,4)                     | (0.2,5)                 | (0,0)              |
| 5  | ~ 31.8            | (5.5,4)                     | (−4.6,2.5)              | (0.9,3.3,48.6)     |
| 6  | ~ 31.8            | (5.5,4)                     | (−9.2,2.5)              | (~0.1,9.2,31.8)    |
| 7  | ~ 31.8            | (5.5,4)                     | (13.8,2.5)              | (~0.3,9.1,1.15)    |
| 8  | ~ 31.8            | (5.5,4)                     | (−18.4,2.5)             | (~2.3,30.6,31.9)   |
| 9  | ~ 31.8            | (5.5,4)                     | (−23.2,5)               | (~2.6,30.6,31.9)   |
| 10 | ~ 31.8            | (5.5,4)                     | (27.6,2.5)              | (~2.9,30.6,31.9)   |
| 11 | ~ 31.8            | (5.5,4)                     | (46.2,5)                | (~0.3,9.3,38.6)    |
| 12 | ~ 31.8            | (5.5,4)                     | (9.2,2.5)               | (~0.1,−9.2,31.8)   |
| 13 | ~ 31.8            | (5.5,4)                     | (13.8,2.5)              | (~0.3,−9.1,1.15)   |
| 14 | ~ 31.8            | (5.5,4)                     | (18.4,2.5)              | (~2.3,−30.9,48.7)  |
| 15 | ~ 31.8            | (5.5,4)                     | (23.2,5)                | (~2.6,−30.6,31.9)  |
| 16 | ~ 31.8            | (5.5,4)                     | (27.6,2.5)              | (~2.9,−30.3,15.1)  |

Fig. 3. Considered scenario showing downtilt angle, A_{dt}, BS height, and user equipment height definitions at a distance d.

is generated with respect to the estimated antenna gain and is rotationally symmetric. The algorithm modifies the HPBW and the beam orientation to maximize the area covered with sufficient high received signal power while keeping the signal outside the defined cell low. After the algorithm has converged, a second beam is generated and directed toward the point with the lowest received power within the cell. Like the first beam, the second beam is generated with respect to the estimated antenna gain and is rotationally symmetric. The HPBW and orientation are then modified of the second beam to provide the best coverage with both beams present while keeping the received power outside the cell low. This is done for all 16 Gaussian beams. The final configurations are listed in Table IV. The table presents the beamwidth of each antenna element in the horizontal and vertical plane. The orientation of the lens-horn antennas is given by an azimuth and downtilt angle, A_{dt}. The downtilt angle is given with respect to the horizon, shown in Fig. 3, where a positive angle describes a down tilt. The azimuth angle is measured clockwise from +x'-direction, shown in Fig. 4. The relative lens-horn position is given in terms of λ with the reference point at the element #4, which is at the BS height of 9.8 m. The contour plot, Fig. 4, shows the area that can be covered by the 16 Gaussian beams that are required for this multiantenna system, where the received power is at least −65 dBm. Note that polarization mismatch loss at the receiver between the incoming signal and the receiver antenna is not taken into account. In this case, 99.7% of the sector can be reached with a received signal strength of above −65 dBm.

B. Maximum Received Power for Array Configuration

The coverage planning in Section III-A presents coverage of 99.7% in which at least data rates of 0.8 Gbit/s with a SISO communication link between a single BS antenna element and a UE can be provided. However, besides SISO mode, we can also make use of multiple BS antenna elements simultaneously if each horn antenna element has its own RF path and analog/digital converter (ADC/DAC) as shown in Fig. 1. In this case, the multihorn system can be operated to utilize multiple channels in an MIMO fashion, similar to a (digital) phased array.

For instance, to maximize the received power, maximum ratio transmission (MRT) on the transmitter side or maximum ratio combining (MRC) on the receiver side can be applied. In each case, weights are applied based on the channel state information to maximize the power at the receiver. The theoretical maximum is shown in the heat map in Fig. 5 under the assumption that the PAs operating at their maximum output power, where the signals are coherently accumulated for every single point. In this case 99.9% of the area receives a power level of at least −65 dBm. As the heat map reveals, theoretically received powers of 10-15 dB above the set threshold can be achieved. This means in phased array mode, higher modulation schemes could be used to increase spectral efficiency without increasing the transmit power per antenna element.

C. Antenna Element Design

In the next step, the antenna elements are designed based on the obtained values listed in Table IV. The Gaussian beam study estimated that there are three different horn elements needed with antenna gains of 16.5 dBi (small), 24.6 dBi (mid) and 31.8 dBi (large). To achieve the required antenna gain of each element and fulfill the physical constraint on the system,
the high gain antenna elements have to be much shorter than their ideal length approximated by (3), which causes a phase deviation. The maximum phase deviation of a conical horn is given by [17] as

\[ \delta = \frac{D^2}{8l} \]

where \( \delta \) expresses the deviation in wavelength. To compensate the phase deviation, a lens can be used. In the following, the design process for the lens-horn antennas is discussed.

The initial design for each lens-horn is based on the defined constraints in Section II-C. The antenna is then further optimized by adjusting the lens in terms of material properties and curvature with respect to the horn element. For correcting the phase error, it is decided to use a plano-convex curved lens, which can be directly attached to the horn opening. Theoretically, any dielectric material can be used for the lens antennas. By using higher permittivity materials, the lens becomes thinner, and therefore the weight of the lens is reduced. Furthermore, a thinner lens means that there is less lossy material through which the signal propagates. On the other hand, the loss tangent, \( \tan \delta \), increases and stronger surface reflections are seen with increasing permittivity. Hence, selecting the permittivity is a trade-off, depending on the application and needs of the antenna system. Besides the electrical characteristics, also good UV resistance and low water absorption are important. These weather resistance abilities have the advantage that the lens can also serve as a radome to protect the inside of the horn from water, insects, and dirt. A suitable material is polytetrafluoroethylene (PTFE), also known as teflon. However, one drawback of PTFE is its high density of 2.2 (g/cm\(^3\)) resulting in high weight for a standard lens design in this case. For weight reduction, steps can be introduced to the lens design as it is presented in [18] and [19]. This lens approach is called a Fresnel lens. Furthermore, to reduce the leverage effect of the lens and maintain the center of gravity within the horn, the convex side of the lens is pointing inward to the feed.

For the lens, the algorithm described in [18] and [19] is applied using the parameters given in Table V for the large, middle, and small lens. For the calculation, only a single step inside the lens is considered. In general, introducing steps reduces the lens weight. However, at the same time, the lens bandwidth decreases. In [19], the \( x_s \)-coordinate of the surface of a single stepped lens is expressed as a function of the \( y_s \)-coordinate by

\[ x_s(y) = \frac{L}{1 + \sqrt{\varepsilon_r}} \left[ \frac{1 + \left( \frac{y_s}{L \sqrt{\frac{\gamma^2 - 1}{\sqrt{\varepsilon_r} + 1}}} \right)^2}{1 + \sqrt{\varepsilon_r}} \right] + L \frac{\sqrt{\varepsilon_r}}{1 + \sqrt{\varepsilon_r}} + \frac{\lambda}{\sqrt{\varepsilon_r} - 1} \]

where \( \varepsilon_r \) is the relative permittivity of the lens and \( L \) is expressed as

\[ L = \frac{F \left( \sec (\gamma) - \sqrt{\varepsilon_r} \right)}{1 - \sqrt{\varepsilon_r}} + \frac{\lambda}{\sqrt{\varepsilon_r} - 1} \]

where \( F \) is the aperture radius and \( \gamma \) is the half of the horn flare angle. First, by applying (5), the 2-D curvature of the lens is generated, and then the 3-D lens is obtained by rotating the curvature. Note that the final conical horns have oval cross sections. Therefore, the calculated circular lenses are fit to the horn openings. The dimensions of the horn antennas are presented in Figs. 6–8. The simulation results are discussed together with the measurement results in Section IV.

### IV. Prototyping and Testing

The final 16-element MLH-system consisting of the three different antenna types (small, mid, large) is depicted in Fig. 9. It shows the manufactured MLH-antennas mounted on a multihorn holder in the anechoic chamber. The design of the multihorn holder allows adjusting each lens-horn element individually so that they can be set up with respect to orientation and position listed in Table IV. In the next step, the lens-horn antennas are characterized in terms of the radiation pattern, antenna gain, and mutual coupling, which is summarized in Section IV-A for the vertical and horizontal polarization. Since the coax-to-waveguide adapters are single polarized, the adapters are rotated to measure the antennas’ vertical and horizontal responses. The measured information is then used to update the coverage simulation from Section III-A. The results of this updated simulation are provided in Section IV-B for the vertical and horizontal polarization.

### Table V

| Material   | Aperture radius | Flare angle | \( \lambda \) | Focal length |
|------------|-----------------|-------------|---------------|--------------|
| large PTFE \((\varepsilon_r = 2.1)\) | 80 mm          | 26.56\(^\circ\) | 10.71         | 160 mm       |
| mid PTFE \((\varepsilon_r = 2.1)\)  | 81 mm          | 26.85\(^\circ\) | 10.71         | 159 mm       |
| small PTFE \((\varepsilon_r = 2.1)\) | 14.5 mm        | 16.3\(^\circ\)  | 10.71         | 48 mm        |
A. Measured Horn Antenna Characteristics

For the measurement, the NSI planar near field scanner in the anechoic chamber of the Eindhoven University of Technology is used to measure the antenna characteristics of the antenna under test (AUT). The measurements are performed for each antenna element, and a standard gain...
horn (SGH) is used as a reference to determine the antenna gain of the constructed lens-horn antennas. Note that each lens-horn antenna is characterized with its coax-to-waveguide adapter since an SMA 2.92 mm calibration kit is used for calibration. In Figs. 10–12 the radiation pattern for the three different antennas are compared to their simulations for the vertical polarization and in Figs. 13–15 for the horizontal polarization. It can be seen that the simulation and measurement show a good agreement. Hence, the presence of the multihorn holder does not significantly affect the radiation characteristics of the measured antennas. Besides the measured co-polarization, also the cross-polarization is plotted. The cross-polarization ratio at 0° is at least 25 dB, which is in alignment with currently used BS antenna systems [20].

To investigate the influence of the adjacent elements in this multiantenna configuration, the coupling between elements is measured when placing them horizontally side-by-side and pointing in the same direction. This setup is similar to the final setup, shown in Fig. 9, where the elements are side-by-side but pointing slightly away from each other. The measurement is performed for large-large, mid-large, and small-large constellation for both polarization separately. The results are plotted in Fig. 16, where the top and bottom graph shows the vertical and horizontal polarization, respectively. Note that the $S_{11}$ refers to the first lens-horn of each pair. For both vertical and horizontal polarization, a low coupling of at least $-58$ dB is observed for all constellations, which is expected since the radiation pattern measurement where all elements are present showed a neglected variation in the radiation pattern. Furthermore, it is found that the lens introduces reflections in the antenna structure. The reflection position can be determined by the ripple distance in the simulated and measured $S$-parameter.

Fig. 10. Large lens-horn (vert. pol.): Simulation versus measurement of the vertically polarized antenna displayed in vertical (top) and horizontal (bottom) cut, at 28 GHz. For the measurement both, co- and cross-polarization, are presented.

Fig. 11. Mid lens-horn (vert. pol.): Simulation versus measurement of the vertically polarized antenna displayed in vertical (top) and horizontal (bottom) cut, at 28 GHz. For the measurement both, co- and cross-polarization, are presented.

Fig. 12. Small lens-horn (vert. pol.): Simulation versus measurement of the vertically polarized antenna displayed in vertical (top) and horizontal (bottom) cut, at 28 GHz. For the measurement both, co- and cross-polarization, are presented.
Fig. 13. Large lens-horn (hori. pol.): Simulation versus measurement of the horizontally polarized antenna displayed in vertical (top) and horizontal (bottom) cut, at 28 GHz. For the measurement both co- and cross-polarization are presented.

Fig. 14. Mid lens-horn (hori. pol.): Simulation versus measurement of the horizontally polarized antenna displayed in vertical (top) and horizontal (bottom) cut, at 28 GHz. For the measurement both co- and cross-polarization are presented.

However, the reflection coefficient is below $-10\,\text{dB}$ for the desired frequency range, and only slight antenna gain variations due to ripples are observed.

Fig. 15. Small lens-horn (hori. pol.): Simulation versus measurement of the horizontally polarized antenna displayed in vertical (top) and horizontal (bottom) cut, at 28 GHz. For the measurement both co- and cross-polarization are presented.

Fig. 16. Two-port s-parameter response for two adjacent lens-horn antennas with the combinations of two large lens-horns, large and mid lens-horns, and large and small lens-horns. The top and bottom plot depicts the vertically and horizontally polarized responses, respectively.

Besides a low coupling, also a similar response over frequency is desired. For the antenna system, the considered frequency range is $26.5$–$29.5\,\text{GHz}$ (n257 band). Therefore,
the radiation patterns of the lens-horn antennas are plotted for 26.5, 28, and 29.5 GHz in Figs. 17–22 for the vertical and horizontal polarization. The plots show that the main beams have minor variations for the large and mid-lens-horn antenna across the frequency band, while there are some fluctuations noticeable in the sidelobes. The largest observed deviations of
the HPBW between the frequencies of the large, mid, and small lens-horn are about 0.5°, 1°, and 7.5°, respectively. The variation of the small lens-horn over frequency can be clearly seen in the horizontal plane in Fig. 19. For instance, at 22 m distance, where the center of the small lens-horn beam is pointing at, the beam footprint at 26.5 GHz increases in the y-direction by about 4 m in comparison to the footprint at 28 GHz. Hence, the large variation of 7.5° has only a minor effect. That is because variations in the beamwidth have a smaller impact on the beam footprint in the near region than at far distances. Slight variations are also found in the antenna gain of the small, mid, and large antenna, see Table VI. A variation over frequency is unavoidable due to the frequency dependencies in (2)–(4). A closer look at how much the beamwidth and gain variations over frequency influence the coverage is discussed for both, vertical and horizontal, polarizations, in Section IV-B.

### B. Optimized Coverage Using Obtained Radiation Characteristics

To verify the lens-horn antennas on system level, the obtained radiation patterns from Section IV-A are fed to the coverage model from Section III-A. Since the initial simulation used ideal Gaussian beams, the orientation of the lens-horn antennas has to be optimized in the simulation for the actual radiation patterns. For instance, with the actual radiation patterns, the center beams should be pointed further down. The updated coverage plots are shown in Figs. 23 and 24 for vertical and horizontal polarization, and the updated

#### TABLE VI

| Antenna Gain Measurement and Simulation Results (Vert. Pol.) |
|-------------------------------------------------------------|
| **Frequency** | 26.5 GHz | 28 GHz | 29.5 GHz |
| **Measured [dBi]** | 29.8 | 30.5 | 30.8 |
| **Simulated [dBi]** | 29.9 | 31 | 31.4 |

#### TABLE VII

| Antenna # | Orientation (Az., Aφ) [°] |
|-----------|--------------------------|
| 1         | (0, 20.7)                |
| 2         | (19, 8.5)                |
| 3         | (-19, 8.5)               |
| 4         | (0, 4.8)                 |
| 5         | (-4.2, 4.7)              |
| 6         | (-8.9, 4.2)              |
| 7         | (-13.9, 3)               |
| 8         | (-18.8, 3)               |
| 9         | (-23.5, 3)               |
| 10        | (-27.5, 3)               |
| 11        | (4.2, 4.7)               |
| 12        | (8.9, 4.2)               |
| 13        | (13.9, 3)                |
| 14        | (18.8, 3)                |
| 15        | (23.5, 3)                |
| 16        | (27.5, 3)                |

To verify the lens-horn antennas on system level, the obtained radiation patterns from Section IV-A are fed to the coverage model from Section III-A. Since the initial simulation used ideal Gaussian beams, the orientation of the lens-horn antennas has to be optimized in the simulation for the actual radiation patterns. For instance, with the actual radiation patterns, the center beams should be pointed further down. The updated coverage plots are shown in Figs. 23 and 24 for vertical and horizontal polarization, and the updated
orientation of the MLH-system are listed in Table VII. In the updated coverage plots, the three different frequency cases are displayed. With the updated orientation, the sector provides a received power of above $\sim65$ dBm in 98.9% (at 28 GHz) of the area. At 26.5 GHz 98.3% and at 29.5 GHz 98.5% of the area is covered. The slight variations in coverage show that the MLH-antenna system is robust against the variation over frequency. The coverage for the horizontal polarization MLH-antenna system is plotted in Fig. 24. Here, a received power of above $\sim65$ dBm is provided in 96.2% (at 28 GHz) of the area. At 26.5 GHz 96.2% and at 29.5 GHz 95% of the area is covered. The variation in coverage is due to differences in the beamwidths for each polarization, leading to a larger spread between the high-gain beams. In this case, the orientation could be adjusted to provide an even higher coverage in the horizontal polarization. However, this would cause more inter-site interference for the vertical polarization, and therefore it is always a trade-off between coverage and inter-site interference at the edge of a sector. Furthermore, in the design process, an NLOS path loss model is assumed, for which the coverage percentage is calculated. Hence, the achieved coverage of at least 95% in all considered cases is a pessimistic estimate, and it is likely to be higher.

In general, a lower received power level does not necessarily mean that no link can be established due to the safety margin and a low error rate that is assumed in this study, see Section II-B. However, in case the received power level is too low, the bandwidth can be reduced, or the signals of adjacent antenna elements can be combined, see Section III-B.

V. CONCLUSION

In this article, the use of a high-gain antenna system for fixed-beam coverage in mobile communications is proposed and analyzed. The proposed MLH-antenna system consists of three different lens-horn antenna designs and has a total dimension of 0.82, 0.76, 0.38 m (x, y, z). It is shown that the MLH-antenna system partitions with 16 antenna elements a 60° sector in a multisite scenario with an ISD of 300 m such that it provides a received power of $\sim65$ dBm in 98.9% (vert. pol.) and 96.2% (horiz. pol.) of the area at 28 GHz. It is demonstrated by coverage simulations that the MLH-antenna system is robust against variations in antenna gain and HPBW, which are seen for different frequencies and polarizations. The system is designed to provide a data rate of at least 0.8 Gbit/s (SISO mode, single-polarization) in most of the cases under NLOS conditions with a low TX power of 26 dBm. However, the system can also be used in MIMO mode to achieve higher data rates, which is studied for a general fixed-beam antenna solution in [6]. All in all, it is concluded that the proposed system can be realized and is a viable option for future mm-wave mobile communications.

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