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Wide-Scan Focal Plane Arrays for mmWave Point-to-Multipoint Communications

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ABSTRACT  A new antenna system concept is presented where a wide-scan focal plane array antenna is used to achieve point-to-multipoint communication. The focal plane of a parabolic toroid reflector is populated with several antenna arrays, the positions of which determine the directions of the beams. This concept is investigated for beams pointed towards 0° (broadside) and 28° in azimuth. Each array allows for scanning an additional ±1° in azimuth and elevation. This allows for compensation of twist and sway of the antenna mast. Several array configurations are compared in terms of directivity and scan loss for such a system at E-band. It is found that an 8-by-8 array with an inter-element spacing of 0.7λ results in an optimal directivity with a scan loss lower than 1dB when scanning ±1° in azimuth and elevation. For the 0° beam direction the directivity is 45.5dBi and for the 28° beam direction the directivity is 44.4dBi, showing the wide angle scanning properties of this system. An experimental system is built at K-band and measurements are performed showing this system in action. In the measurements an array of 8-by-8 is used with an inter-element spacing of 0.5λ. The scan loss when scanning ±2° in azimuth and elevation is below 1dB. The directivity is 37.0dBi and 35.4dBi for the 0° and 28° beam directions, respectively. The spillover losses and aperture efficiencies are also found, as well as a relative metric for the transmitted power and the effective isotropic radiated power for both the E-band and K-band systems.

INDEX TERMS  Focal plane array, point-to-multipoint communication, reflector antenna, mmWave, 5G, 6G.

I. INTRODUCTION

FUTURE fifth generation (5G) and sixth generation (6G) wireless communication systems promise data-rates in the order of gigabits per second (Gbit/s) per user for multiple users simultaneously [1]. To achieve such high data rates, new frequency bands in the millimeter-wave (mmWave) range are being allocated in order to take advantage of the large amount of available spectrum. Furthermore, the distance between cells is reduced in comparison to current networks, resulting in shorter link distances. Hence, mmWave E-band (71-86 GHz) wireless fronthaul point-to-multipoint (P2MP) systems are proposed [2]. High-gain beamforming antenna systems are required to generate a high effective isotropic radiated power (EIRP) to overcome the large path loss and attenuation due to atmospheric conditions [3]. Due to the high antenna gain, these antenna systems have a narrow beam-width, in the range of 1° or smaller [4]. This necessitates electronic beam steering to simplify the alignment during installation and to solve the issue of twist and sway of the mast on which the antenna is mounted [5], [6]. In addition, electronic beam steering facilitates the flexible use of wireless fronthaul systems, allowing multi-beam operation and flexible routing of data streams. Such an antenna system should provide wide-angle scanning capabilities to provide coverage to various points at different locations.

An antenna system with high antenna gain, electronic beam steering and wide scan range can be realized using phased arrays. Such systems have been widely used in radio-astronomy and radar applications [7], [8]. Achieving a
directivity high enough for fronthaul or backhaul at E-band with a phased array is cost prohibitive, as several hundreds, if not thousands, of antenna elements are required, each having a phase shifter or vector modulator. Furthermore, this approach will require expensive active cooling since every element in the array is always active [9]. Therefore we propose the use of focal plane array (FPA) antennas as a solution for such application [10]. An FPA antenna is a system that consists of a reflector and an array of antenna elements in its focal plane. By changing the excitation weights of the antenna elements it is possible to change the beam shape and direction. This can make initial alignment easier and solves the mast twist and sway problem. Wide-scan FPA antennas have been used in various applications such as radio astronomy, passive remote sensing, and satellite communication [11]. With an FPA the required number of active antenna elements can be significantly reduced compared to a phased array. The benefit is that this decreases cost. However, with fewer elements active, the transmit power is also reduced. It is possible to intentionally increase the number of active antenna elements of an FPA by shaping the reflector or sub-reflector [12], [13], or by axially displacing the array [14], increasing the EIRP. Regardless of this design choice, to achieve wide angle scanning it is still needed to populate the full focal plane also resulting in an expensive system.

For point-to-multipoint operation it is not required to populate the full focal plane because it is known at the time of deployment in which directions the antenna system should scan. Therefore only the regions in the focal plane that contribute to the main beams need be populated with antenna elements. As such we propose an FPA antenna system where the focal plane is populated with several smaller arrays, each of which is responsible for communication with one site. This situation is depicted in Fig. 1. Each subarray optimally illuminates the reflector resulting in a highly directive beam. This novel combination of a reflector with subarrays for P2MP operation and twist and sway compensation allows for reliable and adaptive fronthaul and backhaul for 5G and 6G infrastructure.

This paper includes the following original contributions:

1) A study on the required size and inter-element spacing of a feed array for the proposed P2MP antenna system at E-band. Included are simulations of the directivity, aperture efficiency, spillover and estimations of the transmitted power and EIRP for such a system, relative to the maximum transmitted power of one antenna element \( (P_{TX,rel}, EIRP_{rel}) \). The reflector that is used allows for scanning to ±30°. Details about the design are given in Section III. The simulations are done in the cases where the beam is pointed towards 0° and 28° in azimuth. The margin of 2° is used for evaluating the performance when scanning an additional ±1° in azimuth and elevation. During the measurements the scan range is extended to ±2°. It is shown how the scan loss depends on the array size, the inter-element spacing, and the position of the array in the focal plane.

2) A comparison of the proposed system with antenna arrays as feeds to a system with typical horn antennas as feeds. To this end, a feed is defined with a Gaussian shaped radiation pattern, which is compared to the array case in terms of directivity, aperture efficiency and spillover loss.

3) Validation of the proposed system concept with measurements. Included are measurements of the embedded element patterns of an active beamforming array at K-band, and measurements of the FPA system when this beamforming array is used as the feed of the reflector. This includes measurements of the 0° and 28° configurations and for both the horizontal and vertical polarizations, as well as measurements when scanning an additional ±1° and ±2° in azimuth and elevation. Estimations of the directivity, aperture efficiency, spillover, relative transmitted power and relative EIRP are given. The excitation weights and phases of the antenna elements are shown. Measurements are performed at K-band because an E-band beamforming array was not available during this study. Given the losses are increased at E-band, it is difficult to compare the systems at K- and E-band in terms of radiation efficiency, and losses in the feed network. As such these kind of losses and inefficiencies are ignored, and we only look at a relative measure for EIRP and transmitted power. The reflector itself is considered lossless at both frequencies, and the operation and challenges are similar for E-band and K-band which allows for a useful comparison.

The remainder of this paper is structured as follows. In Section II the system concept is described in more detail and in Section III the model of this system is presented. In Section IV simulations are conducted at 80GHz and the optimum array size is determined. In Section V measurements are conducted with a demonstrator at 25.9GHz. The simulation results are compared to the experimental results in the discussion in Section VI and finally the conclusions are given in Section VII.

II. PROPOSED SYSTEM CONCEPT AND REQUIREMENTS
In the proposed concept the parabolic toroid reflector design as described in [15] is used. Rather than a single focus, this type of reflector has a circularly shaped focal arc. By moving a feed along the arc, the beam can be scanned in a different direction. The advantage of this type of reflector...
is that it allows for nearly identical beams when scanning the feed along the focal arc. The reflector from [15] is used here as well. The parabolic generating curve (Fig. 2) has a focal distance $F$ of 26.3cm. The distance from the origin of the parabola to the rotational axis $P$ is 65.5cm. The angle of the rotational axis with the horizontal axis is $\alpha = 70^\circ$. The $z$-axis of the feed is parallel to the rotational axis. By rotating the feed and the parabola around the rotational axis the focal arc and the parabolic toroid are generated. The opening angle of the feed $\gamma$ is $14.14^\circ$. The reflector is traced out of the parabolic toroid by removing the part that falls outside the opening angle of the feed for the full scanning range of $\beta = \pm 30^\circ$. The resulting reflector has a surface area of 0.1$m^2$. The diameter $D$ and width $W$ of the reflector, as depicted in Fig. 2, are 22cm and 55cm, respectively. For more information and a more extensive explanation on the design and operation principles of such reflector the reader is referred to [15], [16] and [17].

For the feeds, we propose to use square antenna arrays of dual-polarized patch antennas. For the power amplifiers (PAs), we propose to use silicon technologies (such as CMOS or BiCMOS), because large arrays of III-V technologies such as GaN at E-band are cost prohibitive. By using [18] and specifying a power back-off, one can arrive at an estimate of the absolute measure of transmitted power ($P_{TX}$) for various technologies and frequency ranges, which allows one to contextualize the relative metrics presented in this paper.

Finally, the requirements for such a system must fulfill are summarized as follows.

1) According to [19] the directivity of point-to-point antenna systems is required to be higher than 38dBi. With automatic gain control implemented the maximum transmission power is limited to 35dBm, and the total EIRP is limited to 85dBm.

2) For twist and sway compensation the requirement is to scan between $\pm 1^\circ$ in azimuth and elevation with a scan-loss of at most 1dB. The $\pm 1^\circ$ target is based on [6], where a study is conducted on twisting and swaying on different types of masts, and the effect on the quality of the link.

3) The system needs to allow dual-polarized operation such that two simultaneous data streams can be transmitted or received on the same channel, increasing the capacity.

III. SYSTEM MODEL

For the investigation of the proposed system concept, a model is required that can be used to calculate the directivity, relative transmitted power and relative EIRP. This model is defined as follows.

1) A mechanical model of the reflector is created in MATLAB and exported to GRASP, which is simulation software used for simulating reflector antennas [20].

2) In CST Microwave Studio a model of a square pinned fed balanced patch antenna is defined that is resonant at 80GHz. The far-field of one patch antenna is also exported to GRASP.

3) An array of patches is defined in GRASP that illuminates the reflector. Each patch has the farfield pattern as found in the previous step. The array is square and has a size of $N$-by-$N$ antennas, with inter-element spacing $d$, which is the same in both the $x$ and $y$ directions. The resulting situation is depicted in Fig. 2.

4) In GRASP the resulting secondary element patterns are found for each array element separately. These patterns $E_{co}$ and $E_{cross} \in \mathbb{C}^{N_{points} \times N_t}$ are electric fields normalized to the free-space impedance and the wavenumber, as will be explained later in this section. $N_t$ is the total number of antennas equal to $N^2$ for square arrays. $N_{points}$ is the number of spatial points the electric fields are calculated on.

5) The secondary element patterns are retrieved from GRASP and loaded into MATLAB where maximum ratio transmission (MRT) beamforming is applied.

MRT is a beamforming strategy that maximizes the received signal power at the intended user under the condition that the total transmitted power is limited to some power $P$. This is the optimal beamforming strategy when there is only one user, but not when there are multiple users, because the interference is not taken into account [21]. Here, the users are separated by a large angle, and the beam widths are small. The interference in this case is therefore ignored. To scan the beam in the intended direction using MRT beamforming, the excitation vector $w \in \mathbb{C}^{N_t \times 1}$ is found using

$$w = h^\dagger$$

where $h \in \mathbb{C}^{1 \times N_t}$ is the channel vector. The dagger symbol $^\dagger$ represents the hermitian transpose. In this context, the coefficients of $h$ represent the co-component of the electric field of antenna $n$ in the intended direction of the main beam. As such, the complex coefficients $h_n$ that build $h$ are found...
by evaluating the complex value of $E_{co}$ for antenna $n$ in the intended beam direction.

After finding the weights, and taking into account that the electric fields are normalized to the free-space impedance, we can find the directivity vector resulting from the beamforming $D \in \mathbb{C}^{N_{\text{points}} \times 1}$ using [22]

$$D = \frac{4\pi U}{P_{\text{rad}}} \tag{2}$$

with $U \in \mathbb{C}^{N_{\text{points}} \times 1}$ the radiation intensity and $P_{\text{rad}}$ the total radiated power. Here $D$ represents the directivity function $D(az, el)$ as a vector with $N_{\text{points}}$ combinations of $az$ and $el$. GRASP normalizes the electric fields such that

$$E = \frac{1}{k\sqrt{2Z_0}}E_{SI} \tag{3}$$

with $E_{SI}$ in the standard SI definition of electric field, $k$ the wavenumber and $Z_0$ the free-space impedance. With this definition the unit of $E$ is $\sqrt{W}$. $U$ is related to $E_{SI}$ by [22]

$$U = \frac{1}{2Z_0} \left( |E_{SI,co}|^2 + |E_{SI,cross}|^2 \right) \tag{4}$$

and as such to $E$ by (3) resulting in

$$U = k^2 \left( |E_{co}|^2 + |E_{cross}|^2 \right). \tag{5}$$

With this the radiation intensity can be found by multiplying $E$ by the weights as in

$$U = k^2 \left( |E_{co}|^2 + |E_{cross}|^2 \right) \tag{6}$$

which together with (4) results in

$$D = \frac{4\pi k^2 (|E_{co}|^2 + |E_{cross}|^2)}{P_{\text{rad}}} \tag{7}$$

where the total radiated power $P_{\text{rad}}$ is computed by integrating $U$ over a spherical grid that encloses the antenna array, meaning this can be written as

$$D = \frac{4\pi (|E_{co}|^2 + |E_{cross}|^2)}{\int \int |E_{co}|^2 + |E_{cross}|^2 \, d\Omega} \tag{8}$$

which is not dependent on $k$ anymore. The gain function is related to the directivity function by the radiation efficiency $\eta_{\text{rad}}$ as in

$$G(az, el) = \eta_{\text{rad}} D(az, el). \tag{9}$$

Given the reflector is considered lossless at the frequencies we are interested in, and the losses in the feed network and the losses in the antenna elements themselves are ignored, this means $\eta_{\text{rad}} = 1$. We can then find the effective aperture $A_e$ and the aperture efficiency $\eta_{\text{Ap}}$ using [22]

$$A_e = \eta_{\text{Ap}} A = \max(D(az, el)) \frac{\lambda_0^2}{4\pi} \tag{10}$$

with $A$ the physical area of the reflector. In order to calculate the spillover the power that is radiated from the array that does not hit the reflector ($P_{\text{miss}}$) is found and compared to $P_{\text{rad}}$ using

$$\text{spillover dB} = 10 \log_{10} \left( \frac{P_{\text{rad}}}{P_{\text{miss}}} \right) \tag{11}$$

where it should be noted that finding $P_{\text{miss}}$ requires an intensive simulation because it must be calculated by integrating over a sphere (as in the denominator of (8)) surrounding the complete setup, including the reflector. Because of the diffraction of the edges of the reflector, this must be done with a high resolution. Furthermore, there is no clear distinction between the fields that result from reflection and refraction of the reflector, and the fields radiated by the array that actually 'miss' the reflector. As such, to find $P_{\text{miss}}$ we integrated the power on the full sphere, excluding the region of the main beam and its direct side lobes. This requires some hand work, but results in a good approximation of the actual spillover loss.

Since $h$ is found by evaluating $E_{co}$ in the intended beam direction, the elements of $h$ and $w$ have the unit $\sqrt{W}$. We can therefore find the relative power in $W$ of element $n$ by squaring its weight $w_n$. As such, we can estimate the total transmitted power relative to the maximum transmitted power of a single element, $P_{\text{TX,rel}}$, in dB, by summing the powers of all $N$ elements as in

$$P_{\text{TX,rel}} = 10 \log_{10} \sum_{n=1}^{N} |w_n|^2 \tag{12}$$

and finally, we can estimate EIRP$_{\text{rel}}$ in dB with

$$\text{EIRP}_{\text{rel}} = 10 \log_{10} (\max(G)) + P_{\text{TX,rel}}. \tag{13}$$

### IV. SIMULATION RESULTS AT E-BAND

In this section a reference is defined which the focal plane array system can be compared to in terms of directivity, spillover and aperture efficiency. Furthermore we investigate the required number of patch antenna elements and the optimum inter-element spacing in order to perform MRT beamforming with low scan-loss at 80GHz.

### A. REFERENCE SIMULATIONS WITH GAUSSIAN FEEDS

In order to create a reference, a feed is defined with Gaussian shaped radiation pattern. This Gaussian pattern has a 10dB taper at an angle of $\gamma = 14.14^\circ$, corresponding exactly to the edges of the reflector as explained in Section II. The Gaussian feed is scanned along the focal arc by an angle ranging from $\beta = 0^\circ$ to $\beta = 30^\circ$ in steps of $10^\circ$, and the directivity, spillover loss and aperture efficiency are found for each case. These results are summarized in Table 1. From
FIGURE 3. A depiction of the electric fields in the focal plane, when the reflector is illuminated by a plane wave arriving from the directions $\text{el} = -1^\circ$, $\text{el} = 0^\circ$ and $\text{el} = +1^\circ$ respectively, with $\text{az} = 0^\circ$. The squares depict the optimum locations for the $d = 0.5\lambda_0$ (white) and $d = 0.7\lambda_0$ (red) $N = 8$ arrays. The figure shows that the scan loss decreases for an increased array size, because larger arrays are able to sample a larger part of the aperture distribution when scanning.

| Inter-element spacing $d$ | optimum offset [mm] | $N = 4$ | $N = 6$ | $N = 8$ | $N = 10$ | $N = 12$
|-----------------------------|---------------------|--------|--------|--------|--------|--------|
| $d = 0.5\lambda_0$ | optimum offset [mm] | 2.8    | 2.9    | 3.7    | 5.0    | 6.1    |
| $\text{az} = \pm 1^\circ$ scan loss [dB] | 5.6    | 2.94   | 1.46   | 1.22   | 0.8    |
| $\text{el} = \pm 1^\circ$ scan loss [dB] | 6.2    | 3.32   | 1.52   | 1.24   | 0.76   |
| $d = 0.7\lambda_0$ | optimum offset [mm] | 2.5    | 3.7    | 5.4    | 7.1    | 8.8    |
| $\text{az} = \pm 1^\circ$ scan loss [dB] | 3.40   | 1.31   | 0.76   | 0.73   | 0.52   |
| $\text{el} = \pm 1^\circ$ scan loss [dB] | 4.40   | 1.37   | 0.94   | 0.71   | 0.42   |
| $d = 1.0\lambda_0$ | optimum offset [mm] | 3.6    | 6.2    | 8.6    | 11.1   | 13.1   |
| $\text{az} = \pm 1^\circ$ scan loss [dB] | 1.60   | 0.69   | 0.40   | 0.14   | 0.11   |
| $\text{el} = \pm 1^\circ$ scan loss [dB] | 1.60   | 0.83   | 0.62   | 0.35   | 0.30   |

This data it is shown that the directivity is slightly reduced for scanning up to $30^\circ$, by only 0.3dB. However, in this configuration where the feed is scanned across the focal arc, the beam is also scanning in elevation plane. For the $0^\circ$ beam it scans to boresight, e.g., to $0^\circ$ in elevation, but it scans downward in elevation when scanning in azimuth. As such a fifth case is added to the table where the feed is placed away from the focal arc, such that it is scanning to $\text{az} = 28^\circ$, $\text{el} = 0^\circ$. In this case the directivity reduction is 1.7dB and the spillover is increased from 0.26dB to 0.44dB. The aperture efficiency is reduced from 23.9% to 17.3%.

The relatively low aperture efficiency is a consequence from the design of this reflector. By using Gaussian feed patterns, only a relatively small area of the reflector is illuminated. In contrast, patch antennas can illuminate a larger area of the reflector and as such increase the aperture efficiency and therefore the directivity.

B. OPTIMUM ARRAY POSITION FOR BROADSIDE SCANNING

In order to reach the optimum directivity and lowest scan loss using the array of patch antennas, the array must be positioned at the optimum location. Because the reflector is symmetric around the $vw$-plane (see Fig. 2) the optimum $y$-coordinate for broadside radiation is 0mm. However, the reflector is not symmetric in around the $uw$-plane, so the optimum $x$-coordinate is not as obvious. In order to find the optimum coordinate, the array is iteratively moved by an offset in the $x$-direction until the scan loss in the positive and negative elevation directions are equal. The optimum offset depends on the array size and the inter-element spacing. This is shown in Fig. 3. Here the electric fields arriving at the focal plane are plotted resulting from a plane-wave illumination of the reflector. The directions of the plane waves are $-1^\circ$, $0^\circ$ and $1^\circ$ in elevation, respectively. The white square represents the area sampled by an $N = 8$ array with $d = 0.5\lambda_0$ and the red square represents and $N = 8$ array with $d = 0.7\lambda_0$. The arrays both have the optimal offsets, which are clearly different for each size. This illustrates the importance of finding the optimum position for each configuration that is tested.

The described process is done for several array sizes ($N = 4 - 12$) and for several inter-element spacings ($d = 0.5\lambda_0, 0.7\lambda_0, 1.0\lambda_0$). For each configuration the optimum offset was found and the offsets are listed in Table 2. The resulting scan losses are shown in Fig. 4. The figure
FIGURE 5. Directivity towards $\omega = \phi = 0^\circ$ when using MRT beamforming for different setups. An inter-element spacing of $0.7\lambda_0$ is optimal.

shows that the 1dB scan loss requirement is satisfied for the $N \geq 10, d = 0.5\lambda_0$ arrays, the $N \geq 8, d = 0.7\lambda_0$ arrays and $N \geq 6, d = 1.0\lambda_0$ arrays. Clearly the scan loss decreases as either $N$ or $d$ is increased. Looking at Fig. 3 this indeed makes sense because an array with a larger aperture can still sample the focal field when the scan angle is increased.

C. RESULTING PARAMETERS AND COMPARISON TO REFERENCE FEEDS

By using the optimum offsets resulting from this iterative procedure, the directivity can be found. The directivity is calculated for each array size and inter-element spacing and the results are shown in Fig. 5. From this it becomes clear that the $0.7\lambda_0$ setup has the highest directivity when scanning to broadside. Knowing this, and knowing the scan-loss is below 1dB for the $d = 0.7\lambda_0, N = 8$ setup (see Fig. 4), this array meets our specifications. The array that is used for the measurements is 8-by-8 but has an inter-element spacing of $0.5\lambda_0$. Therefore we will include some calculations of an $0.5\lambda_0$ array as well.

Now $P_{\text{TX,rel}}$ and EIRP$_{\text{rel}}$ can be found, as well as $A_e$, $\eta_{\text{Ap}}$ and the spillover losses. These results are summarized in Table 3. It is shown that not only the directivity and scan-loss benefit from the increased inter-element spacing, as the aperture efficiency is higher and the spillover is lower in the $0.7\lambda_0$ case, for both the boresight and scanning beams. Furthermore, comparing the results of the $0.7\lambda_0$ array to the reference in Table 1, it is clear that the directivity and therefore the aperture efficiency is increased when using such array antenna. For the $0^\circ$ beam direction the directivity is improved by 1.9dB, and for the $28^\circ$ beam direction it is improved by 2.5dB. The spillover loss is also decreased for the $28^\circ$ beam by 0.12dB, although it is 0.09dB higher for the $0^\circ$ beam. It should be noted that the reference that uses the Gaussian patterns lacks the benefit from the flexibility of the proposed FPA antenna, since no compensation of twist and sway is possible with the fixed Gaussian excitation.

V. DEMONSTRATOR

To experimentally validate the performance of the proposed system, measurements are conducted at 25.9GHz with an experimental setup. The reflector used here has the same size and parameters as the simulated reflector.

### TABLE 3. Results of simulations at E-band using arrays of patch antennas with $N = 8$ and for a maximum scan angle of $1^\circ$.

| $f_0$ (GHz) | $N$ | $d(\lambda_0)$ | Scan range [$^\circ$] | $x_{\text{opt}}$ [mm] | $y_{\text{opt}}$ [mm] | $D$ (dB) | $P_{\text{TX,rel}}$ [dB] | EIRP$_{\text{rel}}$ [dB] | Scan-loss [dB] | Spillover loss [dB] |
|------------|-----|----------------|----------------------|----------------------|----------------------|--------|------------------------|-------------------|----------------|-------------------|
| 80         | 8   | 0.5            | $\pm 1.0$            | 5.7                  | -30.4                | 44.4   | 13.3                   | 57.6              | 1.52           | 30.5              |
| 80         | 8   | 0.7            | $\pm 1.0$            | 5.4                  | -28.6                | 44.4   | 13.3                   | 57.3              | 0.94           | 31.3              |
| 80         | 8   | 1.0            | $\pm 1.0$            | 2.8                  | -190.6               | 44.4   | 13.3                   | 55.7              | 0.78           | 30.6              |

A. EXPERIMENTAL SETUP

For these measurements the 8-by-8 dual-polarized beamforming array shown in Fig. 6 is used consisting of circular patch antennas. Each antenna element has a vector modulator that allows the phases and weights to be changed digitally with an 8-bit weight. The array in the picture has 10-by-10 elements, but the elements on the edges are dummies, and only the inner 8-by-8 are active. The inter-element spacing is $0.5\lambda_0$, or 5.8mm. The beamformer is controlled by an FPGA, and in turn the FPGA is controlled through USB by a computer running MATLAB.

The full 2D embedded element pattern for each antenna is measured using a planar near field scanner. The results are shown in Fig. 7 in a $(\theta, \phi)$ coordinate system, in terms of cuts in the horizontal ($\phi = 0^\circ$, $xz$-plane) and vertical ($\phi = 90^\circ$, $yz$-plane) directions, for both polarizations. The figure shows that the co/cross-polarization ratio of this array at boresight is on average 10dB. There is also a spread
between the levels of the radiation patterns at boresight of up to 4.5dB for the co-polarization. The average 3dB beamwidth is around 55°.

**FIGURE 7.** Cuts of the embedded element patterns of the 64 antenna elements in the array for both polarizations. The bold lines represent the average of all elements, the grey lines show the responses of each individual antenna element.

**FIGURE 8.** The reflector and the array in first configuration where it can scan to ±2° in azimuth and elevation.

**B. MEASUREMENTS SCANNING TOWARDS AZ = 0°**

The array is placed in the focal plane of the reflector as shown in Fig. 8. A characterization is done as follows.

1) An idle radiation pattern measurement is taken, by turning off all antenna elements and measuring the quiescent radiation.

2) A measurement is taken of the radiation pattern for each antenna element separately, by setting the weight of the element under test to the highest value, and turning off all other elements.

3) The idle radiation pattern is subtracted from the measured patterns of the elements.

4) The channel vector \( \mathbf{h} \) is built by extracting the complex value of co-component of the measured electric fields for each antenna in the intended beam direction. The results are used to find the excitation weights in MATLAB, according to the strategy described in Section III.

5) These weights are normalized such that the element with the highest power has a weight of 255.

6) The weights that were calculated are used to excite the array in the experimental setup.

7) A measurement is taken of the resulting radiation pattern and the directivity function is found.

Steps 4 through 7 are repeated for each measured beam. Because the planar near-field measurement only measures the power on the front half of the sphere, the measured directivity is overestimated. To adjust for this, an estimate of the spillover is found by simulating the system following the same strategy outlined in Section III, using the measured embedded element patterns shown in Fig. 7 and with the frequency set to 25.9GHz. The power is then found on the back half of the sphere. As a result, for the beams towards 0° ± 2° a spillover of 0.40dB is found, and for the beams towards 28° ± 2° a spillover of 0.44dB.
FIGURE 9. The reflector and the array second configuration where it can scan to $28 \pm 2^\circ$ in azimuth and $\pm 2^\circ$ in elevation.

is found. The spillover is subtracted from the measured directivity.

The full characterization for both polarizations (H and V) is done. For beams towards $az = 0^\circ$, $el = 0^\circ$ in both the H-pol and V-pol, the resulting weights and radiation patterns are shown in Fig. 10. Several observations can be made from this figure. It becomes clear that each antenna/amplifier combination has a considerably different power gain resulting in some antennas having a much higher excitation than others. This is the result from inaccuracies in the beamforming ICs, and the chosen beamforming strategy that excludes an extensive calibration of the array. It is also apparent that the cross-polarization is high in both the H-pol and V-pol case. This can be attributed to the cross-polarization levels of the antenna elements themselves as observed in Fig. 7.

To evaluate the performance of this FPA antenna, the main beam is scanned in steps of $1^\circ$ between $-2^\circ$ and $+2^\circ$ in azimuth and elevation. Cuts of these beams are shown in Fig. 10. It shows that the scan-loss is lower than 1dB for both polarizations up to $\pm 2^\circ$ in both directions. What stands out is that the directivity is slightly increased when scanning in positive elevation direction. By moving the array in the focal plane it is possible to decrease this effect and make the scanning in positive and negative elevation behave more similar, as described in Section IV.

For the center beam the directivity is $37.1\text{dBi}$ in H-pol and $37.0\text{dBi}$ in V-pol. The relative transmitted powers are $11.8\text{dB}$ and $9.6\text{dB}$ for the H-pol and V-pol beams, respectively. The difference between the powers results in a difference in EIRP$_{rel}$ as well, $48.9\text{dB}$ in H-pol and $46.6\text{dB}$ in V-pol. The reason for this difference is that one of the vertically polarized antenna elements operates at a much higher power level compared to the surrounding elements. Because of the chosen scaling strategy this leads to a reduction in transmitted power. A method to decrease this difference would be to scale the weights to a total transmitted power, instead of scaling to the weight of the highest excited element, or by slightly lowering the weight of the highest excited element relative to the other elements.

### C. MEASUREMENTS SCANNING TOWARDS AZ = 28°

Now the array is displaced by 28mm in $-x$-direction and 195mm in $-y$-direction. This corresponds to pointing a beam towards an angle of $+28^\circ$ in azimuth and $0^\circ$ in elevation. The reflector itself is rotated by $-28^\circ$ such that the beam is pointed towards the near field scanner. This setup is shown in Fig. 9. It is shown that the array is not rotated around the z-axis, because the experimental setup does not allow this. Due to the geometry of the reflector, this results in a rotation of the polarization. This means that exciting an H-pol element does not result in an H-pol beam. The same holds for the V-pol. Instead, two new planes are defined, H' and V', which are rotated clockwise by $14^\circ$ with respect to the H and V planes, respectively. This corresponds to a scan angle of $28^\circ$. As such, exciting the H-pol elements results in an H'-pol beam, and exciting the V-pol elements results in a V'-pol beam. With this in mind the same measurements are repeated. In post-processing the measured electric fields are found the newly defined H' and V' planes.

The resulting beams in the H'-pol and V'-pol are shown in Fig. 11. The cuts of beams between $\pm 2^\circ$ in azimuth and elevation are also shown. The highest scan-loss is $0.60\text{dB}$ and $0.65\text{dB}$ for the H'-pol and V'-pol beams, respectively. The directivity for the center beams are $35.4\text{dBi}$ in H'-pol and $35.5\text{dBi}$ in V'-pol. The relative transmitted power for H'-pol is $12.7\text{dB}$ and for V'-pol $11.7\text{dB}$. This difference in power is smaller than when scanning to $az = 0^\circ$. The resulting EIRP$_{rel}$ is $48.2\text{dB}$ and $47.3\text{dB}$ for H'-pol and V'-pol respectively. The findings are summarized in Table 4.

### VI. DISCUSSION

To validate the performance of the E-band system concept, which optimally uses an 8-by-8 $0.7\lambda_0$ array, it is possible to compare the simulations with the experimental setup at K-band. The most important difference is that the measurement array has a physical area that is $4.9$ times larger. This is the reason that the experimental setup has a wider scan range, since the location of the focal spot, when scanning, moves by the same distance in the focal plane regardless of

| TABLE 4. K-band measurement results. |
|--------------------------------------|
| Polarization | K-band 0° | K-band 28° |
|--------------|-----------|------------|
| $f_0$ [GHz]  | 25.9      | 25.9       |
| $N$          | 8-by-8    | 8-by-8     |
| $d$ [$\lambda_0$] | 0.5 | 0.5 |
| Scan range [°] | ±2.0 | ±2.0 |
| $P$ [dB]     | 37.1      | 37.0       |
| $P_{T_Xrel}$ [dB] | 11.8 | 9.6 |
| $EIRP_{rel}$ [dB] | 48.9 | 46.6 |
| Max. scan-loss [dB] | 0.70 | 0.80 |
| Estimated spillover loss [dB] | 0.40 | 0.44 |
| $\tau_{spol}$ [%] | 54.8 | 53.5 |
FIGURE 10. Measurement results of the 25.9GHz experimental setup when scanning to broadside. In the left column the excitations for the H- and V-polarized beams are shown in terms of weights and phases. The displayed phases are rounded to degrees. In the middle column the co- and cross-components of the H- and V-polarized beams are shown. In the right column the beam cuts are shown in terms of directivity when scanning to $\pm 2^\circ$ in azimuth and elevation for both polarizations. $\alpha_z$ and $\alpha_{el}$ denote the scanning directions.

frequency. This is evidenced by the measurements, where at most 0.80dB of scan loss is observed when scanning up to $\pm 2^\circ$, whereas this is 0.94dB for the simulated E-band system when scanning only up to $\pm 1^\circ$.

There are also some important similarities between the experimental setup and the simulations. We see in Tables 3 and 4 that the aperture efficiency is lower for the beam in the $\alpha_z = 28^\circ$ direction compared to the beam in the $\alpha_z = 0^\circ$ direction. However, the scan-loss to $\pm 1^\circ$ and $\pm 2^\circ$ is actually lower for the $\alpha_z = 28^\circ$ case.

$P_{tx,rel}$ and EIRP$_{rel}$ are relatively unchanged for the $\alpha_z = 0^\circ$ and $\alpha_z = 28^\circ$ beams, both in simulations and measurements, where the largest decrease in EIRP is 1.03dB for the E-band $d = 0.7\lambda_0$ system. This shows that this system is indeed well suited for wide-angle scanning.
FIGURE 11. Measurement results of the 25.9GHz experimental setup when scanning to \( \text{az} = 28^\circ \). In the left column the excitations for the H'- and V'-polarized beams are shown in terms of weights and phases. The displayed phases are rounded to degrees. In the middle column the co- and cross-components of the H'- and V'-polarized beams are shown. In the right column the beam cuts are shown in terms of directivity when scanning to \( 28 \pm 2^\circ \) in azimuth and \( \pm 2^\circ \) in elevation for both polarizations. \( \text{az}_0 \) and \( \text{el}_0 \) denote the scanning directions.

The benefits of moving from a \( 0.5\lambda_0 \) to a \( 0.7\lambda_0 \) have been shown to be many, including an improvement in directivity, a lowering of the scan-loss, and a decreased spillover loss. It is also shown that the EIRP_{eq} is slightly reduced, because less power is transmitted. However it should be mentioned that one power amplifier can now occupy a larger area, which gives some freedom in thermal design, allowing the power per PA to be improved to compensate. Furthermore the increased spacing is also beneficial in terms of mutual coupling.

VII. CONCLUSION
A new system concept for point-to-multipoint communication was proposed at E-band for fronthaul and backhaul applications. With this concept several beams can be formed by placing multiple antenna arrays in the focal plane. It was shown that a parabolic toroid reflector with an array...
A demonstrator is built at K-band which uses an 8-by-8 analog beamforming array. At this frequency the relative EIRP is higher than 46.6dB and the measured directivity is higher than 45.0dBi for the beams towards 0° and 28° in azimuth. Furthermore, each beam can be scanned up to ±1° in azimuth and elevation by changing the excitation weights of the array, with at most 0.94dB of scan loss. This alleviates twist and sway issues and simplifies initial alignment. It was shown that this system employing an antenna array outperforms the reference which uses Gaussian shaped element patterns.

In conclusion, this system is shown to be a good candidate for robust and adaptive fronthaul or backhaul for 5G and 6G wireless networks.

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