A High-Gain 58-GHz Box-Horn Array Antenna with Suppressed Grating Lobes

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***Abstract*—A low-profile high-gain antenna array of box horns for the frequency band 57.2–58.2 GHz is presented. The antenna consists of 256 radiating elements divided into two subgroups of 128 elements fed by a rectangular waveguide feed network. The radiating elements are fed in parallel and the waveguides are connected with T-junctions. The matching of the T-junctions is improved with a matching pin and a splitter. Because of the waveguide feed network, the element spacing is larger than one wavelength, which causes grating lobes. The grating lobes and sidelobes in the H-plane have been suppressed by the use of a combination of subarrays, a special characteristic of the box horn, and an array amplitude tapering. The measured sidelobe levels in the H-plane are below -30 dB at angles larger than 8° from boresight. A gain higher than 35.7 dBi and a return loss higher than 14.4 dB have been measured for the antenna over the band 57.2–58.2 GHz.**

***Index Terms*— Grating lobes, horn antenna, millimeter-wave antenna arrays.**

I. Introduction

The demand for planar high-gain and high-efficiency antennas is increasing due to a number of millimeter wave applications such as car collision avoidance radar, wireless local area network (LAN), and radio links [1], [2]. A simple and efficient solution is important for commercial applications. Several different types of planar antennas have been reported [3]–[7]. At millimeter waves two factors limit the use of these antennas: one is the efficiency, the other the bandwidth. In the case of radio link antennas for 57.2–58.2 GHz, the required bandwidth is only 1.73%, but it is essential that the direction of the main beam is frequency independent and that all properties, including low sidelobe levels, remain stable over the whole band. The gain should exceed 35 dBi, which is difficult to reach with a wide-band planar antenna even at lower frequencies [5].

The main objective is to achieve low sidelobe levels in the H-plane radiation pattern in addition to a high gain and efficiency. Furthermore, the objectives should be achieved with a simple and low-profile antenna. A high efficiency requires a low-loss transmission line in the feed network. At millimeter-wave frequencies a waveguide has low transmission losses, which makes it a good alternative. Waveguide feed networks have been used before, but they are complex three-dimensional structures that are far from low profile and are demanding to manufacture [8], [9]. The matching of the antenna is also of importance since radio links are duplex systems and reflections from the transmitter to the receiver degrades the performance of the connection. In this paper, a low-profile antenna array with a waveguide feed network is presented for the radio link application band 57.2–58.2 GHz.

The antenna is made by placing waveguide fed horns next to each other in rows and columns with element spacings larger than one wavelength. The large element spacings are a result of the use of rectangular waveguides in the feed network. This, on the other hand, results in grating sidelobes. In this antenna, low sidelobe levels are achieved, despite the large element spacing, by combining two arrays shifted sideways and by using a radiating element, which has been designed to reduce grating lobes in both planes. Here the principle of an array having a radiation pattern, which is the product of the element pattern and the array factors is used [10]. Mutual coupling has been neglected to keep the design procedure simple. High return loss is realized by using well-matched power dividers. Still the construction of the antenna is very simple, it consists of only two parts: one contains the feed network, the other the radiating elements.

II. Suppression of Sidelobes

An element spacing larger than one wavelength leads to grating lobes in the array factor [10]. The radiation pattern of an antenna array is a product of the element pattern and the array factor. The combination of arrays gives additional array factors, which can be used to create a null in the desired direction.

The array factor of the whole antenna array is the product of the array factor of the subarray and the array factor of a two-element array. The normalized H-plane array factor (AF) of a combination of two arrays in Fig. 1 is

$$AF(\theta) = AF_C(\theta)\cdot AF_{\mathrm{sub}}(\theta) = \cos\left[\frac{1}{2}\,(kd\,\sin(\theta+\beta))\right] \cdot\left[\frac{\sin(8kd_h\,\sin(\theta+\beta))}{16\cdot\sin(kd_h\,\sin(\theta+\beta)/2)}\right] \quad (1)$$

where k is the wavenumber, d is the distance between the subarrays in the H-plane, d_h is the element spacing in the H-plane, θ is the angle of radiation, and β is the relative phase of the element [10]. The element spacing in the H-plane $d_h = 1.8\lambda$. Then the array factor of the subarray AF_{sub} has a grating lobe in the direction 34.4° from the main beam and by selecting the shift d properly the array factor of the

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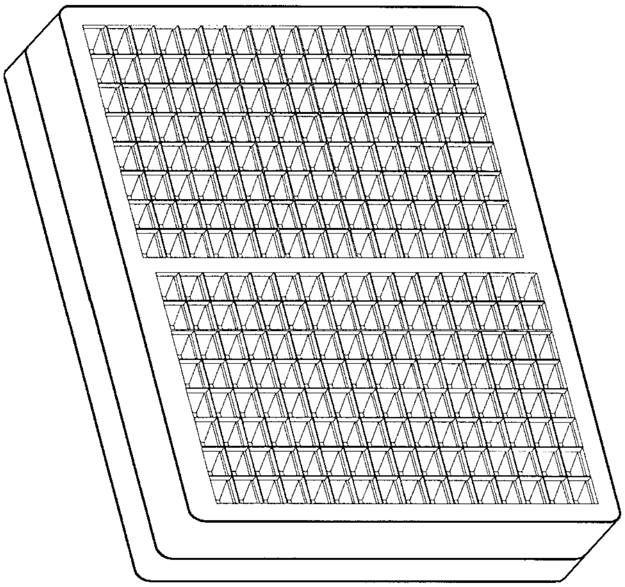

Fig. 1. A 256-element antenna array.

combination of two subarrays AF_c has a zero in the same direction. The grating lobe from the array factor of the subarray is eliminated by the zero in the array factor of the combination of two arrays AF_C, when $d = 0.9\lambda$. The result is the same as having an array with the element spacing 0.9λ, with the elements placed on two parallel lines separated in the E-plane. The grating lobes in the E-plane can also by reduced by a combination of arrays, but this would complicate the design of the feed network.

The null from the additional array factor can be used together with the nulls of the element radiation pattern to further reduce the grating lobes. The direction of the zero in the element radiation pattern determines the angle at which the sidelobes will be reduced. The other sidelobes at large angles from the main beam can be reduced by applying an amplitude tapering over the antenna aperture [10]. An amplitude tapering has been applied in the H-plane on the elements closest to the edge. Their relative amplitudes are 0.36 and 0.6 starting from either edge. By using all three methods presented here the radiation pattern of the antenna array will be highly directive with low sidelobe levels in the H-plane.

III. Radiating Elements

The directions of the grating lobes are determined by the element spacing. A uniform aperture distribution over the interelement distance has zeros in the radiation pattern in the same direction as the grating lobes [10]. The field distribution in a rectangular waveguide is evenly distributed in the E-plane and by enlarging the same field distribution over a larger area a uniformly illuminated aperture can be achieved. The field can be distributed over a larger area with an E-plane horn. The widening of the height of the waveguide results in a phase error in the aperture of the horn. The size of the phase error depends on the size of the aperture and the length of the horn. The radiation pattern of an aperture with a phase error smaller than 45° does not significantly differ from

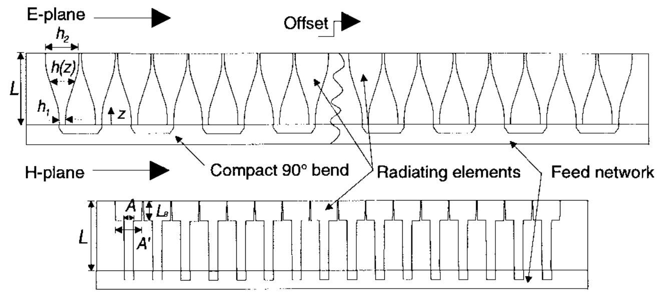

Fig. 2. E- and H-plane cuts.

the pattern of an aperture without phase error [10], but it is still too large to completely eliminate the grating lobes. The length of the radiating element directly affects the thickness of the whole antenna array. Therefore, it is essential to find a short element with a small aperture phase error. These two requirements contradict each other since a small aperture phase error requires a long horn. By using an opening function from [11], the length of the horn can be reduced by one third compared to a linear opening function without increasing the phase error in the aperture. The opening function is

$$h(z) = h_1 + (h_2 - h_1)\left[\frac{z}{L}(1-C) + C\ \sin^2\left(\frac{\pi}{2}\ \frac{z}{L}\right)\right] \quad (2)$$

where $h(z)$ is the dimension in the E-plane at the distance z from the input port, h_1 is the dimension at the input port, h_2 is the aperture size, L is the total length of the horn, and C determines the amount of profiling (see Fig. 2). A value of $C = 0.7$ gives a reasonably gradual change in the E-plane dimension [11]. This opening function has been presented for a corrugated horn, but according to simulations it is valid also for E-plane horns without corrugation. The E-plane dimensions of the radiating element have been designed using (2) and setting the allowed aperture phase error to 27°, resulting in a horn shorter than 3.3λ. The grating lobes in the E-plane will not be completely eliminated since the element wall thickness cannot be infinitely small and the aperture phase error slightly shifts the direction of the zeros in the radiation pattern.

Reduction of the grating lobes in the H-plane with a normal H-plane horn is not as successful as the use of an E-plane horn in the E-plane. The field distribution in the aperture of an H-plane horn is cosine-shaped and the zeros in the radiation pattern of an H-plane horn with the aperture size equal to the interelement distance are therefore at a much larger angle than the grating lobes. A different kind of horn is needed to reduce the levels of the grating lobes.

A special type of a horn called the box horn has been presented in [12]. A higher order wavemode TE_{30} is excited at a sharp transition in the H-plane width from A to A' in Fig. 2. The relative amplitude of the higher order wavemode TE_{30} depends on the size of the transition. The wavemode TE_{30} and the fundamental wavemode TE_{10} propagate at different speeds. Therefore, by choosing the length of the box $L_B = 0.89\lambda$, the two wavemodes TE_{10} and TE_{30} will be in opposite phase at the aperture. Then the direction of the zero in the radiation pattern of the box horn can be adjusted with the relative amplitude of the higher order wavemode TE_{30}. The relative amplitude of the excited wavemode TE_{30} is set to 0.45.

The H-plane dimensions of the element have been designed so that the pattern zero will be at a slightly larger angle from the main beam than the grating lobe. In this way the zero will significantly reduce the sidelobes in the angular region 30°–50° from the main beam. The E- and H-plane cuts have been designed separately and the length of the E-plane horn or allowed aperture phase error will determine the total length of the radiating element.

IV. Feed Network

In this case, the antenna characteristics should remain as stable as possible over the band 57.2–58.2 GHz. Therefore, every radiating element should be fed in parallel. A parallel feed requires long transmission lines in the feed network. Therefore, it is essential that the losses in the transmission line used in the feed network are very low in order to reach a high antenna efficiency. At about 60 GHz, rectangular and circular waveguides have low losses and they are very simple structures. In comparison with the circular waveguide, the rectangular waveguide is easier to make by a milling machine. In addition, making a T-junction of rectangular waveguides is also straightforward. The main drawback of rectangular waveguides is the width of the transmission line. The width of the rectangular waveguide determines the cutoff frequency, which, in turn, determines the frequency range of operation of the waveguide. The width of a rectangular waveguide filled with air is at least half a free-space wavelength. The use of rectangular waveguides in a feed network in one plane will, therefore, result in an element spacing larger than one wavelength. The feed network has to be built in one plane in order to achieve a low-profile antenna design and to keep the antenna structure simple.

A. Element Spacing

The directivity as a function of the element spacing reaches local maxima about every 0.9λ [13]. In order to maximize the directivity, the element spacing in the H-plane has been increased to about 1.8λ from the smallest possible value of approximately 1.3λ. The number of grating lobes is proportional to the element spacing in wavelengths [10] and it therefore remains unchanged despite the larger spacing. The additional space obtained in the feed network with the larger element spacing is used for the power dividers and wider waveguides, thus further decreasing the losses in the transmission line. The element spacing in the E-plane is largely determined by a waveguide section with an electrical length of half a wavelength. This phase corrector is needed to realize the same phase for each element. The E-plane element spacing is set to the smallest possible value, which is about 2.3λ. The width of the rectangular waveguide in the feed network has been set to 3.2 mm. The height of the waveguide is chosen to be 2.39 mm, which is the same as the height of the standard waveguide WR-19. The cutoff frequency for a waveguide with these dimensions is 46.875 GHz. It is sufficiently low to ensure low-loss operation above 57 GHz. The antenna is to be fed from a WR-15 waveguide and the transition from the standard waveguide to the waveguide used in the feed network is made linearly in both planes similarly to a pyramidal horn.

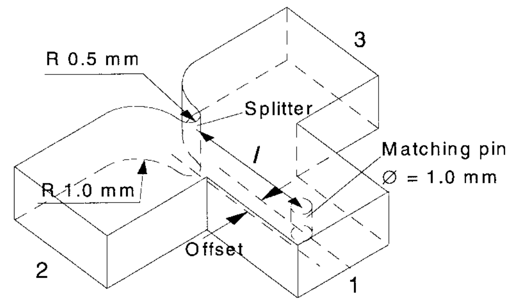

Fig. 3. Power divider.

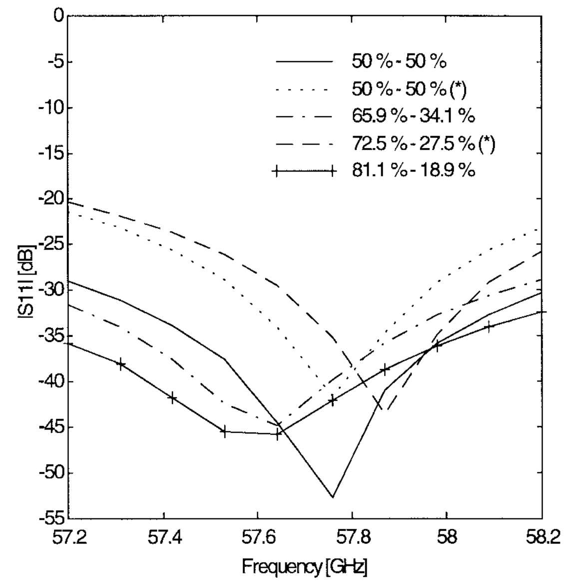

Fig. 4. Simulated reflection coefficients for the power dividers with different power division ratios.

B. Power Divider

A concept of realizing well-matched power dividers for 37.0–39.5 GHz with different power division ratios has been presented in [14]. The matching of the T-junction is improved by introducing a 3-mm-long shaped splitter in the junction. The shape of the splitter has been obtained empirically. The splitter is also used to realize unequal power division ratios without phase difference, by having an offset between the centerlines of the splitter and the input port of the T-junction (see Fig. 3). The matching of the junction is further improved by placing matching pins in the output ports. The same principles have been used here and the power divider used in the feed network is shown in Fig. 3. In this case only one matching pin has been used, but the bandwidth of the T-junction is large enough for this kind of application. The distance l in Fig. 3 between the splitter and the matching pin is 0.76λ–0.91λ depending on the power division ratio. The simulated reflection coefficients of the power dividers with different power division ratios (percentages) are shown in Fig. 4. The offset is, e.g., 0.087λ for the power divider 81.1–18.9%. The different power division ratios are used to taper the amplitude distribution in the H-plane and thereby

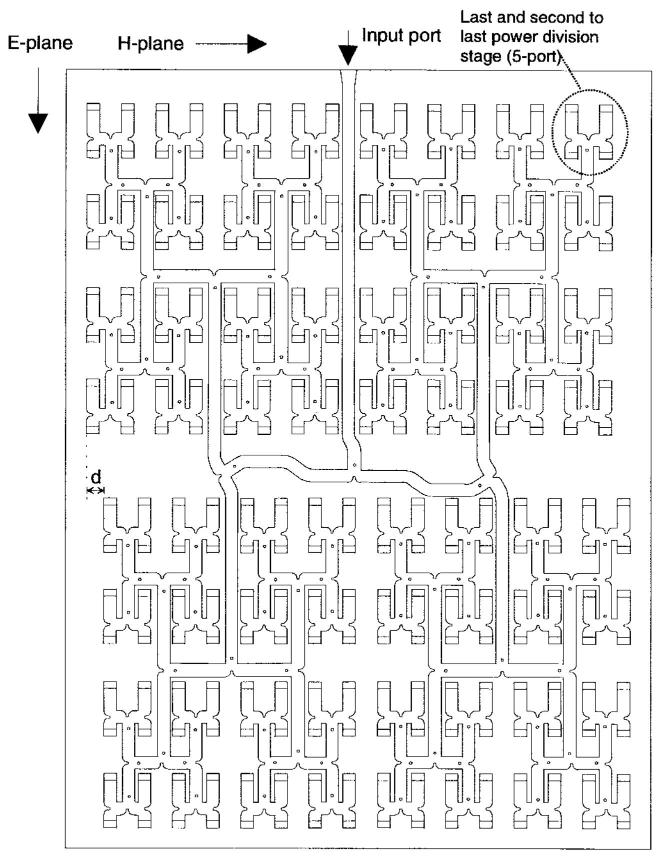

Fig. 5. Feed network.

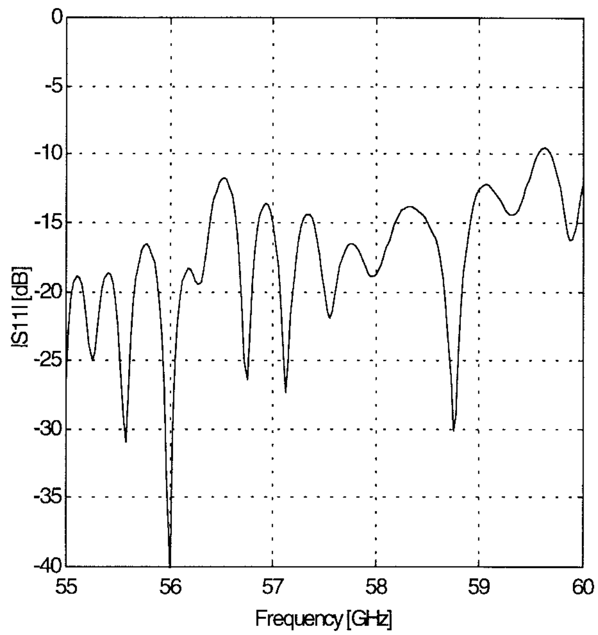

Fig. 6. Measured reflection coefficient.

reducing the sidelobe level at angles larger than 50° from the main beam. There is not enough room for a matching pin in the last power division stage. The matching of the last and second to last stage is done with the matching pin in the power divider in the second to last stage. The simulated reflection coefficients for the last two stages together are indicated in Fig. 4 with (*). The software used in the simulations was the High Frequency Structure Simulator® from Hewlett-Packard. The transition from the feed network to the radiating element was done by a compact 90° bend shown in Fig. 2, where a 45° slant was used to obtain a high return loss for the bend. The simulated return loss of the bend is higher than 43 dB over the frequency band 57.2–58.2 GHz.

The feed network has been designed to parallel feed two 128-element arrays, which are shifted sideways 0.9λ in the H-plane as shown in Fig. 5. The aperture efficiency of a single element is about 50% and the illumination efficiency 93%. With 256-elements the antenna aperture size is approximately 250 cm^2. The distance from the input port to a single radiating element is about 0.3 m and the attenuation in the waveguide is approximately 2.5 dB/m. This indicates an antenna gain of about 36.2 dBi, if the losses due to mismatch in the power dividers are estimated to 0.05 dB per junction corresponding to the reflection coefficients in Fig. 4.

V. Measurement Results

A prototype antenna consisting of 256 radiating elements divided into two subarrays of 128 elements each has been manufactured according to the design presented here using conventional milling techniques. The prototype antenna is made of 18 aluminum parts kept tightly together with screws. The feed network has been milled from one plate, but the part with the radiating elements has been built of 17 pieces to avoid rounded corners in the cross section of the radiating element. The overall dimensions of the prototype is about 18 cm $\times$ 22 cm $\times$ 3 cm (width, height, thickness). The whole antenna is shown in Fig. 1 and the E- and H-plane cuts in Fig. 2.

Great care must be taken in making the waveguides, since the width of the waveguide affects the propagation characteristics of the signal. Small fluctuations in the width of the waveguide can accumulate into large phase differences between the elements due to the long transmission distance from the input port to a single element. Some of the matching pins in the prototype are not correctly made and this degrades the performance of the prototype antenna.

The measured return loss of the antenna is shown in Fig. 6. The return loss is higher than 14.4 dB over the frequency band 57.2–58.2 GHz, but remains higher than 12 dB over the band 56.55–59.45 GHz. The gain of the prototype antenna was measured with the gain comparison method. A 23.15 dBi standard gain horn was used as reference. The results of the gain measurement are shown in Table I. The gain is between 35.75 and 36.15 dBi over the band 57.2–58.2 GHz. This corresponds to an aperture efficiency higher than 33%. The measured gain is close to the estimated gain of 36.2 dBi.

The radiation pattern and gain measurements were performed outdoors at a distance of 25 m since a sufficiently large anechoic chamber was not available. The measured and calculated E-plane radiation patterns at the center frequency 57.7 are shown in Fig. 7. Grating lobes due to the array factor can be seen in the measured radiation pattern at the angles 25.5° and 56° from boresight. These sidelobes are not completely canceled out by the zeros in the radiation pattern

TABLE I
MEASURED ANTENNA GAIN

f [GHz]	G [dBi]
57.2	35.75
57.3	35.85
57.6	35.85
57.7	35.85
57.8	36.15
58.0	36.05
58.2	35.75

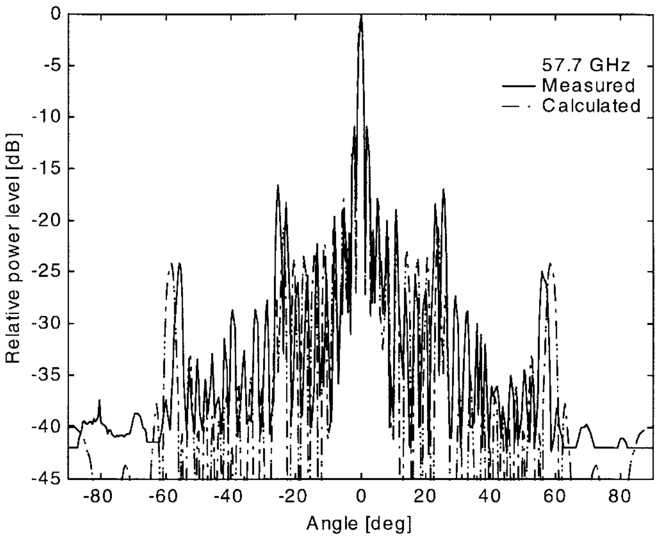

Fig. 7. E-plane radiation pattern at 57.7 GHz.

of the element. The relative power levels of the sidelobes are -17 and -25 dB, respectively. The H-plane radiation pattern has been measured at the frequencies 57.2, 57.7, and 58.2 GHz giving very similar results. The calculated and measured H-plane radiation patterns at 57.7 GHz are shown in Fig. 8. The sidelobe level remains below -35 dB at angles larger than 21° over the band 57.2–58.2 GHz. At angles larger than 36° the sidelobe level is below -40 dB. At 57.7 GHz the measured sidelobe level at the angle 35° is about 5 dB higher than the calculated sidelobe level. The asymmetry in the levels of the first sidelobes in the H-plane is due to manufacturing errors in the feed network.

The dynamic range in the measurement was not sufficient to accurately measure sidelobe levels in the H-plane at larger angles than 40° from the main beam. The measured sidelobe level is very close to the noise floor, which increases slightly the measured sidelobe level from the real sidelobe level. The angular accuracy in the radiation pattern measurement is affected by the inaccuracy of the manually controlled rotation gear, which explains the small differences between the calculated and measured angular directions for the sidelobes. The measurement results agree with the calculated results and confirm the theoretical design.

VI. CONCLUSIONS

A low-profile high-gain antenna has been designed for the frequency band 57.2–58.2 GHz. For suppression of the grating

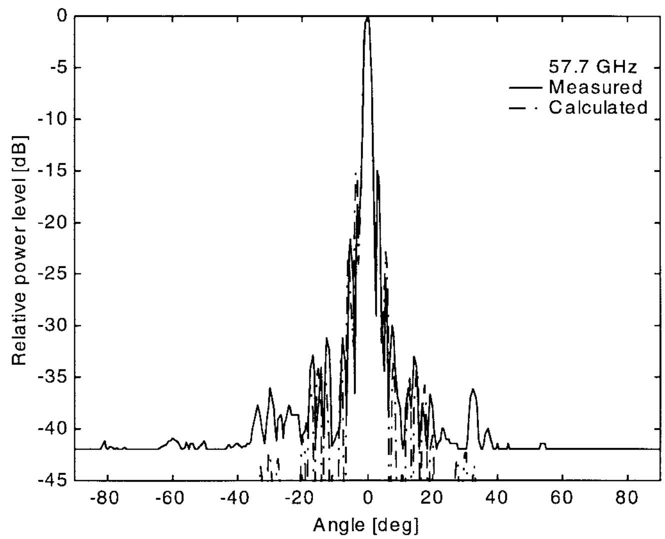

Fig. 8. H-plane radiation pattern at 57.7 GHz.

lobes due to the large element spacing, two techniques have been used: a proper radiating element and combination of arrays. The measured sidelobe level in the H-plane is below -35 dB at larger angles than 21° from the main beam. The estimated gain was 36.2 dBi and the measured gain is higher than 35.75 dBi, corresponding to an aperture efficiency of 33%. The return loss is higher than 14.4 dB over the band of operation, but stays higher than 12 dB over the range 56.55–59.45 GHz (5%). The experimental results confirm the theoretical analysis. The results have been achieved with a simple planar antenna, where every radiating element is fed in parallel. A combination of two box horn arrays has showed its usefulness as a high-gain planar antenna suitable for millimeter wave frequencies.

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