High-Gain Phased Array Antenna with End-Fire Radiation for 26 GHz Wide-Beam-Scanning Applications

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Abstract—In this paper, a high-gain phased array antenna with wide-angle beam-scanning capability is proposed for fifth-generation (5G) millimeter-wave applications. First, a novel, end-fire, dual-port antenna element with dual functionalities of radiator and power splitter is designed. The element is composed of a substrate integrated cavity (SIC) and a dipole based on it. The resonant frequencies of the SIC and dipole can be independently tuned to broaden the impedance bandwidth. Based on this dual-port element, a 4-element subarray can be easily constructed without resorting to a complicated feeding network. The end-fire subarray features broad beam-width of over 180 degrees, high isolation, and low profile, rendering it suitable for wide-angle beam-scanning applications in the H-plane. In addition, the methods of steering the radiation pattern downwards or upwards in the E-plane are investigated. As a proof-of-concept, two phased array antennas each consisting of eight subarrays are designed and fabricated to achieve the broadside and wide-angle beam-scanning radiation. Thanks to the elimination of surface wave, the mutual coupling between the subarrays can be reduced for improving the scanning angle while suppressing the side-lobe level. The experimental predictions are validated by measurement results, showing that the beam of the antenna can be scanned up to 65 degrees with a scanning loss only 3.7 dB and grating lobe less than -15 dB.

Index Terms—Antenna, antenna array, beam-scanning, end-fire, millimeter-wave, fifth-generation (5G).

I. INTRODUCTION

Since the International Telecommunication Union (ITU) and Federal Communications Commission (FCC) licensed several frequency bands in millimeter-wave (mm-Wave) for the fifth-generation (5G) applications in 2015 [1] and 2016 [2], the research on mm-Wave antenna/array has attracted significant attention, from both academia and industries. Compared with the conventional cellular system operating in the microwave band, the mm-Wave cellular system has the advantages of wide operating bandwidth and compact footprint due to reduced wavelength, thus less interference. On the other hand, the mm-Wave wireless system suffers from severe path loss in free space, which is usually improved by increasing the antenna gains at both base stations and mobile terminals [3]–[4]. As a trade-off, sub-millimeter wave (26 GHz) is believed to be a good candidate for commercial 5G application due to the properties of wideband and relative low path loss.

Several high-gain mm-Wave array antennas based on different structures were proposed, including microstrip patch antenna (MPA) [5], planar aperture antenna (PAA) [6], substrate integrated waveguide (SIW)-fed magneto-electric (EM) dipole [7], and log-periodic dipole

antenna (LPDA) [8]. However, these antennas are not suitable for beam scanning applications and thus the coverage is very limited due to the constraints of structures or feeding networks. Recently, several phased arrays operating at mm-Wave band for wide scanning angle were reported [9]–[12]. In [10], one-dimensional (1-D) stacked patch antenna array was presented for 5G applications. In [11], both 1-D and 2-D beam scanning properties are achieved based on multiple-layer meta-material structure. The mm-Wave antenna was also realized by a SIW fed tapered slot antenna in [12], but it is not based on conventional printed circuit board (PCB) method and thus leads to a high fabrication cost. Recently, mm-Wave antennas with the end-fire radiation characteristics achieved by using a low-profile substrate has attracted strong research interest due to the reduced physical dimension in the mm-Wave frequency and enhanced beam-width, which are desired features in phased array antenna design for realizing a wide-angle beam-steering capability [13]–[14].

In this paper, an antenna with end-fire radiation is proposed for mm-Wave application. Unlike traditional designs, the antenna is composed of a dipole extended from a substrate integrated cavity (SIC) and features two asymmetric ports, which are used as input and output, respectively. This antenna element is an extended design from our previous work in [15], where a planar omni-directional antenna was reported. In this work, the end-fire radiation property is achieved by utilizing only one dipole. The wide beam-width, the down-tilt capability in the E-plane, and the mutual couplings between the subarrays are investigated for the first time. In addition, a high-gain phased array is demonstrated to show the improved wide-angle beam-scanning performance over conventional phased arrays. The proposed dual-port antenna can not only function as a radiator but also a power splitter. Due to a very broad beam-width and high isolation property, this antenna is further developed to form a wide-angle beam-scanning phased array. As a proof-of-concept, two 4×8 antenna arrays are designed to demonstrate the broadside radiation and beam-scanning capabilities. Experimental results demonstrate that phased array can scan over a wide angle up to 65 degrees with a scan loss less than 4 dB. Moreover, the method of realizing the down/up-tilt beam is investigated for potential 5G cellular applications.

The reminder of this paper is organized as follow. Section II illustrates the dual-port element, subarray, and the realization of the beam-tilt. Section III presents two phased array antennas and their feeding networks, which is followed by concluding remarks in Section V.

II. ANTENNA IMPLEMENTATION

In this section, a novel dual-port antenna element will be first presented and followed by the implementation of the corresponding subarray and the method of tilting the beam in E-plane.

A. Dual-Port Antenna Element

Fig. 1(a) shows the geometry configuration of the proposed end-fire mm-Wave antenna, which is composed of a substrate integrated cavity (SIC) and a pair of dipole extended from the SIC. Different from its

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Fig. 1. Configurations of the dual-port SIC-based antenna element. $L_d = 2.1$ mm, $h_2 = 1.5$ mm, $L_1 = 1.1$ mm, $L_2 = 1.2$ mm, $L_3 = 1.2$ mm, $L_4 = 0.8$ mm, $L_5 = 0.8$ mm, $L_6 = 0.4$ mm.

conventional counterparts, the proposed antenna features two asymmetrical ports (Port-1, Port-2) at both sides of the cavity, functioning as the power input and power output, respectively. The SIC is realized by connecting the top and bottom layers of the substrate using circular metallic via holes. The size and distance of the via holes follow the designing rules presented in [16]. On one side of the SIC, a dipole is formed by extending two strips from the top and bottom of the SIC, which can generate the end-fire radiation. The length of the dipole is around a half-wavelength at the designed center frequency (26 GHz). The RO4003C substrate with a permittivity of 3.55 and a loss tangent of 0.0027 is used in the design. The thickness of the substrate is 0.3 mm. The design and optimization were performed using the high frequency structure simulator (HFSS) and the optimized parameters are listed in the caption.

In order to elucidate the operating principles, an equivalent circuit is shown in Fig. 1(b). The SIC can be regarded as a resonator and represented by a shunt $L_1 C_1$, whose dominant resonant frequency can be approximately derived by using the substrate integrated waveguide (SIW) methods [16]. The dipole as the radiator, is represented by a shunt $R_L L_0 C_0$, where R_L is the radiation resistor. The input power from Port-1 is split in the SIC, a part of power is coupled to the dipole and radiated out and the rest of power is output from Port-2. Therefore, the proposed antenna combines the functions of antenna and power splitter into one device. Based on this property, the element can be readily expanded to an array antenna, which will be detailed in Part-B. Fig. 2 shows the simulated current distribution on the antenna when port-1 is excited. Evidently, the input power is split and transmitted to the dipole and Port-2, simultaneously. From the current distribution, it is observed that the output power is much weaker than the input power at Port-1, which concurs with our conjecture.

It should be noted that the L_1 and the L_4 have different lengths in Fig. 1 and thus, the antenna is asymmetrical in geometry. In this work,

Fig. 2. Simulated current distribution of the dual-port antenna element at 26 GHz.

Fig. 3. Extracted radiated power portion and transmitted power portion of the input power of the dual-port element with different L_4 .

the power is input from Port-1 and output from Port-2. By adjusting the value of L_4 , the power coupled from Port-1 to Port-2 can be altered. In order to illustrate this point, a parametric study is carried out. Fig. 3 shows the extracted radiated power portion (P_r) and the extracted transmitted power portion (P_t) with different values of the L_4 . Results show that the power portion for radiating is decreased from 0.7 to 0.3 when the L_4 increases from 0.2 mm to 1.2 mm. On the contrary, the portion of the power transmitted from Port-1 to Port-2 is increased from 0.28 to 0.58. Therefore, the ratio of the incident power for radiation and transmission can be readily controlled.

It is observed from Fig. 1(b) that there are two resonant structures in the design, *i.e.* SIC and dipole, which contribute to two independent resonant frequencies in the frequency response. Thus, the impedance bandwidth of the antenna can be enhanced by intentionally combining these two resonant frequencies together [17]. Fig. 4(a) presents the simulated S_{11} of the antenna element with different lengths of the dipole, L_d . As can be observed from the figure, the first resonant frequency is shifted to the lower frequency band when L_d is increased from 2.2 mm to 2.4 mm, while the second resonant frequency almost remains unchanged. Experimental results show that the resonant frequency of the SIC can also be adjusted by changing the length of the slots etched in the SIC. Thus, by increasing the length of the grounded CPWG (L_2), the second resonant frequency can be shifted to the lower band while the first resonant frequency is kept unchanged, as shown in Fig. 4(b).

B. End-Fire Subarray

Based on the design of the dual-port antenna element, a 1×4 subarray is formed by connecting four elements, and the configuration is shown in Fig. 5(a). The distance between two adjacent elements is set to 6 mm to reduce side-lobe level. The feed lines between the elements are meandered to ensure a consistent phase for all elements. According to the chart provided in Fig. 3, the input power can be

Fig. 4. Simulated S_{11} with different (a) lengths of the dipole, L_d ; (b) lengths of the grounded CPWG, L_2 .

Fig. 5. (a) Configuration of the series-fed end-fire subarray; (b) simulated current distribution at 26 GHz: $L_{d1} = 0.6$ mm, $L_{d2} = 1.0$ mm, $L_{d3} = 1.3$ mm.

approximately distributed among the four elements by adjusting the lengths of cut-in slots, denoted as L_{d1} , L_{d2} , and L_{d3} in Fig. 5(a). To maintain an evenly power distribution among the elements, the ratios of the radiated power to the transmitting power at the three interfaces are set 1:1, 1:2 and 1:3 from top to bottom. Thus, the lengths of the cut-in slots can be determined: $L_{d1} = 0.6$ mm, $L_{d2} = 1$ mm, and $L_{d3} = 1.3$ mm. Fig. 5(b) presents the current distribution characteristics of the subarray at 26 GHz. As can be seen, the input power is evenly distributed at the four dipoles and radiated out in-phase. Moreover, the L_6 in Fig. 1 for each element is re-optimized for better impedance matching performance since the conditions of the elements are different from each other.

Fig. 6. Simulated radiation patterns of the four-element subarray.

Fig. 7. Simulated E-plane radiation patterns of the subarray with different L_t .

Fig. 6 shows the simulated E- and H-plane radiation patterns of the subarray. As can be seen, the subarray exhibits an end-fire radiation towards the -y-axis direction. For the E-plane, the side lobes are lower than -12 dB and the cross-polarization level is below -23 dB, respectively. In the H-plane, the antenna exhibits a very wide beam with the 3-dB beam-width of over 184 degrees due to the end-fire radiation of the antenna element. Such a fan-shaped radiation pattern is very suitable for achieving the wide-angle beam-scanning phased array, which will be exemplified in Section III.

C. Beam-Tilt

Another important merit of this series-fed subarray is the beam-tilt ability, which is widely required in practical cellular applications for the coverage extension. By adjusting the length of the feeding line (denoted as L_t in Fig. 5), a constant phase delay can be easily achieved between every two adjacent elements. As a result, the main beam can be readily tilt in the E-plane. Fig. 7 shows the simulated E-plane radiation patterns with different L_t . As observed, the main beam can be steered from 90 degrees to about 110 degrees when L_t is increased from 0.5 mm to 1.5 mm. It is noted that the gain keeps almost unchanged when the antenna is steered to up 20 degrees in the E-plane.

D. Isolation

For phased array application, one important issue to be concerned is the mutual coupling between the antenna elements or subarrays. For illustration purpose, two identical subarrays are placed side-by-side with a distance of S between them, as the inset figure shown in Fig. 8. Since the two subarrays are separated physically, the mutual coupling

Fig. 8. Simulated mutual coupling of two side-by-side placed subarrays with different spacings.

behavior is different from the conventional phased array shared the same substrate, where the surface wave is always excited at the interface between the substrate and air. This inevitably leads to a strong mutual coupling when antennas are placed closely. As observed in Fig. 8, the mutual coupling is relatively low even when the distance between the subarray is reduced to 4 mm, corresponding to 0.35λ at 26 GHz. The mutual coupling in this work is mainly attributed to the electromagnetic coupling via free space rather than the surface wave propagating along the interface between the substrate and free space due to the two subarrays being separated by free space. As a result, the isolation between the subarrays can be maintained below -17 dB when S is reduced from 7 mm to 4 mm, demonstrating that the spacing is insensitive to the antenna isolation.

III. WIDE-ANGLE BEAM-SCANNING ARRAY

Owing to the advantages of wide beam-width, low mutual coupling, and compact structure, the proposed end-fire subarray is very suitable for designing the phased arrays. In this section, two antenna arrays are implemented based on the proposed subarray. One is for broadside radiation (Array-I) and the other is for beam-scanning demonstration (Array-II). These two antennas have the same configuration, but with different feeding networks to alter the steering capability.

A. Configurations

Fig. 9(a) shows the configuration of the proposed phased array, which is composed of eight subarrays with down-tilt beam and a feeding network at the bottom. In this work, two phased arrays are exemplified by adopting different power splitting networks. Each of phased array has eight subarrays oriented in the x-direction and each of unit has four radiating elements in the z-direction for gain improvement. This layout was determined to maintain a comparable beam-width in both E- and H-plane while improving the beam-scanning performance in the horizontal plane (H-plane). Besides, $Lt = 1.2$ mm is chosen to achieve the down-tilt beam. Since the subarrays have a small dimension in the x-direction and comparable high isolation between them, a small distance of 4 mm is chosen between the subarrays, which is about 0.35λ at 26 GHz. Fig. 9(b) shows the interface between the vertical subarrays and the horizontal feeding boards. To firmly connect the outputs of the power splitting networks and the inputs of the subarrays, a protrusion is first manufactured on the subarray board and several rectangular apertures are cut in the feeding board. Then, the boards for subarrays are inserted in the feeding board and soldered together.

B. Feeding Networks

Fig. 10(a)-(b) show the configurations of the two eight-way power

Fig. 9. (a) Configuration of the proposed down-tilt phased array, (b) interface between the subarrays and feeding array board.

Fig. 10. Fabricated 8-way power splitters for the phased arrays with different phase delays ($\Delta\phi$) at outputs: (a) Array-I, $\Delta\phi = 0$ degrees, (b), Array-II, $\Delta\phi = 120$ degrees.

Fig. 11. Simulated amplitudes and phase characteristics of the two power splitters: (a) amplitude, splitter-1, (b) phase, splitter-1, (c) amplitude, splitter-2, (d) phase, splitter-2.

Fig. 12. Simulated and measured S-parameters of the three array antennas.

splitters used in the phased arrays, denoted as splitter-1 and 2, respectively. The splitter-1 has the same phase delay at the outputs and the splitter-2 has a phase delay of 120 degrees at the outputs. The eight outputs are connected to the subarrays, respectively. The phase delay is realized by adjusting the lengths of the microstrip lines at the outputs. In order to reduce grating lobes, a tapered power distribution is designed purposely, which is also indicated in the figures.

Fig. 11 shows the simulated amplitudes and phases characteristics of the two power splitters. As observed, the two power splitters have the maximum output power in the center, which is gradually decreased towards both ends. In addition, the splitter-1 has no phase delay at the outputs, while the splitter-2 has a consistent phase delay of 120 degrees at the outputs in the band of interest.

IV. RESULTS AND DISCUSSION

The two phased arrays are prototyped and characterized. Fig. 12 shows the simulated and measured S-parameters of the two phased

Fig. 13. Simulated and measured normalized radiation patterns of the antenna arrays at 26 GHz: (a) E-plane, Array-1, (b) H-plane, Array-1, (c) E-plane, Array-2, (d) H-plane, Array-2.

array antennas. The measured results agree reasonably well with the simulations, exhibiting a good impedance matching performance from 24.8 GHz to 27.5 GHz. The minor discrepancy is attributed to the fabrication and measurement tolerance.

Fig. 13 shows the simulated and measured normalized radiation patterns of the two array antennas at 26 GHz. As can be seen from Fig. 13(a) and (c), the main beams are down-tilted by about 15 degrees in the E-plane and the side-lobe is lower than -10 dB. In the H-plane, Array-1 shows a broadside radiation with a sidelobe lower than -16 dB, as shown in Fig. 13(b). For Array-2, the main beam is steered to 65 degrees in the H-plane with the sidelobes suppressed to below -15 dB. The measured gains of the Array-1 and Array-2 at 26 GHz are about 18.7 dBi and 15 dBi, respectively, showing a scan loss of 3.7 dB. This improved scanning angle and the low scanning loss are attributed to the wide beam width, high isolation of the subarrays and reduced spacing between them. It should be noted that the scanning angle varies as frequency band changes. Fig. 14 shows the maximum scanning angle of the two phased arrays over a wide frequency range. As observed, the maximum beam of Array-1 maintains in the broadside direction in the band of interest. However, the maximum scanning angle of Array-2 is reduced from 80 degrees to 60 degrees as frequency increases from 24 GHz to 27 GHz. This is mainly due to the variations of the electrical dimensions and the phase delays of the power splitting network. In other words, the maximum scanning angle can be further increased by reducing the distance and increasing the phase delay between the subarrays, which, however, will lead to a higher scanning loss.

Fig. 15 presents the experimental realized gains and the radiation efficiency of the two fabricated phased arrays over from 23 to 29 GHz. As can be observed, the two proposed phased arrays have a high radiation efficiency of over 80% in the band of interest. The efficiency of the Array-2 is slightly lower than that of Array-1 due to the increased loss of the feeding network. One can also observe that the gains of Array-2 is about 14.5 dBi, which is about 3.5 dB lower than the gains of Array-1. This gain reduction is mainly caused by the scanning loss of Array-2.

Fig. 14. Variation of scanning angle of the phased arrays over a broadband.

Fig. 15. Experimental realized gains and radiation efficiency of the two antenna phased arrays.

V. CONCLUSION

In this paper, a sub-millimeter-wave antenna with end-fire radiation was proposed for 5G cellular applications. The antenna element features a dual-port structure, which combines the functions of radiation and power splitting into one device. Based on the element design, an end-fire subarray was implemented. The subarray has the advantages of wide beam-width of over 180 degrees, high isolation when closely placed, and the ability of down/up beam steering, it provides a viable solution for the phased array design. As a proof-of-concept, two phased arrays composed of eight subarrays have been developed and characterized. The results showed that the antenna has a low sidelobe level, and wide beam-scanning ability up to 65 degrees with a low scanning loss of 3.7 dB and a low side-lobe level (SLL) of -15 dB. These promising features make the proposed mm-Wave end-fire antenna very suitable for 5G cellular mm-Wave applications.

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