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Wideband Double Dipole Quasi-Yagi Antenna Using a Microstrip-to-Slotline Transition Feed

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Abstract

This paper describes a wideband double dipole quasi-Yagi antenna fed by a microstrip-to-slotline transition. The transition feed consists of a microstrip radial stub and a slot radial stub, each with the same angle of 90° but with different radii, to achieve wideband impedance matching. Double dipoles with different lengths are utilized as primary radiation elements to enhance bandwidth and achieve stable radiation patterns. The proposed antenna has a measured bandwidth of 3.34～8.72 GHz for a −10 dB reflection coefficient and a flat gain of 6.9±0.6 dBi across the bandwidth.

Key words: Slotline Transition, Double Dipole, Tapered Coplanar Stripline.

Ⅰ. Introduction

Quasi-Yagi antennas are commonly used for many applications in the microwave and millimeter-wave bands because of their broad bandwidth, high gain, low cost, and high-radiation efficiency, as well as their ease of fabrication. These antennas can be fed by several types of feedlines, including microstrip lines (MS) [1]～[6], coplanar waveguides (CPW) [7], [8], coplanar striplines (CPS) [9], [10], or slotlines [11]. Quasi-Yagi antennas utilize a regular dipole as the driver, which means that the bandwidths are approximately 50 % or less and may not be sufficient for some applications. Eldek enhanced the bandwidth by introducing MS-fed planar antennas with double dipoles [12] and double rhombuses [13] as the main radiation elements. These antennas achieved a wide bandwidth with two parallel strip feedlines printed on opposite sides of the substrate. However, the radiation patterns were degraded due to the asymmetric structure of the antennas.

This paper presents a microstrip-to-slotline transition-fed quasi-Yagi antenna with wide bandwidth and flat gain. The transition consists of a microstrip radial stub and a slot radial stub, each with the same 90° angle, but with different radii for impedance matching between the microstrip line and the slotline [14]. The bandwidth is enhanced by replacing the regular dipole driver with two parallel dipoles of different lengths. These double dipoles are connected to the slotline by a coplanar stripline, which is tapered to improve the impedance matching conditions. The antenna incorporates two parasitic strips as directors, and a truncated ground plane as the reflector, to achieve small gain variation across the operating bandwidth. Compared with the double dipoleYagi-Uda antenna [15], the presented antenna has a more compact and simpler structure and a wider impedance bandwidth.

Ⅱ. Antenna Geometry

Fig. 1 shows the geometry of the wideband microstrip-to-slotline transition-fed quasi-Yagi antenna. The antenna was designed on a 60×70 mm RT Duroid 6010 substrate with a dielectric constant of 10.2, a loss tangent of 0.0023, and a thickness of 0.635 mm. It consists of a microstrip-to-slotline transition as the feed, two parallel dipoles as the main radiation elements, two printed strips as the directors, and a ground plane as the reflector. The microstripline was designed on the back side of the substrate and had a characteristic impedance of 50 Ω. The slotline was designed with a characteristic impedance of approximately 70 Ω, due to a compromise between the narrow slot width required to obtain a characteristic impedance of 50 Ω and the limitations of the available fabrication technique. Therefore, a microstrip radial stub and a slot radial stub, each having the

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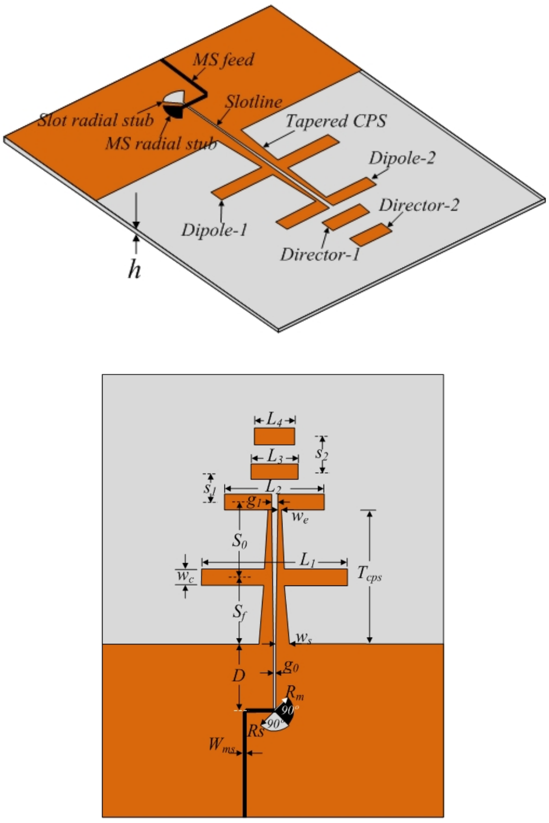


Fig. 1. Geometry of the wideband double dipole quasi-Yagi antenna.

same angle of 90° but with different radii, were inserted into the transition to provide impedance matching between the microstrip-line and the slotline. Two parallel dipoles acted as the primary radiation elements, which were directly connected to the slotline by a coplanar stripline. The initial lengths of dipole-1 and dipole-2 were approximately one-half of the effective wavelengths ($\lambda_{eff}/2$) at 4 and 6.5 GHz, respectively. However, the input impedance of the double dipoles is higher than the characteristic impedance of the slotline and the input impedances at each dipole are different. Therefore, the coplanar stripline was shaped from a wide-line width with a smallgap to a narrow-line width with a largegap to improve the impedance matching condition. Additionally, two dipoles were arranged to minimize the gain variation in the low-frequency region (3.5~6.5 GHz) with the dipole-1 to reflector (ground plane) spacing of S_f and the dipole-1 to dipole-2 spacing of S_0. The directors were designed to have a resonance near 8.5 GHz and to achieve an optimal phasing for a small gain variation in the high-frequency region (6.5~9 GHz). The antenna characteristics were investigated with an Ansoft high-frequency structure simulator (HFSS) and the optimized antenna design parameters are given in Table 1. The next section is a parametric study that shows the

Table 1. Optimized antenna design parameters.

Parameter	Value (mm)	Parameter	Value (mm)
D	10	S_f	10
g_0	0.2	S_0	11.2
g_1	0.8	S_1	4.4
L_1	22	S_2	5.4
L_2	16	W_{ms}	0.56
L_3	7	W_s	2
L_4	6	W_e	0.6
R_m	2.8	W_c	2.4
R_s	3	h	0.635

effects of several main design parameters on the antenna characteristics.

Ⅲ. Parametric Study

As mentioned before, the key motivation for the use of two parallel dipoles is to achieve multi-resonances, and consequently, to further enhance the bandwidth. This is clearly observed in Fig. 2, which shows the simulated reflection coefficient of the antenna with/without dipole-1 (L_1). In the absence of dipole-1, the antenna exhibited three-resonances around 6.6, 7.4, and 8.4 GHz, and the bandwidth was 6.55~8.75 GHz for a −10 dB reflection coefficient. In the presence of dipole-1, the antenna showed five-resonances around 3.8, 4.5, 6.3, 7.3, and 8.4 GHz, and the bandwidth was 3.65~8.90 GHz for the −10 dB reflection coefficients. The impedance bandwidth of the microstrip-to-slotline transition fed quasi-Yagi antenna was enhanced from 29 % to 83.7 % by the use of double dipoles as the primary

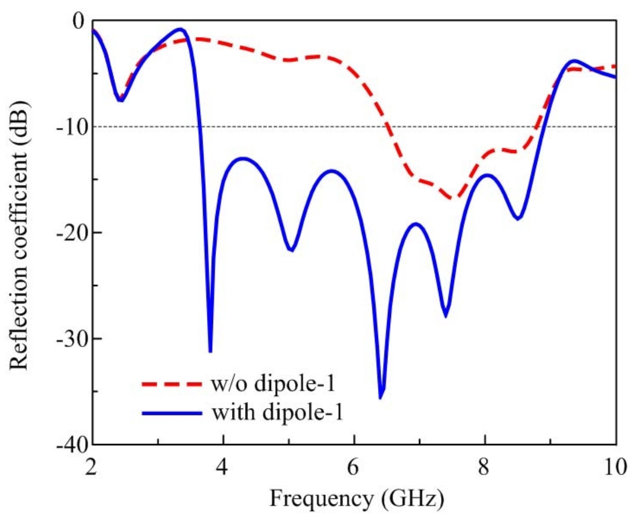


Fig. 2. Simulated reflection coefficient of the antenna with/without dipole-1.

radiation elements.

Fig. 3 shows the simulated reflection coefficient as a function of frequency for different lengths of dipole-1 (L_1). As L_1 was increased from 20 mm to 24 mm in 2 mm steps, the lowest resonant frequency decreased while other resonances changed only slightly. Fig. 4 shows the simulated reflection coefficient as a function of frequency for different lengths of the dipole-2 (L_2). As L_2 was increased from 14 mm to 18 mm in 2-mm-steps, the third resonance (near 6.3 GHz) decreased but other resonances hardly changed. Fig. 5 shows the antenna reflection coefficient as a function of frequency for different lengths of the director-1 (L_3). As L_3 was increased from 6 mm to 8 mm in 1 mm steps, the resonances in the high-frequency region decreased, while the resonances in the low-frequency region remained almost the same. These results indicate that the lengths of $L_{1\sim3}$ mainly determine the resonances of the antenna. The effect of these parameters on the antenna gain is negligible and is not shown here.

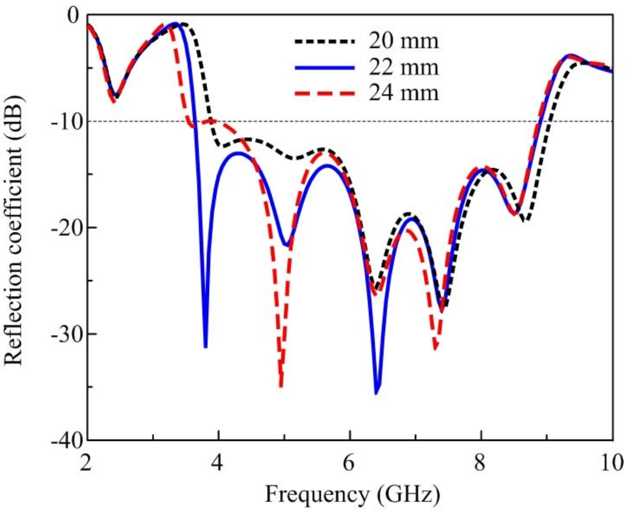


Fig. 3. Simulated reflection coefficient as a function of the dipole-1 length (L_1).

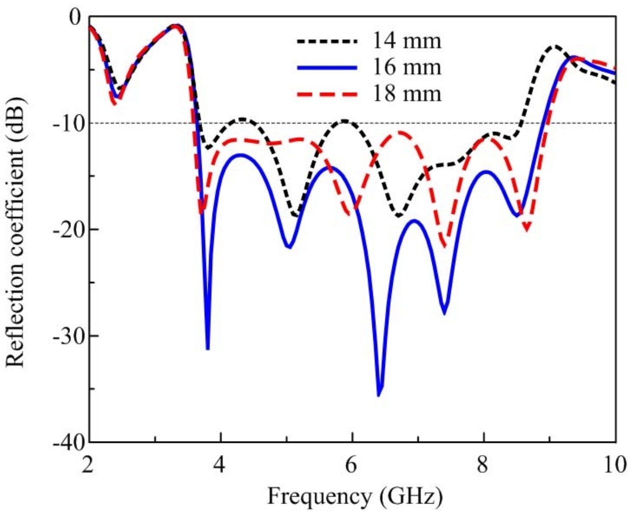


Fig. 4. Simulated reflection coefficient as a function of the dipole-2 length (L_2).

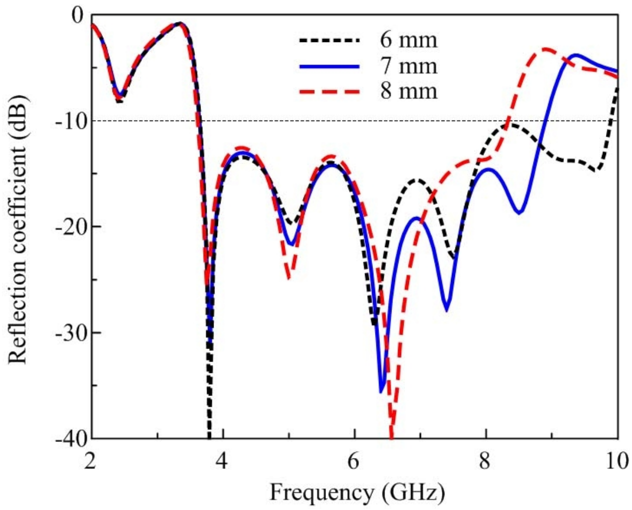


Fig. 5. Simulated reflection coefficient as a function of the director-1 length (L_3).

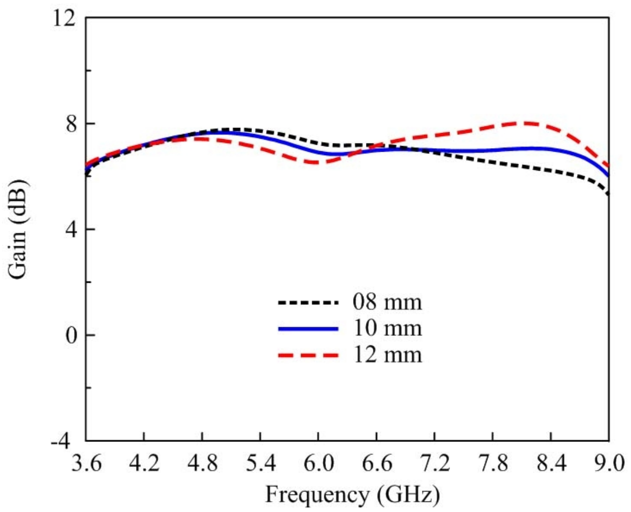


Fig. 6. Simulated gain as a function of the space between the ground plane and double dipoles (S_f).

This study also investigated the antenna design parameters that negligibly affect the reflection coefficient (not shown), but strongly influence the gain. Fig. 6 shows the simulated gain as a function of frequency for different spacings between the ground plane and longer dipole (S_f). As S_f was varied from 8 mm to 12 mm in 2-mm-steps, the gain degraded in the low-frequency region, but improved in the high-frequency region. Fig. 7 shows the simulated gain as a function of frequency for different spacing between two parallel dipoles (S_0).

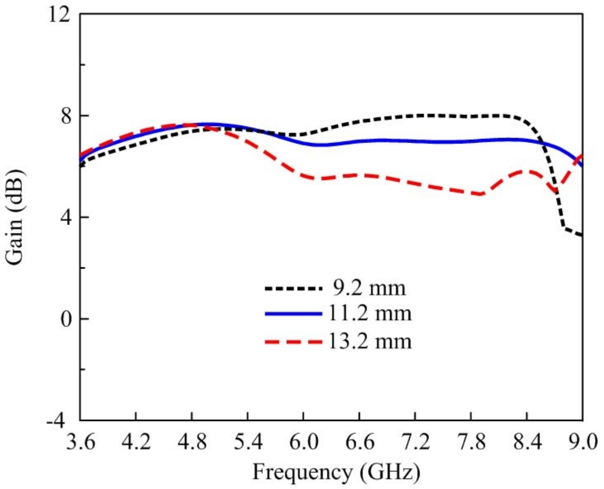


Fig. 7. Simulated gain as a function of the space between two parallel dipoles (S_0).

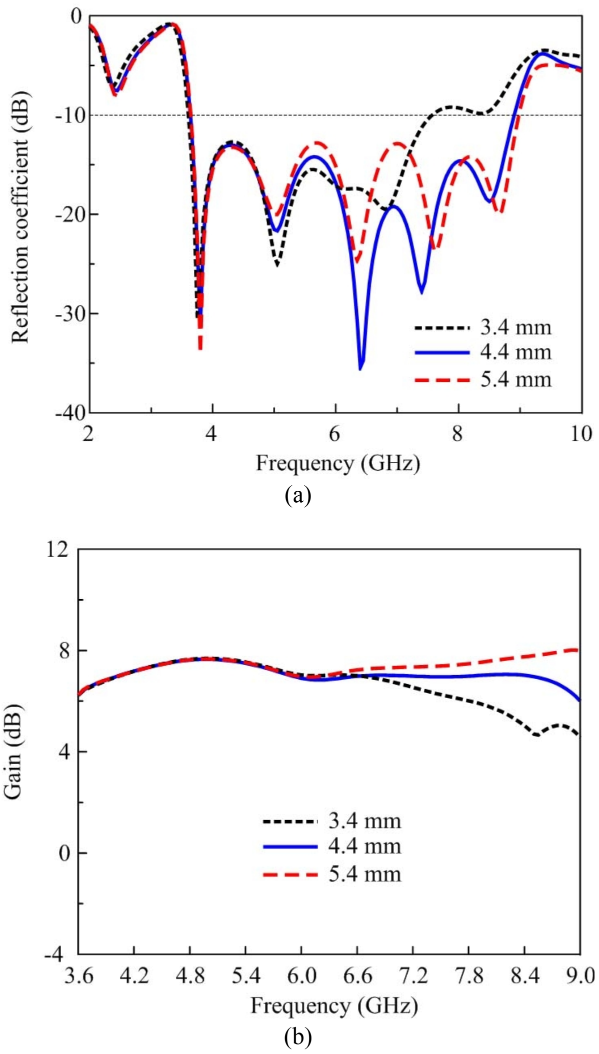


Fig. 8. (a) Simulated reflection coefficient and (b) gain as a function of the spacing between the double dipoles and director (S_1).

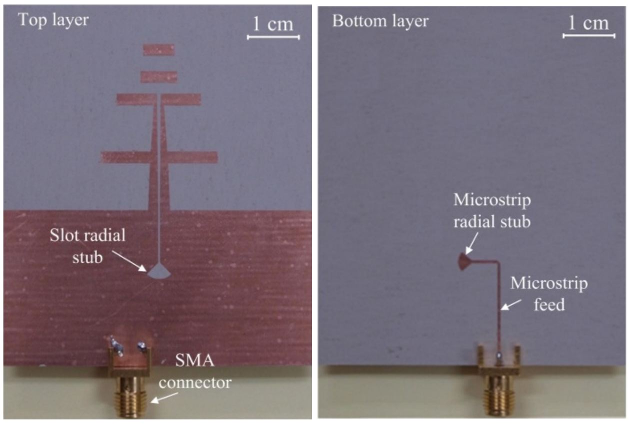
As S_0 was varied from 9.2 mm to 13.2 mm in 2 mm steps, the gain slightly improved in the low-frequency region (3.6~4.8 GHz), but was significantly degraded in the high-frequency region. These results indicate that the gain of the proposed antenna can be controlled by adjusting S_f and S_0.

Fig. 8 shows the variations in the antenna characteristics as a function of frequency for different spacings between the shorter dipole and director-1 (S_1). Increments of 1 mm in this space, from 3.4 mm to 5.4 mm, induced significant changes in the reflection coefficient and gain in the high-frequency region but negligible changes in low-frequency region. This indicates that the space between the parallel dipoles and the directors mainly affects the antenna characteristics in the high-frequency region.

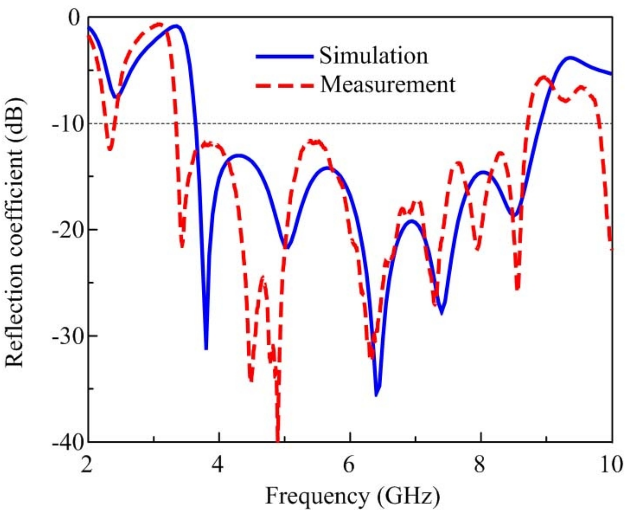
Ⅳ. Simulation and Measurement Results

The antenna was fabricated on both sides of an RT/Duroid 6010 substrate with a 20-μm copper thickness, and a subminiature type-A (SMA) connector was used as a microstrip-to-coaxial line transition (not included in HFSS simulations). An Agilent N5230A network analyzer and an Agilent 3.5-mm 85052B calibration kit were used for measurements of the prototype [Fig. 9(a)]. As shown in Fig. 9(b), the measured and simulated reflection coefficients of the antenna agreed rather closely. The measured bandwidth was 3.34~8.72 GHz for the −10 dB reflection coefficient while the simulated bandwidth was 3.65~8.9 GHz. The slight difference between the measured and simulated results could be attributed to a misalignment at the transition and the effect of the SMA connector.

The normalized radiation patterns of the antenna at 4, 6, and 8 GHz are illustrated in Fig. 10, and showed a good agreement between the measurements and simulations. The radiation patterns are symmetric and stable, with a front-to-back ratio and cross-polarization level better than 15 and −17 dB, respectively. At a frequency of 4 GHz, the measurements resulted in a half power beamwidth (HPBW) of 83° and 113° along the E- and H-planes, respectively. At 6 GHz, the measurements resulted in a HPBW of 92° along both E- and H-planes. At 8 GHz, the measurements resulted in a HPBW of 83° and 77° along the E- and H-plane, respectively. As shown in Fig. 11, the measured gain of the antenna was 6.3~7.5 dBi across the bandwidth, which agreed closely with the simulated gain of 6.4~7.6 dBi. Some ripples in the measured results could be attributed to scattering effects arising from the measurement setup. The antenna also exhibited a relatively small gain variation (±0.6 dB)



(a)



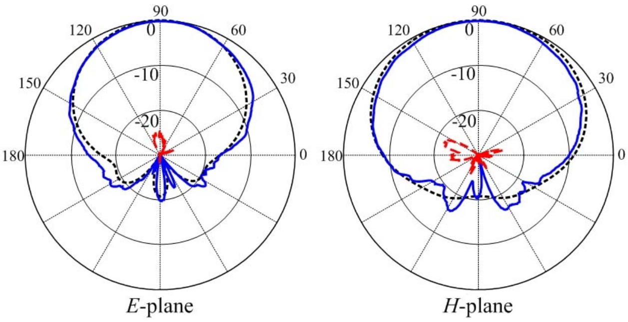
(b)

Fig. 9. (a) Fabricated wideband double dipole microstrip-to-slotline transition fed quasi-Yagi antenna, (b) measured and simulated reflection coefficient.

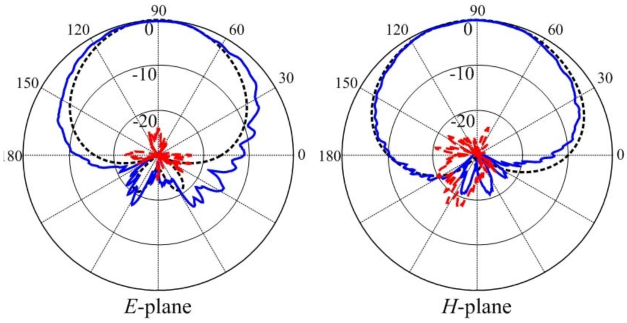
at 3.6～9 GHz; thus, it can be used for transmitting or receiving applications in wideband wireless communication systems.

Ⅴ. Conclusion

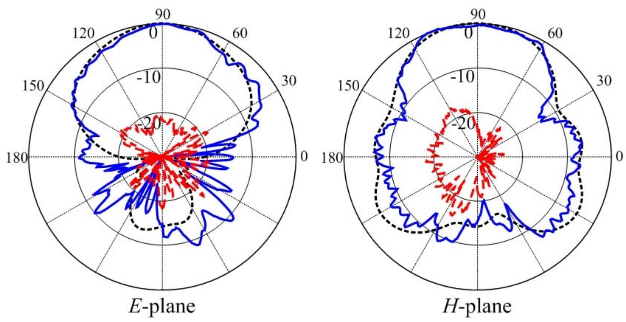
This paper introduced a microstrip-to-slotline transition-fed quasi-Yagi antenna with a wide bandwidth and a flat gain. Wideband characteristics and stable radiation patterns were achieved by replacing the regular dipole driver with two parallel dipoles of different lengths, which were directly connected to the slotline by a tapered coplanar stripline. The planar structure, stable radiation pattern, wideband characteristics (3.34～8.72 GHz), small gain variation (6.9±0.6 dBi), low cross-polarizations (<－17 dB), and high front-to-back ratio (>15 dB) of the proposed antenna make it widely applicable to wideband wireless communication systems.



(a) 4 GHz



(b) 6 GHz



(c) 8 GHz

Fig. 10. Normalized radiation patterns of the antenna: ------ co-polarization simulation, —— co-polarization measurement, and - - - - cross-polarization measurement. Simulated cross-polarization is excluded because value is too small to show in Figure.

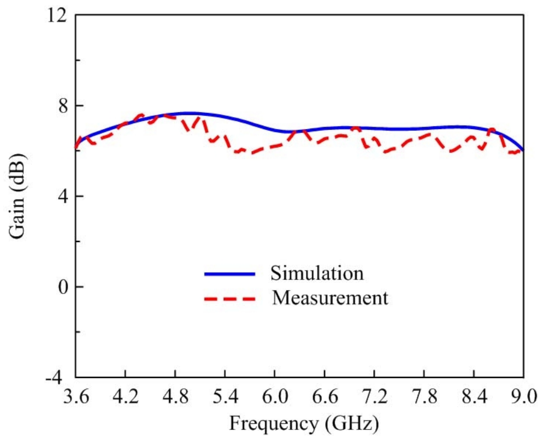


Fig. 11. Gain of the optimized antenna.

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