Research Article

Design of a Small-Size Broadband Circularly Polarized Microstrip Antenna Array

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A small-size 2×2 broadband circularly polarized microstrip antenna array is proposed in this article. The array has four broadband dual-feed U-slot patch antenna elements with circular polarization, and the sequential feeding technique is used to further enhance the 3 dB axial ratio bandwidth. The lateral size of the fabricated array is as small as $1.33\lambda_0 \times 1.33\lambda_0$, and the profile is only $0.04\lambda_0$. Measured results show that the overlapped -10 dB reflection coefficient and the 3 dB AR bandwidth is 53%, and the variation of the measured realized gain is less than 1 dB for S-band satellite communications (1.98–2.2 GHz).

1. Introduction

Circularly polarized (CP) microstrip antennas and arrays are widely used in wireless communications due to their insensitivity to polarization mismatch and multipath effects. Although CP microstrip arrays (MSAs) have many merits, such as low profile, lightweight, and easy conformability, they inherently have narrow bandwidth in terms of impedance and axial ratio (AR). Wideband and compact designs of CP MSAs are increasingly demanded in modern communication systems.

The sequential feeding technique (SFT) has been proven to be an effective way to improve the impedance and AR bandwidth of CP MSAs [1, 2]. Several designs of 2×2 CP MSAs based on the SFT are presented [3–10]. In [3–5], single-layer single-feed CP microstrip antenna elements are used and the sequential feeding networks are embedded in the middle of the arrays to reduce the overall sizes. These MSAs achieve more than 18% -10 dB reflection coefficient ($|S_{11}|$) bandwidth, and the 3 dB AR bandwidth is over 12.7%. At the expense of complex feeding network and an extra metallic reflector, an MSA consisting of dual-feed slot-coupled CP elements obtains 50% -10 dB $|S_{11}|$ and 30%

3 dB AR bandwidths in [6]. In [7, 8], linearly or circularly polarized patch elements are fed by sequentially rotated L -probe feeding networks, and the 3 dB AR bandwidth exceeds 40%. However, the lateral sizes of these arrays exceed $1.8\lambda_0 \times 1.8\lambda_0$, and the minimum height is also larger than $0.15\lambda_0$. In [9], a stacked microstrip antenna array with size of $1.46\lambda_0 \times 1.33\lambda_0 \times 0.11\lambda_0$ is presented and 51% -10 dB $|S_{11}|$ and 27% 3 dB AR bandwidths are achieved. In [10], each array element is composed of four corner-truncated square patch and U-shaped slots are embedded in the ground plane; 54.2% -10 dB $|S_{11}|$ and 39.3% 3 dB AR bandwidths are achieved with a relatively large height of $0.4\lambda_0$.

In this article, a 2×2 broadband CP MSA with a small size is designed. The structure of a wideband CP dual-feed U-slot patch antenna recently reported in [11] is modified to lower the profile and is used to construct the MSA, and a feeding network using the SFT is adopted to further enhance the polarization purity and AR bandwidth. The design principles and the related numerical results are validated by the measured ones on the fabricated prototype, and wide overlapped impedance and AR bandwidth, small lateral size, and low profile of the 2×2 array are achieved.

FIGURE 1: Structures of the antenna element. (a) Geometry of the antenna element. (b) Side view of the antenna element. Dimensions of the antenna element (unit: mm): $W = 95$, $W_p = 50.5$, $L_s = 21$, $S_s = 20$, $W_s = 1$, $d_1 = 1.8$, $d_2 = 10.2$, $t = 6.75$, $h = 1.016$, and $h_0 = 4$ mm.

2. Design of the Antenna Element and Its Performance

The geometry of the antenna element and the layout of the MSA are illustrated in Figure 1. The antenna element structure is adapted from the design proposed in [11]. In [11], the antenna was printed on a thick high-permittivity substrate to meet the requirements of maximal size reduction in lateral direction, while the bandwidth and the profile are not the most concerned properties. In the present design, to enlarge $|S_{11}|$ and AR bandwidths and to reduce the antenna's profile, we use two dielectric substrates with the same thickness of h and with the relatively low permittivity of $\epsilon_r = 3.5$ and loss tangent of $\tan \delta = 0.0018$. As shown in the two figures, the dual-feed U-slot patch antenna elements are printed on the top substrate and the feeding network resides on the bottom one. The two substrates are separated by an air gap of height h_0 . The width of the square patch is W_p , and the horizontal and vertical arm lengths of the etched U-shaped slot are L_s and S_s , respectively. The width of the slot is W_s , and the distance from the patch center to the vertical arm is t . The antenna element is fed by two metallic pins that are connected to the two output ports of a microstrip Wilkinson power divider with the power ratio of 1:1, and the distances of the two pins measured from the patch center are d_1 and d_2 , respectively. It should be noted that the phase difference between the two feeding ports is not 90° as for the commonly used dual-feed CP antennas, and its value is tuned to about 60° due to the presence of the embedded U-slot.

The performances of the element patch antenna are shown in Figure 2. As can be seen in Figure 2(a), excellent impedance matching at two feeding ports (with the reference impedance 50Ω) is observed at two series resonant frequencies 2 GHz and 2.1 GHz, and the joint impedance bandwidth is 10% (1.95~2.15 GHz). Meanwhile, the isolation between two ports is more than 20 dB over a wide bandwidth, which indicates a weak mutual coupling between the two feedings. Figure 2(b) shows the realized gain and

AR when the two feeding ports (*feed_1* and *feed_2*) of the element patch antenna are excited simultaneously and respectively. The peak realized LHCP gain with simultaneous excitation is 8.2 dBic, which is almost equal to the realized linearly polarized (LP) gain with respective excitation. The 3 dB AR bandwidth with simultaneous excitation is 22% (1.86~2.32 GHz).

3. Design of the Array and Parameter Studies

In Figures 3(a) and 3(b), the 2×2 array is constructed by 4 antenna elements (indexed by A_1 to A_4), and the SFT is implemented by using a three-stage feeding network composed of 7 Wilkinson power dividers to improve the S_{11} and AR bandwidths. For the left-handed circular polarization (LHCP) design, the 4 antenna elements rotate clockwise from A_1 to A_4 , and the feeding phase at the input port of the third-stage power divider before each element is 90° lagged as compared to that for the former element. The center distance between antenna elements is $0.65\lambda_0$, and the overall size of the array is about $1.33\lambda_0 \times 1.33\lambda_0 \times 0.04\lambda_0$ at the center frequency of 2.1 GHz.

In order to optimize the performance of the proposed element and MSA, parametric studies are carried out. The Wilkinson power divider is designed for the element to achieve the circular polarization. In what follows, the effects of several key dimensions on the element and array are investigated, and the final optimum geometrical sizes of the antenna element and the array are listed in the captions of Figures 1 and 3.

4. Effects of the Length of the Patch Element

PAUSE Figure 4 shows the variations of the realized gain, AR, and $|S_{11}|$ of the element and array versus the frequencies when W_p is changed. When W_p is increased, the frequencies at which both the peak gain and the minimal AR appear are shifted downward, as seen in Figure 4(a). The lowest

FIGURE 2: Performances of the patch element. (a) S-parameters. (b) Realized gain and AR.

FIGURE 3: Structures of the corresponding 2×2 array. (a) Top view of the array. (b) Side view of the array. Dimensions of the array (unit: mm): $G = 190$.

resonant frequency is also lowered with increasing W_p , but the -10 dB impedance bandwidth remains almost unchanged. Although the AR of the array is insensitive to the variation of W_p and the $|S_{11}|$ is always less than -10 dB from 1.6 to 2.5 GHz, the frequency at which the maximal

realized gain appears is lowered (see Figure 4(c)). Moreover, as seen in Figure 4(d), the lowest and highest resonant frequencies in the $|S_{11}|$ curve are reduced with the increasing W_p , while other resonances remain almost unchanged. These results indicate that the intermediate resonant frequencies

FIGURE 4: Effects of the length of the patch element. (a) Realized gain and axial ratio of the element. (b) Reflection coefficient of the element. (c) Realized gain and axial ratio of the array. (d) Reflection coefficient of the array.

determined by the embedded U-slot are quite independent of W_p .

5. Effects of the Length of the U-Slot's Horizontal Arm

Figure 5 illustrates the effects of the horizontal arm length L_s of the U-slot on the performance of the element and array. For both the element and array, when L_s deviates from the optimum value of 20 mm, the peak realized gain will decrease

to 1.5 dB or so, as illustrated in Figures 5(a) and 5(c). When L_s is reduced, the AR of the element deteriorates and improves at the low and high ends of the frequency range, respectively. The AR of the array deteriorates at both the low and high ends of the frequency range. In addition, it is seen that all the resonant frequencies of the element and array change with the varying L_s , as indicated in Figures 5(b) and 5(d). These observations indicate that due to the variation of the slot's length, the phase difference between the two output ports of each power divider cannot provide the optimum value for circular polarization of the

FIGURE 5: Effects of the length of the U-slot's horizontal arm. (a) Realized gain and axial ratio of the element. (b) Reflection coefficient of the element. (c) Realized gain and axial ratio of the array. (d) Reflection coefficient of the array.

antenna element any more, which leads to the gain reduction of the array. So, the dimensions of the third-stage power dividers should be tuned alongside with the change of L_s to provide appropriate phase difference for the CP antenna elements.

6. Effects of the Length of the U-Slot's Vertical Arm

Figure 6 shows the performance variations of the element and array versus the frequencies when S_s is changed. As can be seen in Figure 6(a), when S_s is increased, the peak realized gain of

the element is slightly lowered and the AR at the low frequency range is reduced. Although the high-resonant frequency is lowered with increasing S_s , the -10 dB impedance remains almost unchanged, as shown in Figure 6(b). Figure 6(c) shows that the AR is insensitive to the variations of S_s from 1.6 GHz to 2.5 GHz, and the frequency at which the maximal realized gain appears is slightly lowered and then remains stable when S_s increases to the optimum value of 15.4 mm. In Figure 6(d), the intermediate and highest resonant frequencies depart from each other with the increasing S_s . So, it is evident that S_s affects the intermediate and highest resonant frequencies, as well as the impedance bandwidth, of the element and array.

FIGURE 6: Effects of the length of the U-slot's vertical arm. (a) Realized gain and axial ratio of the element. (b) Reflection coefficient of the element. (c) Realized gain and axial ratio of the array. (d) Reflection coefficient of the array.

7. Effects of the Excitation Modes of the Array

The layouts of the arrays using only *feed_1* or *feed_2* are shown in Figure 7. When the array is only excited by *feed_1*, *feed_2* is connected with matching load ($50\ \Omega$) and vice versa. The three-stage power divider is replaced by a two-stage power divider to ensure the 90° phase difference of SFT. The other dimensions of the array are listed in the caption of Figure 1. The gain, AR, and reflection coefficient of the array with simultaneous and respective excitations are illustrated in Figure 8. The frequencies at which the peak gain appear are shifted slightly, and the peak gains using only

feed_1 or *feed_2* are reduced around 3.7 dB compared with the one simultaneously using two feeds, as seen in Figure 8(a). Besides, the AR deteriorates in the working frequency range using only *feed_1* or *feed_2*. The reflection coefficients of the array using simultaneous and respective excitations are shown in Figure 8(b). Although the all the resonant frequencies are changed, the reflection coefficient still remains less than -10 dB from 1.6 to 2.5 GHz. These results indicate that the gain and circular polarization of the array with simultaneously using the two feeds are enhanced significantly, compared with the case of using only *feed_1* or *feed_2*.

FIGURE 7: Respective excitations for the array: (a) Only *feed_1* is excited; (b) only *feed_2* is excited.

FIGURE 8: Effects of different excitation modes for the array. (a) Realized gain and axial ratio. (b) Reflection coefficient.

FIGURE 9: The fabricated prototype of the antenna array.

8. Experimental Results

A 2×2 MSA is fabricated, and the prototype is shown in Figure 9. The two dielectric substrates are fixed together by screws, and four hollow pillars are used to ensure the height

FIGURE 10: Simulated and measured reflection coefficients of the proposed array.

FIGURE 11: Simulated and measured radiation patterns and axial ratios of the array. (a) 1.6 GHz. (b) 2.1 GHz. (c) 2.6 GHz.

FIGURE 12: (a, b) Simulated and measured axial ratio and realized gain of the array.

TABLE 1: Measured performance comparison of the proposed design and the previous 2×2 CP arrays.

Array	Overall sizes (λ_0^3)	-10 dB $ S_{11} $ BW (%)	3 dB AR BW (%)	3 dB gain BW (%)	Peak gain (dBic)	Global bandwidth (%)
Proposed	$1.33 \times 1.33 \times 0.04$	75	53	20.8	12.3	20.8
[4]	$1.50 \times 1.50 \times 0.06$	20.8	17.6	13	11.5	13
[5]	$1.82 \times 1.82 \times 0.03$	18	12.7	16.6	12	12.7
[6]	$1.72 \times 1.72 \times 0.44$	48	30	30	12.8	30
[7]	$1.80 \times 1.80 \times 0.16$	46.6	40	40	10.9	40
[8]	$1.84 \times 1.84 \times 0.15$	46.8	45	47	10.5	45
[9]	$1.46 \times 1.33 \times 0.11$	51	27	27	12.7	27
[10]	$1.53 \times 1.53 \times 0.05$	54.2	39.3	57	11.3	39.3

of the air gap. Comparison of the measured and simulated $|S_{11}|$ of the array is shown in Figure 10. The measured reflection coefficients agree very well with the simulated ones, and the -10 dB $|S_{11}|$ bandwidth is about 75% (1.25–2.75 GHz). The radiation patterns at 1.6, 2.1, and 2.6 GHz in the $\varphi = 0^\circ$, $\varphi = 45^\circ$, and $\varphi = 90^\circ$ planes are shown in Figure 11, in which the experimental results show good agreement with the numerical predictions. The corresponding 3 dB beamwidths in the three planes are 46.8° , 45.45° , and 44.75° at 1.6 GHz, 38.25° , 38.10° , and 37.55° at 2.1 GHz, and 30.4° , 29.75° , and 28.85° at 2.6 GHz, respectively, which suggests good symmetries in the measured radiation patterns of the proposed array.

As seen in Figure 12(a), the measured realized gains are less than the simulated ones by 1 dB or so, which may be caused by the uncertainties in the measurements and by the losses in the feeding network and other parts of the array as well. The measured peak realized gain of the array is 12.3 dBic at 2.07 GHz, and the realized gain is quite stable within the frequency range of 1.98–2.2 GHz that is designated for S-band satellite communications. The fractional 3 dB gain bandwidth is

20.8%, ranging from 1.94 to 2.39 GHz with the center frequency being 2.09 GHz. The realized gain bandwidth is comparable with that presented in [6], but the profile of the proposed design is only one-tenth of the thickness of the array in [6]. Although nearly 40% bandwidth is achieved in [8], the peak realized gain of that array is 1.8 dB lower than our design, and its occupied volume is 7 times larger than that of the presented one. Since the product of the gain and the bandwidth would be approximately a constant for an array, the measured results are reasonable. The simulated efficiency of the array is shown in Figure 12(a), which is greater than 42% within the 3 dB gain bandwidth. Figure 12(b) illustrates the measured and simulated AR, and good agreement and LHCP performance are observed. The measured 3 dB AR bandwidth is 53% (1.54–2.65 GHz), which is entirely covered by the -10 dB $|S_{11}|$ bandwidth.

Comparisons of the overall performances of the proposed MSA with those of other referenced 2×2 arrays are listed in Table 1. As shown in the table, when compared with the arrays in [4, 5], the proposed array provides wider global

bandwidth (20.8%, 1.94–2.39 GHz) and higher peak gain (12.3 dBic) with smaller occupied volume, even though the volumes of the array in [4, 5] are 1.91 times and 1.41 times of the proposed one, respectively. Although the global bandwidth is 30% in [6]; however, the profile of the proposed design is only 1/10 of the thickness and almost 1/20 of the volume of the array in [6], respectively, and the peak realized gains are comparable. Although nearly 40% global bandwidth is achieved in [7, 8], the peak realized gains of that arrays are 1.4 dB lower than our design, and their occupied volumes are 7 times larger than that of the proposed one. Although the global bandwidth array is 27% in [9], one more substrate and extra foam layers are needed compared with the structure of the proposed array, which means that the proposed array owns relatively low cost and simpler construction. The global bandwidth is 39% in [10], but the proposed array provides higher gain with smaller occupied volume. These comparisons indicate that there exists the compromise between radiation efficiency and antenna size, and the reduction in radiation efficiency with smaller size is inevitable. There are some references with a much wider 3 dB bandwidth gain, but these antennas always have relatively large sizes in both lateral and lengthways directions. Our main concern is focused on the reduction of size of the array antenna while maintaining good electrical performances. Compared with the existing references of compact array antennas, especially with [4, 5], the proposed array antenna provides better electrical performances with smaller occupied volume.

9. Conclusion

In this article, a 2×2 MSA consisting of dual-feed wideband CP U-slot patch antenna elements is presented. The overall size of the array is as small as $1.33\lambda_0 \times 1.33\lambda_0 \times 0.04\lambda_0$ corresponding to the center frequency of 2.1 GHz, and the -10 dB $|S_{11}|$ and 3 dB AR bandwidths reach 75% (1.25–2.75 GHz) and 53% (1.54–2.65 GHz), respectively, which means 53% usable overlapped bandwidth. The measured peak realized gain is 12.3 dBic, and the realized gain is stable within the frequency range of satellite communications at S-band. Compared with other designs reported in [4–8], the proposed MSA possesses wider overlapped impedance and AR bandwidth, smaller lateral size, and lower profile. Being wideband performance and small in size, the proposed array is promising in its applications for modern wireless communication systems.

Data Availability

The dimensions of the array used to support the findings of this study are included within the article.

Conflicts of Interest

The authors declare that there is no conflict of interest regarding the publication of this paper.

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