A Novel Low-Profile Phased Antenna with Dual-Port and Its Application in 1-D Linear Array to 2-D Scanning

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Abstract—A novel low-profile phased antenna with dual-port and its two types of 1-D linear array to 2-D space wide-angle beam scanning are proposed. The antenna unit is composed of fishbone structure and rhombus parasitic patch, which is folded into n-shaped to low profile 0.08λ. By adjusting the phase difference (Δφ) from 0° to 180°, the antenna working mode is changed from even-mode to hybrid-mode, and to odd-mode, eventually. Therefore, the proposed antenna unit realizes wide-angle scanning characteristics, and continuous switching state from broadside lobe to end-fire lobe. In further, the proposed phased antenna element is extended to two kinds of 1-D linear phased array (Array#1 and Array#2) based on the generalized principle of the pattern multiplication. The fabricated Array#1 can realize beam scanning range -116° < θn < 0° (xoz-plane), 0° < θn < 63° (yoz-plane), and 90° < φm < 169° (xoy-plane). And the fabricated Array#2 can achieve 0° < θn < 60° (xoz-plane), 0° < θn < 65° (yoz-plane), and 90° < φm < 152° (xoy-plane). The proposed phased antenna unit and its 1-D array can realize 2-D space wide-angle scanning, improving scanning performance and expanding scanning dimensional space in phased array.

Index Terms—1-D linear array, 2-D space wide-angle scanning, Phased antenna unit, Phased array, the generalized principle of the pattern multiplication (GPPM).

I. INTRODUCTION

Phased array have broad application prospects in wireless communication technology, such as system capacity increasement, spectrum efficiency improvement, channel anti-interference ability enhancement, and beamforming. In particular, its flexible continuous beam scanning and high gain characteristics are widely used in wide-angle scanning such as radar, satellite, wireless communication systems, and other application scenarios, especially in high-speed moving equipment, which has an urgent need for the phased array with low-profile, high-gain, and multi-dimensional space scanning characteristics [1]-[5].

Generally, the phased array has the advantages of wide-angle scanning and high gain [6]-[7]. Nevertheless, it has some problems. For example, the mutual coupling among antenna unit limits the scanning range, the active impedance mismatch leads to gain roll-off in the process of wide-angle scanning, and the traditional linear array only realizes beam scanning characteristic in one-dimensional (1-D) space [8]-[12].

In general, the wide-beam antenna unit can effectively improve the wide-angle scanning performance of the phased array [13]-[17]. In the previews report [18], the planar linear array based on the wide-beam antenna unit could realize the wide-angle scanning characteristics from the broadside direction to the end-fire direction (-89°– 90° in H-plane).

Usually, the conventional linear array can only scan in 1-D space, and two-dimensional (2-D) space scanning can be realized by 2-D array or multi-dimensional array. That is to say, it is a challenge to realize high radiation performance in the finite space of the wireless systems.

The antenna unit based on the reconfigurable technology can adjust the surface current distribution, to exchange its resonance mode and obtain the multi-beam characteristic. Actually, the switching characteristics of the reconfigurable network mainly depend on the on-state or off-state of the PIN diodes. And the characteristics of the PIN diode affect switching time and electromagnetic signature of the reconfigurable network, which limit the scanning performance of the reconfigurable antenna [19]-[21], and may worsen the radiation performance of the antenna. Furthermore, the complex reconfigurable network takes up lots of the system spaces, which brings the problems of the system miniaturization and noise [22]-[23].

In recent years, the leaky-wave antennas have attracted attention due to low cost, frequency beam scanning [24]-[27], and compact structure. In general, the electromagnetic energy of the leaky-wave antenna propagates in one direction, by adjusting the phase constant and the periodic structure to realize the energy is being leaked and radiated from the slot structure [28]-[32]. The spoof surface plasmon polaritons (SSPPs) structures are widely used in microwave equipment, such as antennas [35]-[36], filters [37]-[38], and metasurface [39]. To improve the performance of the wide-angle beam scanning, the dual-port antenna represented by the SSPPs structure in the leaky-wave antenna shows great potential [33]. The basic operating mode of the SSPPs antenna includes the even-mode and odd-mode, while shows the broadside lobe and the end-fire lobe characteristics, respectively [34]. Generally, the one feed port SSPPs antenna mainly excites a single working mode at a given frequency. In order to achieve new operating modes to obtain multi-beam scanning characteristics, the reconfigurable network is introduced to adjust the beam direction of the SSPPs antenna. When switching the reconfigurable network state, the surface current phase of the SSPPs antenna changes, and its radiation patterns also changes accordingly [34]. However, the parasitic radiation effects, RF-link insertion loss [40], and limited switching states of the reconfigurable network all limit the performances of the reconfigurable antenna.

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For convention phased array, beam scanning characteristics mainly depend on the array layout method, array factor function, and other parameters, while the element factor in the traditional array wide-angle scanning is almost negligible.

In the previous studies [41]-[42], the generalized principle of pattern multiplication (GPPM) is proposed based on the generalized element factor (GEF). And the GEF is realized by the phase mode antenna (PMA) with dual-port, and the PMA can scan with the different phase difference between dual-port. Thus, the generalized array with PMA can scan more wide-angle range, and further realize 2-D or multi-dimension scanning characteristics with 1-D PMA array by two stage phase control.

In the previous reports [43], the dual-port antenna unit based on the SSPPs structure is excited for lobe scanning in 2-D space. By adjusting the phase difference between the dual-port feeding, the odd-mode and even-mode of the antenna are superimposed to achieve hybrid-mode. Accordingly, the continuous beam scanning from the end-fire to the broadside radiation lobe is performed. Furthermore, the two linear array based on this phase mode antenna element are constructed, and change the array factor and the element factor of 1-D linear array, in order to realize the beam scans in θ-direction and ϕ-direction are realized, respectively. The 2-D space scanning characteristics are achieved by 1-D line array, which shows good beam scanning performance. One thing to note, the profile of the antenna unit is about 0.7λ, which is too high for equipment with low profile and conformal, and the scanning range and performance of the array should still be further optimized.

In order to realize the low-profile phased antenna, expand the beam scanning range and optimize the array scanning performance in 2-D space, a dual-port phased antenna with low-profile fishbone structure and its two kinds of 1-D linear array are proposed. The designed low profile antenna unit can work in odd-mode, even-mode, and hybrid-modes by adjusting the phase difference (Δφ) between two feed ports. It can be seen, with Δφ between the dual-port of the antenna unit changing, the main lobe direction varies from the end-fire lobe to the broadside lobe (about 0° ~ 68°). In addition, two kinds of 1-D linear array are proposed based on the proposed phased antenna unit and the GPPM. Finally, considering the antenna elements layout method, the generalized element factor (GEF) function, and the generalized array factor (GAF) function are achieved to realize the 2-D space wide-angle scanning characteristics in Array#1 and Array#2.

The experimental results show that the fabricated Array#1 can realize about ±116° wide-angle beam scanning along θ-direction (xoz-plane) with gains about 6.9 ~ 11.7 dBi, and ±63° wide-angle beam scanning along ϕ-direction (yoz-plane) with gains about 11.7 ~ 6.7 dBi, and 90° to 169° wide-angle beam scanning along ϕ-direction (yoz-plane) with gains about 6.2 ~ 4.4 dBi. Meanwhile, the proposed Array#2 can obtain ±60° wide-angle beam scanning along θ-direction (xoz-plane) with gains about 7 ~ 12 dBi, and ±65° wide-angle beam scanning along ϕ-direction (yoz-plane) with gains about 11.9 ~ 7.6 dBi, and 90° to 152° wide-angle beam scanning along ϕ-direction (yoz-plane) with gains about 7.3 ~ 4 dB, respectively.

The remainder of this paper is organized as follows. Section II illustrates the proposed antenna structure, operating principle, key parameters analysis, and measured results analysis. Section III shows the proposed two kinds of 1-D linear array prototype and experimental results. Finally, the conclusion is given in Section IV.

II. ANTENNA ANALYSIS AND DESIGN

A. Antenna configuration

![Fig. 1. Configuration of the proposed antenna unit. (a) Top view. (b) Side view.](image)

The top view and side view of the proposed phased antenna unit based on fishbone structure are shown in Fig. 1(a) and Fig. 1(b), respectively. The radiation patch of the proposed phased antenna unit mainly consists of an H-shaped fishbone structure and a rhombus parasitic patch, which all are sheet copper (th1 = 2 mm). The fishbone structure looks like a very short “n” with the length of the metal arms l, the length in x-direction d, the gap s, and the profile height is h. Here, the length l of the metal arms mainly affects the operating resonant frequency, but the length d is not sensitive for the working resonant frequency and the bandwidth. We set the original value of l = 45 mm, and miniaturized d to 26 mm.

Meanwhile, the gap s between the metal arms in fishbone structure mainly affects the coupling strength, and the width p of the metal arms can appropriately improve the working bandwidth. The profile height h of fishbone structure is 10 mm, about 0.08λ, and the low profile configuration is conducive to the miniaturization of the system. The main axis width w2 in fishbone structure undertakes for the stability of the entire antenna structure, which is set 6 mm. In order to enhance the radiation performance of the antenna unit, a rhombus parasitic patch (m × n) is introduced in the center of fishbone structure. As the phased antenna, the surface current phase distribution of the antenna unit is adjusted and controlled by Δφ between the dual-port. Thus, there are two tuning stubs (w1) being added on the surface of the fishbone’s metal arms, which are used as Port 1 and Port 2 feeding, respectively. In another side, the ground (GND) (th3 = 4 mm) is mainly used for the installation of the proposed antenna and SMA connector. To avoid that fishbone structure might be shorted to the GND, one substrate (th2 = 1.524 mm, εr = 2.5, δ = 0.0013) is placed between the proposed antenna unit and the GND. Finally, the substrate and the GND are fixed by four screws.

B. Working principle of the proposed antenna unit

As shown in Fig. 1(a), the two feed ports of the proposed antenna unit can be fed with equal amplitude, and arbitrary phase difference in the range of 0° to 180°, where the phase of each feed port is defined as φ1 (Port 1) and φ2 (Port 2). Thus, the phase difference between the two feed ports is shown as Δφ = φ2 - φ1.
As mentioned above, the proposed phased antenna unit is based on fishbone structure, which can work in the odd-mode or even-mode as the leaky-wave antenna. When the phase difference $\Delta \phi$ between the two feed ports is changed, the working modes of the proposed phased antenna varies from odd-mode to hybrid-mode and then to even-mode.

1) $E_y$ component varying with the phase difference $\Delta \phi$

![Image](image1.png)

Fig. 2. Simulated $E_y$ component distribution ($\phi-z$-plane) of the proposed antenna unit, (a) front view, (b) side view, (c) only Port 1 excited, (d) $\Delta \phi = 0^\circ$, (e) $\Delta \phi = 30^\circ$, (f) $\Delta \phi = 60^\circ$, (g) $\Delta \phi = 90^\circ$, (h) $\Delta \phi = 120^\circ$, (i) $\Delta \phi = 150^\circ$, and (j) $\Delta \phi = 180^\circ$.

The simulated $E_y$ component distribution ($\phi-z$-plane) of the proposed antenna unit is shown in Fig. 2. The cross-section of the proposed antenna unit in the $\phi-z$-plane is divided into four quadrants (I, II, III, IV), the front view is shown in Fig. 2(a), and the side view is shown in Fig. 2(b).

When only one port, either Port 1 or Port 2 is excited, the $E_y$ component distribution ($\phi-z$-plane) is shown in Fig. 2(c). Meanwhile, the antenna mainly excites the odd-mode characteristic in the left part, and the $E_y$ component distribution is mainly focused on the second quadrant (II). It means that the surface currents distribution is mainly on the left-arms along the $\phi$-axis.

As shown in Fig. 2(d), when $\Delta \phi = 0^\circ$, that is the equal amplitude and the same phase, the $E_y$ component distribution is mainly focused on the first quadrant (I) and the second quadrant (II), and the color diagram are similar and symmetrical distribution. This shows that the antenna surface currents is reverse distribution along the $\phi$-axis, and the even-mode is the main working mode.

Fig. 2(e) and Fig. 2(f) show the $E_y$ component distribution ($\phi-z$-plane) when $\Delta \phi = 30^\circ$, and $\Delta \phi = 60^\circ$, respectively. It is obvious that when the phase difference $\Delta \phi$ is changed of the two feed ports, the $E_y$ component is still distributed in both the first quadrant (I) and the second quadrant (II). It is interesting that the $E_y$ component intensity gradually transfers from the first quadrant (I) to the second quadrant (II). With the phase difference $\Delta \phi$ increasing from $30^\circ$ to $60^\circ$, the colored diagram is changed to become asymmetrically distributed. In other words, although the even-mode is dominate mode, the odd-mode is also excited. In fact, this operating mode is hybrid-mode.

As the phase difference $\Delta \phi$ further changes, the phase difference $\Delta \phi$ increases from $90^\circ$-$120^\circ$ to $150^\circ$, the component distribution of these states is shown in Fig. 2(g), Fig. 2(h), and Fig. 2(i). The $E_y$ component is changed significantly, and it can be seen that the $E_y$ component intensity is not changed obviously in the first quadrant (I), while the $E_y$ component sustained growth in second quadrant (II). It shows that the odd-mode gradually becomes the dominant mode in hybrid-mode, and the influence of the even-mode becomes weaker.

The antenna unit operating in odd-mode ($\Delta \phi = 180^\circ$) is shown in Fig. 2(j), and the odd-mode is the main work mode of the proposed antenna unit. We can see that the $E_y$ component covers the first quadrant (I) and the second quadrant (II) equally, and there is strong red color (strong) electric field along $z$-axis, which means that the surface currents transfer in the same direction along the $\phi$-axis.

By analysis of the $E_y$ components ($\phi-z$-plane) variation trend, it shows that, as the phase difference $\Delta \phi$ is changed between the two feed ports, the antenna unit surface current transmission direction and the phase distribution are changed. That is to say, the proposed antenna unit can work in continuous switching operation mode of the even-mode, hybrid-mode, and odd-mode.

2) Radiation patterns varying with the phase difference $\Delta \phi$

![Image](image2.png)

Fig. 3. Simulated radiation patterns ($\phi-z$-plane) of the proposed antenna unit, (a) only Port 1 excited, (b) $\Delta \phi = 0^\circ$, (c) $\Delta \phi = 30^\circ$, (d) $\Delta \phi = 60^\circ$, (e) $\Delta \phi = 90^\circ$, (f) $\Delta \phi = 120^\circ$, (g) $\Delta \phi = 150^\circ$, and (h) $\Delta \phi = 180^\circ$.

Fig. 3 shows the radiation patterns that the proposed antenna unit works in the different modes with the different phase difference $\Delta \phi$. Only Port 1 is excited as shown in Fig. 3(a), which the radiation pattern is similar to the end-fire lobe, and the gain is about 5.3 dBi, and its operation mode is the odd-mode dominant characteristic.

In Fig. 3(b), when $\Delta \phi = 0^\circ$, the radiation pattern is the broadside radiation characteristic (even-mode). The main lobe gain is about 1.2 dBi. The radiation patterns with $\Delta \phi = 30^\circ$, $60^\circ$ are shown in Fig. 3(c) and Fig. 3(d), which shows the asymmetric distribution characteristic. The reason is that the operation mode of the proposed antenna unit is hybrid-mode, and the gains are about 2.6 dBi and 4.4 dBi, respectively. As the phase difference $\Delta \phi$ increases from $90^\circ$, $120^\circ$ to $150^\circ$, the patterns are shown in Fig. 3(e), Fig. 3(f), and Fig. 3(g). The main lobe of the radiation patterns are mainly focused on the second quadrant (II), while the main lobe continuously transits from the second quadrant (II) to the first quadrant (I). It means that the proposed antenna unit operates in hybrid-mode, and the end-fire radiation (odd-mode) gradually becomes the dominant. The simulated gains are about 6 dBi, 7.1 dBi, and 7.8 dBi, respectively. As shown in Fig. 3(h), when $\Delta \phi = 180^\circ$, the proposed antenna unit behaves as the end-fire characteristic, and its main lobe points to $\phi$-axis. This is the typical odd-mode, with the gain of approximately 8.1 dBi.

In summary, the dual-port phased antenna unit based on fishbone structure can perform continuous beam scanning in ($\phi-z$-plane) by adjusting the dual-port phase difference $\Delta \phi$. A new phased antenna unit is provided for the improvement of the wide-angle scanning performance of the phased array. The proposed phased array, not only...
improves the wide-angle scanning performance through array factor functions and optimized array element spacing, but also introduces flexible and adjustable element factor, to realize wide-angle beam scanning characteristic of phased array in multi-dimensional space.

C. Key parameters analysis

The proposed antenna unit is like short “n” based on fishbone structure, in which the branches (length \(l \times \text{width } w\)) are uniformly-spaced periodic arrangement. As we know, the surface current of each branch shows the odd-mode, even-mode, or hybrid-mode characteristics under excited of different phase differences \(\Delta \phi\). Meanwhile, considering the coupling effects of the branches, which affect the EM characteristics of the antenna unit, eventually. \(l\) and \(w\) are the key parameters of the proposed antenna unit, as shown in Fig. 4. And their effects on resonant operating frequency and coupling coefficients, etc. are further discussed.

1) Parameter \(l\)

By analysis, the resonant frequency variation trend for the \(l = 45\) mm, 46 mm, 47 mm, and 48 mm, are shown in Fig. 4(a).

Obviously, from the S-parameter curves of \(l = 45\) mm (the resonant frequency about 2.55 GHz), \(l = 46\) mm (the resonant frequency about 2.51 GHz), \(l = 47\) mm (the resonant frequency about 2.468 GHz), and \(l = 48\) mm (the resonant frequency about 2.43 GHz). It can be seen that as \(l\) increases from 45 mm to 48 mm, the resonant frequency shifts to lower frequency and vice versa. This indicates that the branch length \(l\) of fishbone structure has significant effect on the resonant frequency of the antenna unit.

Furthermore, it can be observed from Fig. 4(b) that the coupling coefficients is all less than -13 dB at center frequency 2.5 GHz, nevertheless, the coupling coefficients (|S\(_{12}\)|) migration trend to lower frequency as the \(l\) increases from 45 mm to 48 mm. This variation trend characteristic is similar to the resonant frequency change trend.

2) Parameter \(w\)

Fig. 4(c) and Fig. 4(d) show the variation trends of the reflection coefficient and coupling coefficients being related to the main axis width \(w\) in fishbone structure respectively. In Fig. 4(c), the key parameter \(w\) increases from 4 mm to 7 mm, while the resonant operating frequency shifts from 2.51 GHz to 2.572 GHz, respectively. It is obvious that the main axis of the fishbone can not only stabilize the entire antenna unit structure, but also serve as a main transmission path for surface currents. When \(w\) increases, the resonant frequency of the antenna unit is shifted to higher operating band. The Fig. 4(d) shows that the coupling coefficients (|S\(_{12}\)|) is approximately -20.2 dB (\(w = 4\) mm), -14.5 dB (\(w = 5\) mm), -11.5 dB (\(w = 6\) mm) and -10 dB (\(w = 7\) mm) at the center frequency 2.5 GHz. It is noted that the simulated coupling coefficients (|S\(_{12}\)|) deteriorates as \(w\) increases. The main reason is that the surface current transmission path is widened, which further enhances the coupling effect between the two feed ports.

### TABLE I

| Optimized Geometric Parameters for the Proposed Antenna Unit |
|---------------------------------------------------------------|
| Parameters | \(s\) | \(p\) | \(m\) | \(n\) | \(w_1\) | \(w_2\) |
| Values / mm | 1.5 | 2 | 21.3 | 16 | 4 | 4 |
| Parameters | \(l\) | \(d\) | \(h\) | \(t_1\) | \(t_2\) | \(t_3\) |
| Values / mm | 46 | 27 | 10 | 2 | 1.524 | 4 |

Fig. 4. Key parameters \(l\) and \(w\) effect on (a) and (c) \(|S_{11}|\), (b) and (d) \(|S_{12}|\) of the proposed antenna unit.

Fig. 5. The simulated and measured S-parameters, and the prototype antenna unit.
The above simulation and analysis of the key parameters reveal that the factors affecting the performance of the proposed antenna unit are manifold. Comprehensive assessments of the effect of each parameter on the antenna characteristics, have laid the groundwork for the linear array. Finally, the optimal parameters of the proposed antenna unit are shown in Table I.

D. Comparison between simulated and measured results

1) Fabricated prototype of the proposed antenna unit

Based on the analysis above, evaluation and Table I, the proposed prototype antenna unit was fabricated, and measured in microwave anechoic chamber. The prototype antenna unit is shown in Fig. 5.

2) S-parameters

The S-parameters of the proposed antenna unit are shown in Fig. 5. It can be seen that the available bandwidth with measured $|S_{11}| < -10$ dB is about 2.42 - 2.58 GHz (6.4%), meanwhile, the simulated $|S_{11}| < -10$ dB bandwidth is about 2.42 - 2.58 GHz (6.4%). Note that the measured and simulated results of $|S_{11}|$ agree well. In addition, the measured $|S_{21}| < -10$ dB bandwidth is about 2.44 - 2.59 GHz (6%), and the simulated $|S_{21}| < -10$ dB bandwidth is approximately 2.42 - 2.58 GHz (6.4%). Furthermore, the measured $|S_{12}| < -10$ dB is about 2.37 - 2.68 GHz, and at 2.5 GHz it is -13.3 dB. The simulated $|S_{12}| < -10$ dB is about 2.33 - 2.64 GHz, while at 2.5 GHz it is -16.6 dB. It can be seen that the measured S-parameters trends and available range are close to the simulated results.

It should be noted, the results for simulated $|S_{11}|$ and $|S_{21}|$ overlap, and for simulated $|S_{12}|$ and $|S_{22}|$, as well, which is due to structural symmetry and symmetric feeding. However, fabrication errors and the test environment can lead to the inconsistency between the simulated and measured results.

3) Radiation patterns

The simulated and measured radiation patterns of the proposed antenna unit are shown in Fig. 6. It can be seen, when $\Delta \varphi = 0^\circ$, the radiation patterns of the proposed antenna unit at 2.5 GHz show broadside radiation characteristic, and the simulated results agree well with the measured results. There are two radiation lobes approximately symmetrically distributed in the $\text{yoz}$-plane as shown in Fig. 6(a). Thus, the proposed antenna unit shows the typical broadside radiation characteristics from the even-mode. Meanwhile, it can be observed that the phase difference $\Delta \varphi = 0^\circ$, the main lobe direction $\theta_{\text{max}} = -68^\circ$ in the $\text{yoz}$-plane. The measured half power beam width (HPBW) is about $77.4^\circ$ and the simulated HPBW is about $123^\circ$.

In particular, there exists deviation between the measured and simulated results for the HPBW. The experimental results show that the radiation lobe of the proposed antenna unit has small valleys from about $100^\circ$ to $142^\circ$, and from $218^\circ$ to $262^\circ$. As mentioned above, the radiation characteristics are very sensitive to the machining accuracy of the antenna metal arms.

As shown in Fig. 6(b), the proposed antenna unit works in hybrid-mode ($\Delta \varphi = 30^\circ$). In other words, $\Delta \varphi = 30^\circ$, the even-mode is still the dominant operating mode, while the odd-mode is also excited. It is obvious that the main lobe direction is shifted to about $\theta_{\text{max}} = -43^\circ$, and the measured HPBW is about $84^\circ$. The reasons for the inconstancy of the HPBW results are the same as above, and will not be restated. Fig. 6(c) to Fig. 6(e) show that the proposed antenna unit can achieve wide-angle scanning characteristics, and the main beam lobe ($\theta_{\text{max}}$) scanning from $-28^\circ$, $-18^\circ$, to $-5^\circ$. The measured HPBW are about $80.8^\circ$, $70.7^\circ$, and $66^\circ$, respectively. As discussed above, with the phase difference $\Delta \varphi$ increasing from $60^\circ$, $90^\circ$, to $150^\circ$, the proposed antenna unit operates in hybrid-mode. That is to say, the broadside radiation (even-mode) gradually weakens when the phase difference $\Delta \varphi$ changes from $60^\circ$ to $150^\circ$, and the end-fire radiation (odd-mode) gradually grows. In the meantime, as the odd-mode and the even-mode continue to shift,
the main lobe direction of the proposed antenna unit is changed continuously.

As shown in Fig. 6(f), when the phase difference $\Delta \phi = 180^\circ$, the antenna works in odd-mode with end-fire, and the main lobe direction is $\theta_{oa} = 0^\circ$. The measured HPBW is about $64^\circ$ which is close to the simulated HPBW about $64.2^\circ$.

All in all, with phase difference $\Delta \phi$ increases from $0^\circ$, $30^\circ$, $60^\circ$, $90^\circ$, $150^\circ$, to $180^\circ$, operation mode switches from the even-mode, to hybrid-mode, and finally to odd-mode.

As a result, the main lobe (in $yoz$-plane) of the proposed antenna unit prototype shifts from $-68^\circ$, $-43^\circ$, $-28^\circ$, $-18^\circ$, $-5^\circ$, $0^\circ$, i.e. the proposed antenna unit exhibits continuous wide-angle beam scanning characteristic, which lays the foundation for the wide-angle beam scanning and multi-dimensional space scanning of the plane array.

4) Gains

The realized gain varied with phase difference $\Delta \phi$ is shown in Fig. 7. As the phase difference $\Delta \phi$ increases from $0^\circ$ to $180^\circ$, the measured gain increases from 0.75 dBi to 7.9 dBi with the phase difference, which is also according with the variation of the radiation mode of the proposed antenna. And the measured results are basically in agreement with the simulated except little difference.

To sum up, through analyzing on the radiation characteristics of the proposed antenna unit, it shows that as the phase difference $\Delta \phi$ changes, the surface current direction is changed, meanwhile, the odd-mode, hybrid-mode, and even-mode are continuously switched, and continuous beam lobe scanning characteristic from $-68^\circ$ to $0^\circ$ ($yoz$-plane) is further realized. The beam scanning performances of the proposed antenna unit controlled by phase difference $\Delta \phi$ are shown in Table II.

| $\Delta \phi$ | 0$^\circ$ | 30$^\circ$ | 60$^\circ$ | 90$^\circ$ | 120$^\circ$ | 150$^\circ$ | 180$^\circ$ |
|-------------|---------|---------|---------|---------|---------|---------|---------|
| $\theta_{oa}$ | $-68^\circ$ | $-43^\circ$ | $-35^\circ$ | $-28^\circ$ | $-18^\circ$ | $-11^\circ$ | $-5^\circ$ | $0^\circ$ |
| HPBW       | $77^\circ$ | $84^\circ$ | $85^\circ$ | $81^\circ$ | $71^\circ$ | $69^\circ$ | $66^\circ$ | $64^\circ$ |

III. TWO KINDS OF 1-D LINEAR ARRAY WITH SCANNING IN WIDE 2-D CHARACTERISTIC

A. Generalized Principle of Pattern Multiplication

![Diagram](image)

Fig. 8. Configuration of the $1 \times 4$ linear array based on the generalized principle of pattern multiplication.

Based on the previous discussion, to improve the wide-angle scanning performance of the directional antenna unit, to expand array wide-angle scanning and the multi-dimensional space scanning performance are new challenges. The generalized principle of the pattern multiplication (GPPM) [41]-[43] was proposed, where the phase difference $\Delta \phi$ is adjusted between dual-port of the antenna unit to change the transmission direction of the surface currents, meanwhile exciting the different operating modes, such as the odd-mode, hybrid-mode, and even-mode. In other words, the main lobe of the proposed phased antenna unit is no longer limited to the fixed directional pattern, but a flexible and changeable lobe with wide-angle scanning characteristics. In further, the array based on the proposed phased antenna unit can realize beam scanning with the flexible element factor and array factor, or the phase difference $\Delta \phi$ between the dual-port of the antenna unit and the phase difference $\tau$ between the adjacent elements.

The array configuration based on the principle is shown in Fig. 8. As an example, four identical dual-port phased antenna units are denoted as Ant. 1, Ant. 2, Ant. 3, and Ant. 4, with the feed ports numbered as Port 1, Port 2, Port 3, Port 4, Port 5, Port 6, Port 7, and Port 8, respectively.

Therefore, the antenna units have the same initial phase set, which is $(\phi_1, \phi_2)$, where the phase difference between Port 1 and Port 2 is $\Delta \phi = 2\phi_2 - \phi_1$. While the phase difference between every two adjacent antenna units is $\tau$.

It can be seen that the generalized array factor (GAF) of the dual-port antenna unit can also be expressed as

$$f_\alpha(\theta, \phi, \theta_{oa}, \phi_{oa}) = f_\alpha(\theta, \phi, \theta_{oa}(\tau), \phi_{oa}(\tau))$$

$\tau$ is the phase difference between adjacent antenna units. Meanwhile, the generalized element factor (GEF) is denoted as

$$f_e(\theta, \phi, \theta_{oa}, \phi_{oa}) = f_e(\theta, \phi, \theta_{oa}(\Delta \phi), \phi_{oa}(\Delta \phi))$$

Thus, the generalized principle of pattern multiplication (GPPM) can be written as

$$f(\theta, \phi, \theta_{oa}, \phi_{oa}) = f(\theta, \phi, \theta_{oa}(\tau), \phi_{oa}(\tau) \times f(\theta, \phi, \theta_{oa}(\Delta \phi), \phi_{oa}(\Delta \phi)))$$

It is evident that the equation (3) has introduced total of four variables, $\theta_{oa}(\tau)$, $\phi_{oa}(\tau)$, $\theta_{oa}(\Delta \phi)$, and $\phi_{oa}(\Delta \phi)$, which vary along the $\theta$-direction and $\phi$-direction. It means that the degree of freedom of the array in wide-angle beam scanning applications is increased, and which is advantageous for improving the performance of the array in wide-angle scanning as well as in 2-D space scanning. It can be seen from this discussion, the proposed dual-port phased antenna unit based on fishbone structure is just a good GEF element, which may achieve wide-angle beam scanning characteristics in 2-D space for the $1 \times 4$ 1-D linear array, more specifically, in the $xoz$-plane ($\theta_{oa}$), $yoz$-plane ($\phi_{oa}$), and $xoy$-plane ($\phi_{oa}$) by adjusting the generalized element factor (GEF) function, the generalized array factor (GAF) function, and the array layout. The phase states of all ports inside the antenna array are shown in Table III.

| Number | Port 1 | Port 2 | Port 3 | Port 4 |
|--------|--------|--------|--------|--------|
| Port Phase (/ degree) | $\phi_1$ | $\phi_1, \Delta \phi$ | $\phi_1, \tau$ | $\phi_1, \tau, \Delta \phi$ |
| Number | Port 5 | Port 6 | Port 7 | Port 8 |
| Port Phase (/ degree) | $\phi_1, 2\tau$ | $\phi_1, \Delta \phi, 2\tau$ | $\phi_1, 3\tau$ | $\phi_1, 3\tau, \Delta \phi$ |

In the real, the proposed antenna unit can be extended to 1-D linear array in three arrangements, vertical arrangement (Array#1), oblique
45° arrangement (Array#2), and the horizontal arrangement (Array#3).

For Array#3, the y-side length is larger than the x-side length of the antenna unit, so that the interval between the adjacent elements is larger and limits the scanning ability of Array#3. Thus, we only discuss Array#1 and Array#2. To enlarge the scanning angle and suppress grating lobe, the element spacing is usually set to be less than half wavelength [44], or equal to half wavelength.

B. Array#1 ($1 \times 4$ 1-D Linear Array)

The configuration of the proposed Array#1 and the fabricated prototype of the proposed Array#1 are shown in Fig. 9. The proposed Array#1 is linearly arranged along the x-axis, with the array spacing 0.4λ. The phase-shifting network in measurement was composed of a one-to-eight equal-split power divider, eight 8-bit phase shifters, and some phase compensating cable.

1) $S$-parameters of the Array#1

The measured and simulated S-parameters of the proposed Array#1 are shown in Fig. 10(a). The measured $|S_{11}| < -10$ dB band is about 2.43 ~ 2.56 GHz, and the simulated is 2.38 ~ 2.54 GHz, the center of the measured results moved a little to the lower band. Similarly, the measured available bands of $|S_{22}|$, $|S_{33}|$, $|S_{44}|$, $|S_{55}|$, $|S_{66}|$, $|S_{77}|$, $|S_{88}|$, and $|S_{99}|$ are 2.4 ~ 2.56 GHz, 2.43 ~ 2.57 GHz, 2.34 ~ 2.56 GHz, 2.43 ~ 2.57 GHz, 2.43 ~ 2.56 GHz, 2.39 ~ 2.55 GHz, 2.42 ~ 2.5 GHz, respectively. Compared to the simulated reflection coefficient curves of these ports, the measured reflection coefficient curves moved to the lower band a little. Significantly, the simulated S-parameters curves overlap visually.

The coupling coefficients for the proposed Array#1 are shown in Fig. 10(b). It can be seen that the measured $|S_{11}|$, $|S_{22}|$, $|S_{33}|$, $|S_{44}|$, $|S_{55}|$, $|S_{66}|$, $|S_{77}|$, $|S_{88}|$, and $|S_{99}|$ are in good agreement, and the overall bands are from about 2.4 to 2.65 GHz with the coupling coefficients being less than -10 dB. Meanwhile, the measured curves are close to the simulated, besides 0.03 GHz offset to low frequency. In fact, the coupling coefficients are less than -15 dB at the center frequency of 2.5 GHz, whether in simulation or measurement.

2) Measured and simulated beam scanning results in 2-D space for Array#1

From the equation (3), the proposed dual-port phased antenna unit and the corresponding 1-D linear array combined with the GPPM can realize 2-D space beam scanning ($\phi_0$, $\theta_0$) in $xoz$-plane, and $yoz$-plane (by adjusting the GEF, the GAF, and their matching relationship).

(i) Beam scanning in $xoz$-plane

For Array#1, the main beam scanning to 2-D space can be realized with varying phase difference ($\Delta \phi$, $\Delta \tau$). The direction of the main beam with different phase differences ($\Delta \phi$, $\Delta \tau$) for Array#1 are shown in Fig. 11 ($xoz$-plane).

As shown in Fig. 11(a) in the $xoz$-plane, when the phase difference ($\Delta \phi$, $\Delta \tau$) = (180°, 0°), the main beam direction is $\theta_0 = 0°$. The measured and simulated results match relatively well, meanwhile, the measured first side lobe level is about -11 dB and the measured HPBW are approximately 29.4°, which are close to the simulated results. In Fig. 11(b), the main beam direction is $\theta_0 = -30°$, when the phase difference ($\Delta \phi$, $\Delta \tau$) = (90°, 80°). In Fig. 11(c), the main beam direction is $\theta_0 = -53°$, when the phase difference ($\Delta \phi$, $\Delta \tau$) = (120°, 170°). In Fig. 11(d), the main beam direction is $\theta_0 = -93°$ with the phase difference ($\Delta \phi$, $\Delta \tau$) = (100°, 170°), and $\theta_0 = -116°$ with ($\Delta \phi$, $\Delta \tau$) = (150°, 140°) in Fig.
Fig. 11. The radiation patterns (xoz-plane) of the proposed Array#1 with phase difference \( (\Delta \phi, \Delta \tau) \) change, (a) \( \theta_m = 0^\circ \), \( (\Delta \phi, \Delta \tau) = (180^\circ, 0^\circ) \), (b) \( \theta_m = -30^\circ \), \( (\Delta \phi, \Delta \tau) = (90^\circ, 80^\circ) \), (c) \( \theta_m = -53^\circ \), \( (\Delta \phi, \Delta \tau) = (120^\circ, 170^\circ) \), (d) \( \theta_m = -93^\circ \), \( (\Delta \phi, \Delta \tau) = (100^\circ, 170^\circ) \), (e) \( \theta_m = -116^\circ \), \( (\Delta \phi, \Delta \tau) = (150^\circ, 140^\circ) \), and (f) side view of the Array#1.

Fig. 12. The radiation patterns (yoz-plane) of the proposed Array#1 with phase difference \( (\Delta \phi, \Delta \tau) \) change, (a) \( \theta_m = 0^\circ \), \( (\Delta \phi, \Delta \tau) = (180^\circ, 0^\circ) \), (b) \( \theta_m = 22^\circ \), \( (\Delta \phi, \Delta \tau) = (90^\circ, 0^\circ) \), (c) \( \theta_m = 41^\circ \), \( (\Delta \phi, \Delta \tau) = (40^\circ, 0^\circ) \), (d) \( \theta_m = 51^\circ \), \( (\Delta \phi, \Delta \tau) = (20^\circ, 0^\circ) \), (e) \( \theta_m = 63^\circ \), \( (\Delta \phi, \Delta \tau) = (0^\circ, 0^\circ) \), and (f) side view of the Array#1.

Fig. 13. The radiation patterns (xoy-plane) of the proposed Array#1 with phase difference \( (\Delta \phi, \Delta \tau) \) change, (a) \( \Phi_m = 90^\circ \), \( (\Delta \phi, \Delta \tau) = (0^\circ, 0^\circ) \), (b) \( \Phi_m = 127^\circ \), \( (\Delta \phi, \Delta \tau) = (0^\circ, 90^\circ) \), (c) \( \Phi_m = 143^\circ \), \( (\Delta \phi, \Delta \tau) = (0^\circ, 120^\circ) \), (d) \( \Phi_m = 162^\circ \), \( (\Delta \phi, \Delta \tau) = (0^\circ, 160^\circ) \), (e) \( \Phi_m = 169^\circ \), \( (\Delta \phi, \Delta \tau) = (0^\circ, 180^\circ) \), and (f) side view of the Array#1.

11(c). Apparently, the main beam scanning in the xoz-plane is realized by Array#1 with varying \( (\Delta \phi, \Delta \tau) \). With the \( \theta_m \) increasing, the HPBW is widened too. However, there are little errors in the simulated and measured HPBW, which are mainly due to errors in the accuracy of the manufacturing process.

(ii) Beam scanning in yoz-plane

The measured wide-angle beam scanning performances of the proposed Array#1 in the yoz-plane are compared with simulated results, as shown in Fig. 12. It is noteworthy that the beam scanning performance of the Array#1 in the yoz-plane mainly relies on the even-mode. In other words, Array#1 can achieve wide-angle beam scanning characteristic in yoz-plane by adjusting the even-mode, hybrid-mode, and odd-mode of the antenna unit. The operating principle and radiation characteristics of the even-mode have been demonstrated in section II, thus, to avoid repetition, it’s not described in this part.

As shown in Fig. 12(a), when the phase difference \( (\Delta \phi, \Delta \tau) = (180^\circ, 0^\circ) \), the main beam direction is \( 0^\circ \) (yoz-plane), which is the end-fire radiation. The measured HPBW is about 61\(^\circ\), and the measured first side lobe level is about -13.3 dB.

In Fig. 12(b), when the phase difference \( (\Delta \phi, \Delta \tau) = (90^\circ, 0^\circ) \), the main beam direction is 22\(^\circ\) (yoz-plane), the HPBW is about 65.3\(^\circ\). When the phase difference \( (\Delta \phi, \Delta \tau) = (40^\circ, 0^\circ) \), the main beam direction is 41\(^\circ\) (yoz-plane) in Fig. 12(c), it can be seen that the Array#1 exhibits more obvious even-mode characteristics, i.e. the two beams are progressively similar in shape, while the HPBW covers larger area, about 133\(^\circ\). As can be seen from the Fig. 12(d) and Fig. 12(e), when the phase difference \( (\Delta \phi, \Delta \tau) = (20^\circ, 0^\circ) \), and \( (\Delta \phi, \Delta \tau) = (0^\circ, 0^\circ) \), the main beam directions are at 51\(^\circ\) and 63\(^\circ\), respectively. Meanwhile, the measured HPWBWs are about 90\(^\circ\) and 93\(^\circ\), respectively.

The measured and simulated radiation patterns basically match. Fig. 12(e) shows almost the same symmetrical beam (broadside radiation), it suggests the radiation characteristics of the two radiation beams are approximately the same shape. Especially the even-mode.

(iii) Beam scanning in xoy-plane
Next, the measured main beam scanning in xoy-plane are given in Fig. 13. In Fig. 13(a), the phase difference \((\Delta \phi, \Delta \tau) = (0^\circ, 0^\circ)\), the Array#1 works in the even-mode, meanwhile the radiation patterns is broadside, i.e., there are a pair of symmetrically distributed beams, and the main beam direction is \(\Phi_m = 90^\circ\). In Fig. 13(b), the main beam direction is about \(\Phi_m = 127^\circ\) with the phase difference \((\Delta \phi, \Delta \tau) = (0^\circ, 90^\circ)\). And the main beam point to \(\Phi_m = 143^\circ\) with the phase difference \((\Delta \phi, \Delta \tau) = (0^\circ, 120^\circ)\) in Fig. 13(c). It can be seen that with the constant change of the phase difference \((\Delta \phi, \Delta \tau)\), the main beam direction of the proposed Array#1 gradually moves from broadside to end-fire in hybrid operation mode in xoy-plane. In Fig. 13(d), the main beam direction is about \(\Phi_m = 162^\circ\) with the phase difference \((\Delta \phi, \Delta \tau) = (0^\circ, 160^\circ)\), and in Fig. 13(e), the main beam points to \(\Phi_m = 169^\circ\) with the phase difference \((\Delta \phi, \Delta \tau) = (0^\circ, 180^\circ)\). It was clear that, with phase difference \((\Delta \phi, \Delta \tau)\) changing, the main beam performs scanning characteristic in xoy-plane. And the proposed Array#1 radiates from even-mode to hybrid-mode, and then odd-mode, or broadside radiation to mixed radiation and to end-fire pattern finally.

In conclusion, with phase difference \((\Delta \phi, \Delta \tau)\) changing, the Array#1 can perform wide-angle scanning in 2-D space, \(0^\circ \sim -116^\circ\) in the xoz-plane, \(0^\circ \sim -63^\circ\) in the yoz-plane, and \(90^\circ \sim 169^\circ\) in the xoy-plane.

C. Array#2 (1 × 4 1-D Linear 45\(^{\circ}\) Array)

In order to reduce the coupling effects between antenna units and improve beam scanning performance of array in 2-D space, the proposed antenna unit is rotated \(45^\circ\) around the z-axis, and four antenna units are extended along x-axis to construct the Array#2, as shown in Fig. 14.

1) S-parameters of the Array#2

The measured and simulated S-parameters of the proposed Array#2 are given in Fig. 15(a). The measured bandwidth is about 2.35 \(\sim 2.54\) GHz to \(\text{S}_{11} < -10\text{ dB}, 2.46 \sim 2.57\text{ GHz} \text{ to } \text{S}_{21} < -10\text{ dB}, 2.44 \sim 2.57\text{ GHz} \text{ to } \text{S}_{31} < -10\text{ dB}, 2.34 \sim 2.55\text{ GHz} \text{ to } \text{S}_{41} < -10\text{ dB}, 2.2 \sim 2.57\text{ GHz} \text{ to } \text{S}_{51} < -10\text{ dB}, 2.44 \sim 2.58\text{ GHz} \text{ to } \text{S}_{61} < -10\text{ dB}, 2.35 \sim 2.54\text{ GHz} \text{ to } \text{S}_{71} < -10\text{ dB}, \text{ and } 2.44 \sim 2.57\text{ GHz} \text{ to } \text{S}_{81} < -10\text{ dB.}

Comparing to the simulated curves, all the measured curves shift a little to lower band, but the S-parameters at the center frequency 2.5 GHz are all less than -14 dB.

The coupling coefficients between the adjacent ports for the proposed Array#2 are shown in Fig. 15(b). It can be seen that the measured available band with \(\text{S}_{12}, \text{S}_{21}, \text{S}_{23}, \text{S}_{32}, \text{S}_{34}, \text{S}_{43}, \text{S}_{41}, \text{S}_{52}, \text{S}_{53}, \text{S}_{54}, \text{S}_{61}, \text{S}_{62}, \text{S}_{63}, \text{S}_{64}, \text{S}_{73}, \text{S}_{74}, \text{S}_{84}\), and \(\text{S}_{78}\) less than -10 dB ranges from about 2.35 to 2.6 GHz. And the measured and simulated coupling coefficients are less than -11 dB at the center frequency 2.5 GHz.

2) Measured and Simulated beam scanning results in 2-D space for Array#2

From the equation (3), the proposed dual-port phased antenna unit and the corresponding 1-D linear Array#2 combined with the GPPM can realize 2-D space beam scanning (xoz-plane \((\theta_m)\), yoz-plane \((\phi_m)\), and xoy-plane \((\Phi_m)\)) by adjusting the GEF, the GAF, and their matching relationship.

(i) Beam scanning in xoz-plane

For Array#2, the main beam scanning in xoz-plane can be realized with varying phase difference \((\Delta \phi, \Delta \tau)\).

The beam scanning in xoz-plane with different phase differences \((\Delta \phi, \Delta \tau)\) for Array#2 are shown in Fig. 16. The phase difference \((\Delta \phi, \Delta \tau) = (180^\circ, 0^\circ)\), the proposed Array#2 operates in the odd-mode, and the end-fire beam is achieved. The measured HPBW is about 22\(^{\circ}\).

In Fig. 16(b), the measured main beam direction is \(\theta_m = -30^\circ\), and the HPBW is about 29\(^{\circ}\) with phase difference \((\Delta \phi, \Delta \tau) = (120^\circ, 100^\circ)\). Here, the proposed Array#2 operates in hybrid-mode, with the odd-mode still working as the dominant operation mode. Significantly, Fig. 16(b) shows that when \(\theta_m\) is around \(-148^\circ\) there is deviation between measured and simulated results, which is mainly due to machining errors and the uncertain effects of the measurement environment.

From what Fig. 16(c) shows, when the phase difference \((\Delta \phi, \Delta \tau) = (100^\circ, 125^\circ)\), the measured main beam direction is \(\theta_m = -40^\circ\) in xoz-plane, and the measured first side lobe level is about -11 dB, the HPBW is about 31.1\(^{\circ}\). There is also deviation of the side lobe direction around \(-140^\circ\) between the measured and simulated results, due to the same reasons for Fig. 16(b). Fig. 16(d) shows that the measured main
beam direction $\theta_m = -50^\circ$, the phase difference $(\Delta\varphi, \Delta\tau) = (60^\circ, 145^\circ)$, meanwhile, the measured first side lobe level is about $-11.8$ dB, and the measured first side lobe level is about $-11.3$ dB, and the HPBW is about $30.4^\circ$. The measured and simulated curves are in good agreement. In Fig. 16(e), the measured main beam direction is $-60^\circ$ for Array#2, while the phase difference $(\Delta\varphi, \Delta\tau) = (60^\circ, 180^\circ)$, furthermore, the measured HPBW is about $38.2^\circ$, and the measured first side lobe level is about $-5.2$ dB.

(ii) Beam scanning in yoz-plane

For Array#2, the main beam scanning in yoz-plane can be realized with varying phase difference $(\Delta\varphi, \Delta\tau)$. The main beam scanning characteristics of the proposed Array#2 in the yoz-plane are discussed below. The main beam scanning characteristics in the yoz-plane can be achieved, by adjusting the generalized element factor (GEF). Fig. 17(a) shows that the measured main beam direction is $\theta_m = 0^\circ$ with the phase difference $(\Delta\varphi, \Delta\tau) = (180^\circ, 0^\circ)$. In addition, the measured HPBW is about $66^\circ$. It can be observed that the proposed Array#2 shows the end-fire (odd-mode), with a good symmetry characteristic of its radiation patterns. From the Fig. 17(b) to Fig. 17(d), the proposed Array#2 operates in hybrid-mode, which results from the linear superposition of the odd-mode and even-mode, meanwhile, with phase differences $(\Delta\varphi, \Delta\tau) = (80^\circ, 0^\circ)$, $(\Delta\varphi, \Delta\tau) = (50^\circ, 0^\circ)$, and $(\Delta\varphi, \Delta\tau) = (30^\circ, 0^\circ)$. By observing its radiation patterns, it can be seen that the measured main beam directions of the proposed Array#2 are $\theta_m = 29^\circ$, $\theta_m = 41^\circ$, and $\theta_m = 50^\circ$, respectively. Meanwhile, the HPBW are about $89.3^\circ$, $84^\circ$, and $80^\circ$, respectively.

Finally, as seen in Fig. 17(e), the proposed Array#2 operates in the even-mode with the main beam direction $\theta_m = 65^\circ$, and $(\Delta\varphi, \Delta\tau) = (0^\circ, 0^\circ)$. Meanwhile, the HPBW is about $74^\circ$. In fact, as seen in Fig. 17(d) and Fig. 17(e), when the main beam direction $\theta_m = 50^\circ$ and $\theta_m = 65^\circ$, there are some errors between the simulated and measured HPBW. That is to say, the machining accuracy has certain effect on the measured results.

(iii) Beam scanning in xoy-plane

For Array#2, the main beam scanning in xoy-plane can be realized with varying phase difference $(\Delta\varphi, \Delta\tau)$. Next, the beam scanning characteristics of Array#2 in the xoy-plane are shown in Fig. 18. In Fig. 18(a), the Array#2 operates in even-mode when $(\Delta\varphi, \Delta\tau) = (0^\circ, 0^\circ)$ with main beam direction points to $\Phi_m = 90^\circ$ in xoy-plane, meanwhile, which shows broadside radiation patterns. From the linear superposition of the odd-mode and even-mode, meanwhile, with phase differences $(\Delta\varphi, \Delta\tau) = (80^\circ, 0^\circ)$, $(\Delta\varphi, \Delta\tau) = (50^\circ, 0^\circ)$, and $(\Delta\varphi, \Delta\tau) = (30^\circ, 0^\circ)$. By observing its radiation patterns, it can be seen that the measured main beam directions of the proposed Array#2 are $\theta_m = 29^\circ$, $\theta_m = 41^\circ$, and $\theta_m = 50^\circ$, respectively. Meanwhile, the HPBW are about $89.3^\circ$, $84^\circ$, and $80^\circ$, respectively.

Finally, as seen in Fig. 17(e), the proposed Array#2 operates in the even-mode with the main beam direction $\theta_m = 65^\circ$, and $(\Delta\varphi, \Delta\tau) = (0^\circ, 0^\circ)$. Meanwhile, the HPBW is about $74^\circ$. In fact, as seen in Fig. 17(d) and Fig. 17(e), when the main beam direction $\theta_m = 50^\circ$ and $\theta_m = 65^\circ$, there are some errors between the simulated and measured HPBW. That is to say, the machining accuracy has certain effect on the measured results.

(iii) Beam scanning in xoy-plane

For Array#2, the main beam scanning in xoy-plane can be realized with varying phase difference $(\Delta\varphi, \Delta\tau)$. Next, the beam scanning characteristics of Array#2 in the xoy-plane are shown in Fig. 18. In Fig. 18(a), the Array#2 operates in even-mode when $(\Delta\varphi, \Delta\tau) = (0^\circ, 0^\circ)$ with main beam direction points to $\Phi_m = 90^\circ$ in xoy-plane, meanwhile, which shows broadside radiation patterns. From the linear superposition of the odd-mode and even-mode, meanwhile, with phase differences $(\Delta\varphi, \Delta\tau) = (80^\circ, 0^\circ)$, $(\Delta\varphi, \Delta\tau) = (50^\circ, 0^\circ)$, and $(\Delta\varphi, \Delta\tau) = (30^\circ, 0^\circ)$. By observing its radiation patterns, it can be seen that the measured main beam directions of the proposed Array#2 are $\theta_m = 29^\circ$, $\theta_m = 41^\circ$, and $\theta_m = 50^\circ$, respectively. Meanwhile, the HPBW are about $89.3^\circ$, $84^\circ$, and $80^\circ$, respectively.
characteristic. It is clearly seen from Fig. 18(b) to Fig. 18(d), under the hybrid-mode, the proposed Array#2 exhibits double-lobe characteristics in the xoy-plane. When the phase differences $(\Delta \phi, \Delta \tau)$ are $(0^\circ, 60^\circ)$, $(0^\circ, 90^\circ)$, and $(0^\circ, 120^\circ)$, respectively, the main beam directions are $\Phi_m = 110^\circ$, $120^\circ$, and $129^\circ$, respectively. Meanwhile, the HPBW is about $30^\circ$, $31^\circ$, and $31^\circ$, respectively. Furthermore, the first side lobe level is about $-9.7$ dB, $-9.7$ dB, and $-9.3$ dB, respectively. In Fig. 18(e), $(\Delta \phi, \Delta \tau) = (0^\circ, 170^\circ)$ with main beam direction points to $\Phi_m = 152^\circ$, it is close to end-fire radiation.

To sum up, the proposed antenna array has a positive effect for improvement the 2-D space beam scanning performance (in xoz-plane, yoz-plane, and xoy-plane, respectively), when rotated $45^\circ$ along the $z$-axis.

The proposed two kinds of the 1-D antenna array to 2-D beam scanning are compared with the reported wide-angle scanning array in Table IV. It is noted that the presented 1-D antenna array in this paper can realize 2-D beam scanning with profile 0.08λ, which is much smaller than the reported array in [43]. So, our array is very suitable for conformal installation of high-speed platform.

### IV. Conclusion

This paper presents a novel low-profile dual-port phased antenna unit based on fishbone structure. Analysis shows that by adjusting the phase difference $\Delta \phi$ between the dual-port of the antenna, the proposed antenna unit can switch among the odd-mode and even-mode, and hybrid operation mode. And the hybrid-mode is the linear superposition of the odd-mode and even-mode. Finally, the proposed antenna unit achieves continuous beam scanning characteristics from broadside radiation lobe to end-fire radiation lobe.

Then, two kinds of the $1 \times 4$ 1-D linear array (Array#1 and Array#2) are constructed by the 1-D extension of the proposed phased antenna unit. And Array#1 and Array#2 prototypes are fabricated based on the proposed antenna unit applying the generalized principle of the pattern multiplication (GPPM). Further, by analyzing the generalized element factor (GEF), the generalized array factor (GAF), and the arrangement method of the array, it is found that both 1-D array (Array#1 and Array#2) can realize wide-angle beam scanning in 2-D space. More specifically, Array#1 achieves $-116^\circ$ $-0^\circ$ wide-angle scanning along $\theta$-direction (xoz-plane) with gains about $6.7$ $11.7$ dBi and HPBW about $29.4^\circ$ $105^\circ$, $0^\circ$ $63^\circ$ wide-angle scanning along $\phi$-direction (yoz-plane) with gains about $6.7$ $11.7$ dBi and HPBW about $59^\circ$ $133^\circ$, and $90^\circ$ $169^\circ$ wide-angle scanning along $\phi$-direction (yoz-plane) with gains about $4.4$ $6.2$ dBi and HPBW about $32^\circ$ $123^\circ$, respectively. Meanwhile, the proposed Array#2 obtains $-60^\circ$ $0^\circ$ beam scanning along $\theta$-direction (xoz-plane) with gains about $7$ $12$ dBi and HPBW about $22^\circ$ $32^\circ$, $0^\circ$ $65^\circ$ wide-angle scanning along $\theta$-direction (yoz-plane) with gains about $7.6$ $11.9$ dBi and HPBW about $66^\circ$ $90^\circ$, and $90^\circ$ $152^\circ$ wide-angle scanning along $\phi$-direction (xoy-plane) with gains about $4$ $7.3$ dBi and HPBW about $27^\circ$ $267^\circ$, respectively.

The proposed phased antenna unit and its 1-D linear array (Array#1 and Array#2) based on fishbone structure combined with the GPPM show influential significance for improving the wide-angle beam scanning performance and scanning dimension space of the 1-D linear array. Meanwhile, it has positive effect on improving future wireless communication systems and radar systems.

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