Ka-band microstrip array antenna with compact series-parallel hybrid strip-line feeding network

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Abstract This study proposes a Ka-band high-performance microstrip array antenna having a planar array front-end and strip-line back-end. 10 × 4 hybrid-fed rectangular patches with Taylor-synthesized excitation is adopted to form weighted patterns. Preferable symmetrical patterns are obtained covering 34.5–35.5 GHz working band due to the symmetric feeding arrangement. The measured reflection coefficient is lower than −10 dB, and peak co-polarization gain and first side-lobe level are better than 19 dB and −17 dB, respectively. Additionally, the strip-line back-end ensures an ideal front-to-back ratio of approximately 40 dB. The measurement results coincide with the simulation results, proving the immense prospect of our design in airspace detection or other applicable scenarios.

key words: Airspace detection, narrow beam, compact structure, Taylor synthesis, planar microstrip array, strip-line feeding network

Classification: Microwave and millimeter wave devices, circuits, and hardware

1. Introduction

Array antennas have been extensively used in millimeter-wave (MMW) technologies because of their capability to radiate highly directional patterns owing to specific array arrangements [1, 2, 3]. Generally, to reduce clutter interference, low first side-lobe level (FSLL) and low front-to-back ratio (FBR) are expected in antenna design [4]. In addition, their low profile characteristics and light weight meet application requirements [5, 6]. Motivated by these situations and the developing technology of printed circuit boards (PCBs), many types of substrate integrated waveguide (SIW) slot arrays [7, 8] and series [9, 10, 11, 12, 13,14, 15] or parallel [16, 17, 18, 19, 20, 21] fed microstrip arrays are gradually replacing the traditional metallic waveguide or plate slot array antenna, as in [22, 23]. Such PCB processing-based antennas have become the primary option in modern radar systems. The commonly used methods for achieving low FSLL performance include adjusting each element’s pattern, as in [24], or adjusting the array’s excitation distribution [25, 26, 27, 28, 29]. In [25] and [27], large-scale co-planar parallel-fed networks were adopted to realize required current coefficients, and the measured FSLL reached approximately −15 dB and −20 dB, respectively. Such a structure may degrade the cross-polarization level, and the air gap used in [25] results in a thick antenna profile. References [13, 24] employed a low-profile single-layer microstrip layout, but their edge feed method caused asymmetry of the antenna patterns to some extent. A dual-layer substrate and bottom-printed microstrip feeding were used in [26] to improve the asymmetry issue in [24], however, this structure had strong backward radiation. SIW and metallic waveguide slot arrays were used as the back-end components in [28] and [29], respectively, which ensured low FSLL and low FBR of the antenna while preserving the ideal symmetrical pattern. However, both designs require connecting inconvenient waveguide-coaxial transformers in their actual applications. In this study, we propose a Ka-band center-fed planar microstrip array antenna. This design comprises front-end and back-end segments. The front-end consists of four series-fed linear microstrip subarrays. Each subarray includes ten rectangular patch elements with a tapered width distribution and a center-placed U-shaped phase inverter. The back-end adopts a compact comb strip-line feeding network [30] with a symmetrical layout in the xoz plane. Four via holes of 0.4 mm diameter are used to connect the inverter and strip-line conductors. In addition, because the via holes located on the left and right sides of the x-axis connect the upper and lower ends of the counterpart inverters, respectively, this antenna has the preferable symmetric patterns and “tongue-shaped” 3-D patterns across the 34.5–35.5 GHz working band. To prove the validity of our design, the proposed model was fabricated using multilayer PCB technology. The measurement results show reasonable agreement with the simulated results obtained using ANSYS HFSS 18.0 software.

2. Theoretical design

2.1 Overall scheme and flow chart

As mentioned above, the 2-D planar microstrip array antenna containing a multilayer substrate was designed...
as separate front-end and back-end segments. The specific design steps are presented in Fig. 1. The first step involved synthesizing the patterns based on the performance requirements of the concerned application. The expected half-power beamwidth (HPBW) of 10° in the E plane and 20° in the H plane were set as the initial goal values. The E-plane patterns were obtained using the Taylor-weighted patch width distribution of the 1-D linear subarray. Four subarrays were excited by a series-parallel hybrid network with adjusted impedance transform sections in the back end to form Taylor-weighted H-plane patterns. All components are introduced in the following subsections in the order of the design steps shown in Fig. 1.

2.2 Taylor pattern synthesis in E/H plane
As shown in Fig. 2, the designed planar array is composed of non-identical elements discretely arrayed in the xoz plane with the coordinates of origin located at the plate center position. Each subarray consists of M rectangular patches with different widths arranged uniformly along the x-axis. Then, N identical subarrays expand uniformly along the z-axis to form a 2-D configuration. According to the mechanism introduced in [1], the element coordinate at \((x_m, z_n)\) radiates according to the equivalent magnetic currents at its two radiation edges. Under far-zone, time harmonic, and infinite ground conditions, the complex vector of the E-field generated by each element can be calculated by integrating its surface magnetic current sources:

\[
\vec{E}_{\text{mm}} = \vec{e}_p \cdot j \frac{2U_n \cdot e^{-jk \cdot r}}{\pi \cdot r} \sin\left(\frac{1}{2}kW_m \cos \theta \right) \frac{\cos \theta \cdot \sin \theta}{\cos \theta} .
\]

It should be noted that the edge voltage phasors \(U_n\) is treated as constant in each subarray under the standing wave series-fed structure approximation. In the ideal case, the E-field has only the \(\vec{e}_p\) direction component for the z-direction equivalent magnetic currents. In an actual model, the \(\vec{e}_e\) and \(\vec{e}_o\) direction components act as the co-polarization and cross-polarization components, respectively. Furthermore, by superposing the radiation of each element, as shown in Eq. (1), the E-field intensity pattern in the E plane (xoy) and H plane (yoz) can be obtained as follows:

\[
f_E(\phi) \approx \left| \cos \left(\frac{1}{2}kL \cdot \cos \phi \right) \sum_{m=1}^{M} W_m \cdot e^{jk(x_m \cos \phi + z_n \sin \phi)} \right| , \tag{2}
\]

\[
f_H(\theta) = \left| \sum_{n=1}^{N} W_n \cdot \left( \frac{1}{2}kW_m \cdot \cos \theta \right) \sin \theta \sum_{m=1}^{M} U_n \cdot e^{jk(x_m \cos \phi + z_n \sin \phi)} \right| . \tag{3}
\]

In our design, the element number M and subarray number N are set to ten and four, respectively, with all elements located symmetrically in the xoz plane. Hence, the element positions can be determined by the arithmetic sequence characters. Based on Eq. (2) and Eq. (3), the maximum radiation occurs in the broadside (+y) direction. The Taylor synthesis method is used to optimize the patch widths \(W_m\) and edge voltage phasors \(U_n\) under a given array arrangement for suppressing the FSLL at the center frequency of 35 GHz. The theoretically synthesized patterns are plotted in Fig. 3, where the targeted values of the FSLL are set to \(-21\) dB and \(-18\) dB in the E and H planes, respectively.

### 3. Simulation procedure

3.1 Linear sub-array consisting of non-identical elements
Fig. 3 shows the Taylor synthesized patterns with suppressed FSLL. The HPBWs in the two principal planes were designed to be approximately 10° and 20°. Hence, the entire 3-D pattern presents a tongue-like shape. Next, we describe the design of the actual antenna structure according to the synthesis results presented in Subsection 2.2. The first task is creating the 1-D subarray with the desired E-plane pattern. The patch width distribution \(W_m\) obtained by the theoretical method was used as the initial parameter to construct the simulation model shown in Fig. 4. Unlike the theoretical analysis model in Fig. 2, a U-shaped phase inverter, which acts as the equivalent of the slot coupler used in [26, 28, 29], is loaded at the center position of each subarray. With this design, the elements on both sides of the subarray can be
fed in the reverse phase, thus forming a sum pattern in the E plane. Specifically, introducing such a phase inverter may cause degradation of the E-plane pattern symmetry to some extent. However, this drawback can be addressed by placing the feeding points at different ends of the U-shaped section of each subarray, as shown in Fig. 6(a).

The H-plane pattern should be achieved by weighted excitation in the yoz plane. The 1-D subarray mentioned in Subsection 3.1 is used as a generalized element, and four such elements are arranged along the z-axis uniformly to form a 2-D planar array front end. The corresponding back-end feeding network is illustrated in Fig. 5. To restrain the backward radiation, our design adopts a strip-line feeding layout. The upper feeding layer connects the front end via four metalized blind holes at the U-shaped section ends of each subarray. The lower layer can be transferred into a 50-ohm coaxial converter and cable by using a metal probe of 0.3 mm diameter. The entire middle feed line wiring has a comb-like shape. The size of each quarter-wavelength impedance transformer is adjusted to bring each voltage phasor amplitude to the corresponding subarray’s center port, following the distribution rules explained in Subsection 2.2. In addition, as mentioned earlier, the subarray feed points at the x-axis on both sides should be placed at different ends of the corresponding U-shaped section to realize the sum pattern in the H plane. This out-of-phase feature is achieved by setting the length of the feed lines that connect both sides of the lower layer probe as follows:

$$l_2 - l_1 = \frac{\lambda_g}{2}, \quad (4)$$

where $\lambda_g$ denotes the guide wavelength of the strip-line at 35 GHz center frequency.
3.3 Integrated model and dimension parameters
The front end of the 2-D planar array was integrated with the strip-line back end to obtain the entire antenna structure, as presented in Fig. 6. As indicated above, this design adopts a custom composition of a microstrip array radiation layer and strip-line feeding layer comprising two parts. Considering the soft materials of the Rogers RT-5880 substrate used in our design, a hard aluminum plate with a thickness of 4 mm was pasted to the bottom ground to ensure ideal array flatness, which introduced a vacuum coaxial cavity section formed by the coaxial converter probe and aluminum plate hole. The resulting transmission mismatch between the two media was improved by adjusting the bottom ground hole diameter \( d_1 \); thus, a preferable reflection coefficient was achieved across the device operating band of 34.5–35.5 GHz. Each optimized parameter is summarized in Table I, and the simulation results under these parameters are presented in Subsection 4.2 for comparison with the homologous measurement results.

4. Fabrication and measurement

The antenna processing prototype is illustrated in Fig. 7. In this model, a series of nylon screws with a diameter of 2 mm was used to press the multilayer PCB and duralumin back-up plate together. A TACONIC HT-1.5 bonding film layer with 0.1 mm thickness was added between the top microstrip and upper strip-line substrate as well as between the upper and lower strip-line substrates in the multilayer PCB processing. It is worth noting that the probe should be trimmed appropriately to ensure that it can be inserted into the blind hole of the strip-line bottom layer effectively to achieve a better connection between the coaxial cable and antenna. The insulating properties of these nylon screws have little influence on the antenna radiation performance, which can be illustrated by the following comparison between the simulation and measured results.

![Fig. 6 Configuration of the antenna structure: (a) top view and side view, (b) top view of microstrip ground and bottom ground.](image)

| Table I Antenna structure parameters (mm). |
|-------------------------------------------|
| Para. | \( W'_1 \) | \( W'_2 \) | \( W'_3 \) | \( W'_4 \) | \( W'_5 \) | \( W'_6 \) |
| Value  | 1.376 | 1.714 | 2.512 | 3.325 | 3.8 | 0.424 |
| Para. | \( W'_7 \) | \( W'_8 \) | \( W'_9 \) | \( W'_{10} \) | \( W'_{11} \) | \( W'_{12} \) |
| Value  | 0.2 | 0.6 | 0.08 | 0.248 | 0.128 | 0.1 |
| Para. | \( L_1 \) | \( L_2 \) | \( L_3 \) | \( L_4 \) | \( L_5 \) | \( L_6 \) |
| Value  | 2.693 | 3.1 | 1.1 | 1.35 | 4.95 | 5.18 |
| Para. | \( L_7 \) | \( L_8 \) | \( L_9 \) | \( L_{10} \) | \( l_1 \) | \( l_2 \) |
| Value  | 1.445 | 4.335 | 1.445 | 1.445 | 8.3 | 11.19 |
| Para. | \( d'_1 \) | \( d'_2 \) | \( d'_3 \) | \( d'_4 \) | \( h'_1 \) | \( h'_2 \) |
| Value  | 0.4 | 0.3 | 0.1 | 0.69 | 0.508 | 0.508 |
| Para. | \( h'_3 \) | \( L'_4 \) | \( W'_7 \) | \( L'_5 \) |
| Value  | 3 | 80 | 45 | 5.78 |

![Fig. 7 Photograph of the fabricated model.](image)

The measured reflection coefficient sweeping curve drawn using the Agilent-N5246A PNA-X network analyzer and two groups of simulated results for comparison are shown in Fig. 8; the values are lower than -10 dB and -15 dB, respectively, across the entire working band. First, in comparison with the ideal
simulation model without bonding layers, the modified result obtained with the bonding film deteriorates and rises by approximately 5 dB on average when compared with the ideal curve. The introduction of the bonding films changes the equivalent circuit structure of the transmission line, as shown in Fig. 6. From the signal transmission perspective, the non-ideal dielectric constant between the substrate and bonding material causes energy reflection in the antenna structure; hence, impedance matching deteriorates as the frequency increases, which can also be observed in the measured result. However, when compared with the simulation results, the measured results show deterioration and fluctuation to some extent. During process debugging, two main reasons were identified for the difference in the measured curve. One was the unsatisfactory fluctuation in the port calibration results obtained using KEYSIGHT N4694A under the matching load connection situation. Second, the imperfectly trimmed and hard extruded probe during antenna assembly as well as the coaxial adapter used in the measurement contributes to the overall mismatching and fluctuation.

Fig. 8 Simulated and measured results of reflection coefficient sweeping curve.

The antenna patterns in two orthogonal cut planes were measured in a far-field anechoic chamber. The normalized results are presented in Fig. 9. The HPBW in the E plane approaches 8°, which is slightly narrower than the simulated result. FSLL of approximately −19 dB, which is close to the simulation result, is achieved simultaneously. High consistency between the simulated and measured main lobes can be observed in the H plane; the measured FSLL is −17 dB and deteriorates slightly when compared with the simulated value. Undesirably, the measured level is 8 dB greater than the simulation result, which is beyond the scope of the main lobe. The main reason for this surprising difference is the probe position error introduced in the antenna assembly procedure, which causes unsatisfactory excitation phases of the four subarrays, thereby significantly enhancing the unexpected radiation toward the z direction. The peak

![Fig. 9 Simulated and measured results of radiation patterns at 35 GHz center frequency: (a) E plane patterns, (b) H plane patterns.](image)

| Ref. | Array layout | Substrat e | Operating band | FSLL (dB) | Gain (dB) | FBR (dB) |
|------|--------------|------------|----------------|-----------|-----------|---------|
| [5]  | 4×1 aperture feed | Three layers | LS − XC | −10 | 12.65 | Not given |
| [10] | 4 microstrip feed 3×3 | Single layer | Ka | −11.8 | 12.8 | 28 |
| [11] | 4 microstrip coupling feed | Single layer | C | −28 | 12.8 | 34 |
| [15] | 4/4 CPW feed 2×2 via hole feed 10/8 SIW feed 21/4 microstrip feed 8×8 | Single layer | K | −13 | 14.8 | Not given |
| [16] | 4 layer | Single layer | M | −10 | 16.7 | 28 |
| [17] | 8 layer | Dual layer | X − Ku | −8 | 11.52 | 15 |
| [24] | 8 layer | Single layer | W | −15 | 18.88 | 30 |
| [25] | 8 layer | Three layers | X | −15 | 24.4 | Not given |
| [26] | 2 layer | Dual layer | K | −17 | 18 | Not given |
| [27] | 2 layer | Single layer | C | −20 | 14 | 28 |
| This work | 4 stripline feed | Dual layer | Ka | −17 | 19.2 | 40 |
cross-polarization level reaches −20 dB across the omnidirectional scope in both the E and H planes. Moreover, owing to the use of the strip-line and duraluminum plate in our design, the measured front-to-back ratio (FBR) and forward gain were 40 dB and 19.2 dB, respectively, which are ideal. The performance of this study and other representative outcomes are summarized in Table II for comparison.

5. Conclusion

This paper presented a planar microstrip array antenna composed of 10×4 compact hybrid-fed rectangular patches. The tapered patch distribution and loading impedance transformer method were used in the front and back end designs, respectively, to achieve FSLLs lower than −17 dB in both the E and H planes. Suppressed backward radiation, as proven by the measured FBR of 40 dB, was obtained because of the strip-line feeding structure. The ideal FSLL and FBR values resulted in the measured peak gain of 19.2 dBi achieved in the 34.5–35.5 GHz working band, which shows the immense potential of the proposed design in miniaturization or portable MMW wireless applications.

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