Hybrid antenna arrays with high angular resolution for 77 GHz automotive radars

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Abstract An antenna array system with high angular resolution is proposed to adapt the demands of both medium-range radar (MRR) and long-range radar (LRR) detections for 77 GHz automotive radars. Both the MRR and LRR modes are integrated into one substrate based on the optimized sparse array topology, which makes full use of the antenna aperture size to improve the angular resolution of the proposed system. Two-dimensional series-fed weighting arrays are designed via the Taylor synthesis method to effectively heighten the antenna gain and restrain the sidelobe level. After completing the fabrication, measurement results of the proposed antenna array are in good agreement with the simulation results. Moreover, the angular resolution is verified to be 0.5° by adopting the coherent signal space method (CSM) with stepped frequency transmitting waveform, which validates the effectiveness of the proposal.

key words: 77 GHz automotive radar, high angular resolution, medium-range radar, long-range radar, sparse array, angular measurement

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

1. Introduction

Automotive radars have attracted great attention due to their safety precautions [1, 2, 3, 4]. Nowadays, the 77 GHz radars represent the development trend of future automotive radars [5, 6, 7, 8]. According to detection range, automotive radars can be divided into three types, i.e., short-range radar (SRR) [9], medium-range radar (MRR), and long-range radar (LRR) [10]. Generally, different field of views (FOVs) are required for the transmitting antennas in different types of automotive radars. Since both MRR and LRR are exploited to detect the forward targets, it is effective and meaningful to integrate these two antennas on a single substrate [5, 11, 12].

Historically, different antenna types have been used in automotive radars [13, 14, 15]. For instance, [16] investigated an antenna with two pyramidal horns and dielectric lenses for a long-range automotive radar. [17] proposed an antenna structure with a horn array, a lens and a prism. Horn antennas with lens or reflectors usually perform a high gain [18], however, the bulky structure may limit their applications.

Moreover, patch array antennas are convenient in design owing to their simple structures, compact sizes, and ease of integration [11, 19, 20, 21, 22]. Among them, [20] reported a null-filling antenna with four gap-coupled antenna elements and 14 direct-coupled antenna elements. [11] and [21] introduced the substrate integrated waveguide slot array with flat-shoulder-shaped radiation pattern. In particular, series-fed antenna array with weighting method was presented with low E-plane sidelobe, which is helpful to suppress the interference of sidelobe clutter to vehicle detection [23].

In addition, the angular resolution is a critical requirement for automotive radars. [24] presented that the multiple-input multiple-output (MIMO) antennas have better angular resolution than traditional phased-array antennas. [25] employed a uniform linear array consisting of 5 switching-transmit and 12 receive antennas to achieve a high angular resolution. It is known that the angular resolution is related to the aperture of antenna array [5]. Thus, a larger antenna system allows higher angular resolution. However, the increased antenna size makes the automotive radars more difficult to integrate. Since the antenna placement strategy affects the FOV and the angular resolution in direction of arrival (DOA) estimation, sparse antenna arrays can be applied to improve the angular resolution with a smaller antenna dimension [24].

In this letter, a novel antenna array for medium- and long-range radar (MLRR) applications at 77 GHz is presented. By integrating these two models on a single substrate, the MRR antenna is a sparse MIMO array with three transmitters, whereas the LRR antenna is a sparse single-input multiple-output (SIMO) array with one transmitter. Both the MRR and LRR share a common receiving structure of four receivers. Through optimizing the sparse array topology, the proposed antenna can achieve a high angular resolution of 0.5°. The measurement results are consistent with the simulation results, which verify the effectiveness of the proposed antenna and the superiority in angular resolution.

The letter is organized as follows. Section 2 elucidates the detailed design and simulations for the proposed MLRR antenna array. The performance measurement results are presented and analyzed in Section 3. Finally, Section 4 concludes the letter.
2. Antenna system design and simulation

2.1 Preliminary

Considering $M$ transmitting antennas and $N$ receiving antennas, they can form $M \times N$ transmit-receive channels. Assume the position coordinates of each transmitting and receiving antenna are $\vec{x}_t = [x_{t1}, x_{t2}, ..., x_{tM}]^T$ and $\vec{x}_r = [x_{r1}, x_{r2}, ..., x_{rN}]^T$, respectively. Thus, the array steering vector $\vec{a}(\theta)$ can be expressed as:

$$\vec{a}(\theta) = g(\theta) \cdot e^{j k_0 \vec{x}_r \sin \theta} \otimes e^{j k_0 \vec{x}_t \sin \theta}$$  

(1)

where $g(\theta)$ denotes the normalized gain of each element. Let the azimuths of two targets be $\theta_i$ and $\theta_j$ respectively, and the ambiguity function (AF) between arbitrary two steering vectors is used to measure the resolution of the antenna array to any two directions, given by:

$$AF(\theta_i, \theta_j) = \frac{\left| \vec{a}^H(\theta_i) \cdot \vec{a}(\theta_j) \right|}{\left\| \vec{a}(\theta_i) \right\| \cdot \left\| \vec{a}(\theta_j) \right\|}$$  

(2)

As a binary function respecting the angles, its value belongs to $0 \leq AF(\theta_i, \theta_j) \leq 1$. When the maximum value is 1, the antenna array cannot distinguish the two incoming wave directions of $\theta_i$ and $\theta_j$ completely; however, when the minimum value is 0, the array has the strongest resolution. Since the $AF$ of antenna array relates to the position of each element, a multivariate objective function is established to obtain the optimal array element distribution as $f = \vec{f}(\vec{x}_t, \vec{x}_r)$. In our antenna array, the genetic algorithm (GA) [26] is invoked to get the optimal solution of this objective function under specific constraints. More details can be found in our previous work [27]. By solving this problem, the scheme of evenly-arrayed transmitting antenna and sparsely-arrayed receiving antenna is selected as the optimal option.

2.2 Transceiver array

Fig. 1(a) exhibits the proposed MLRR radar antenna array in this letter. As shown in Fig. 1(b), the MRR transceiver array is composed of one-dimensional microstrip series-fed weighted arrays. To be specific, T1, T2, and T3 constitute the MRR transmitting array, and R1, R2, R3, and R4 constitute the receiving arrays. It is visible that the transmitting array is uniform whereas the receiving array is sparse. Starting from the first transmitting array, the positions of components T1, T2, T3, R1, R2, R3, and R4 are $0, 1.54\lambda_0, 3.08\lambda_0, 4.11\lambda_0, 7.96\lambda_0, 17.71\lambda_0$, and $20\lambda_0$ respectively, which are optimized by the GA algorithm. As shown in Fig. 1(c), the
LRR transmitting array is comprised of two-dimensional microstrip series-fed weighted array T4, and the receiving array includes R1, R2, R3, and R4. It implies that MRR and LRR share the same receiving array. As a series-fed weighted array designed by Taylor synthesis method, the whole antenna array can accomplish the desired excitation distribution by adjusting the width of each cell patch [28].

In this antenna system, the patch width weighting method with single feed is employed to design the receiving and transmitting antennas. Among them, the one-dimensional microstrip series-fed weighted antenna array consists of 16 element patches, which are connected by the microstrip line with 100 Ω. The position of the feeding point is shown in Fig. 2. Since the characteristic impedance of the main feeder selected in this design is 50 Ω, a quarter wavelength impedance matching section needs to be added between the main feeder and the 16-element series-feed array to achieve the effect of adjusting the impedance matching. Here, the width of center unit patch is \( \text{patch}_w = 1.54 \text{ mm} \), and the length is \( \text{patch}_l = 1.24 \text{ mm} \), which are calculated based on the theoretical formulas of microstrip patch antenna. As shown in Fig. 2(b), the two-dimensional microstrip series-fed weighted array adopts a series feed network to feed the \( 8 \times 16 \) series-fed patch arrays. The method of impedance matching segment weighting [27] is utilized in the series feed network of the horizontal dimension (zoy plane). By adjusting the impedance value of the quarter wavelength impedance matching segment, the feed to each linear array can be regulated, so as to reduce the sidelobe level in the radiation pattern of the horizontal dimension. In order to reduce the influence of the horizontal clutter, the horizontal dimension is also weighted according to the Taylor synthesis method with \( -30 \) dB. The excitation distribution of the 8-element in the horizontal dimension can be calculated by MATLAB, and the linewidth of each impedance matching section can be obtained according to the calculation formulas of characteristic impedance for microstrip line [29]. The spacing of each array is \( d = 2.8 \text{ mm} \) in this design.

By optimizing the parameters, the simulation results of the input standing-wave ratio (SWR) of one-dimensional microstrip antenna array based on the patch width weighting method and the simulation patterns of E-plane and H-plane are shown in Fig. 3. It can be seen from the simulation results that based on the patch width weighting method, the resonance frequency of one-dimensional microstrip antenna array is located at 77 GHz, and the bandwidth of S11 < \(-10 \) dB is about 0.8 GHz. The gain of single line array antenna is 18.52 dB, the sidelobe level of E-plane is \(-24.5 \) dB, the beam widths of E-plane and H-plane are 6.18° and 69.8°, which meets the requirements of MRR for the FOV.

![Fig. 3. Simulation results of one-dimensional microstrip antenna array based on patch width weighting method. (a) S11. (b) Far-field pattern.](image)

Furthermore, according to the aforementioned design parameters, the input SWR, E-plane and H-plane pattern of the two-dimensional microstrip series-fed weighted antenna array are simulated as shown in Fig. 4. From these figures, it can be observed that the simulation gain of the two-dimensional microstrip series-fed weighted array is 25.69 dB, the side lobe level of E-plane is \(-20.8 \) dB, and the beam width of E-plane is 5.93°, while these two values of H-plane are \(-20.86 \) dB and 12.18°. Specifically, the main lobe of
E-plane deviates from the central axis by 0.6°, which also matches the requirements of LRR for the FOV.

2.3 Physical processing
With the aid of a simple broadband waveguide microstrip transition feed structure [30], the proposed MLRR antenna array can be formed by integrating the LRR and MRR systems on the same substrate. Fig. 5 presents the test assembly drawing of antenna experimental prototype. The substrate adopts Rogers5880 plate with thickness of 0.127 mm. The actual array size of the MLRR antenna is 80 mm × 43 mm in horizontal and pitch, respectively. In the test panel, four transmitting antennas T1, T2, T3, T4 and four receiving antennas R1, R2, R3 and R4 are fed by WR10 waveguide port and UG-387/UM flange interface.

Fig. 5. Assembly drawing of antenna experimental prototype.

3. Measurement results and analysis

3.1 Test on SWR
Fig. 6 gives the test results of SWR for the proposed antenna. As shown, the resonance point of S11 at each port is shifted to 77.5 GHz, and the S11 value is less than −10 dB from 76 GHz to 78 GHz.

Fig. 6. SWR test results of each port.

3.2 Test on coupling degree
Fig. 7(a) exhibits the coupling degree between the transmitting antenna T3 and the four receiving antennas R1, R2, R3 and R4 in the case of three transmitting and four receiving antennas. Similarly, Fig. 7(b) illustrates the coupling degree between transmitting antenna T4 and four receiving antennas R1, R2, R3 and R4 in the case of one transmitting and four receiving antennas. It can be seen from the results that the coupling degree between the transmitting antenna T3 and the four receiving antennas is less than −72 dB in the band of 76~78 GHz, and the coupling degree between the transmitting antenna T4 and the four receiving antennas is less than −87 dB.

3.3 Test on pattern
Fig. 8(a) provides the test results of the far-field pattern for R3, R4, T1 and T3 at 77 GHz. It is evident that the H-plane beam widths of T1 and T3 are relatively wide, i.e. 88.3° and 82.9° respectively. However, the H-plane beam widths of R3 and R4 are relatively narrow, both about 60°. Fig. 8(b) compares the far-field pattern of H-plane and the simulation data of the two-dimensional microstrip series-fed weighted antenna T4. It can be concluded that the H-plane beam width of the transmitting antenna T4 is 10.5°.

3.4 Test on angular resolution
To verify the angle measurement performance of the proposed antenna, the wide band coherent signal space method (CSM) [31] is adopted for DOA estimation. The stepped frequency signal is adopted as transmitting waveform, since it can be easily generated by vector networks. For this reason, the antenna receiving and transmitting links excited by stepped frequency signals can be simulated by setting step frequency mode in vector networks. Each transmitting/receiving antenna forms a transceiver channel. In this way, the echo I/Q data of each transmitting and receiving channel is equal to S21 data of vector networks.
As shown in Fig.9, two metal pillars are vertically placed as targets with a distance of 2.6 cm. The height of the center point of the metal pillars is equal to that of the antenna arrays. By defining the center point of the antenna arrays as the origin O, the normal of the plane of the two metal pillars is coincident with the normal of antenna arrays. The angles between two metal pillars and the normal at the O-point are $\theta_{11} = 0.25^\circ$ and $\theta_{12} = -0.25^\circ$, respectively. The distance from each transmitting/receiving antenna to the target can be approximated as $R_{tm} \approx R - x_{tm} \sin \theta$, $m = 1, \ldots, M$, and $R_{rn} \approx R - x_{rn} \sin \theta$, $n = 1, \ldots, N$, where $M$ and $N$ stand for the numbers of transmitting and receiving arrays respectively. Then, the transmission coefficient of each transmitting and receiving channel can be measured by the vector network as [31]:

$$S_{21}(f)|_{T_{m}, R_{n}} \approx A_{mn} \cdot e^{j2\pi f(R_{tm} + R_{rn})}$$

$$\approx A \cdot e^{j2\pi f\left[R - (x_{tm} + x_{rn}) \sin \theta\right]}$$

where $c$ is the velocity of light, $A_{mn}$ and $A$ refer to the echo amplitudes. As a result, the azimuth of targets can be estimated with the wideband CSM algorithm.

Fig.10 show the estimated results of targets azimuth spectrum for the MRR and LRR systems, respectively. It can be seen that two metal cylinder targets with angle interval can be distinguished through the DOA algorithm processing. However, the spatial peak positions slightly deviate from the actual values $\theta_t = \pm 0.25^\circ$. The peak of targets shifts to right, which may be caused by the fact that the antenna phase center varies with different channels.

### 3.5 Comprehensive comparison

Finally, the proposed antenna array is compared with several reported 77 GHz antennas for automotive radars. As summarized in Table I, our antenna obtains approximately 2.1 dB higher than [11] with both MRR and LRR modes, due to the use of two-dimensional series-fed weighting method. However, the increase of antenna gain leads to a decrease of azimuth with 10.5° at the LRR mode. The antenna with horn and lens in [16] yields a better antenna gain and lobe-width, however, its bandwidth is the narrowest below 1 GHz. Compared with [25], our antenna outperforms in angular resolution with 0.5°, revealing that the sparse array can increase the aperture utilization and further improve the antenna angular resolution.

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4. Conclusion

In this letter, a hybrid MLRR antenna array combining the MRR and LRR systems is proposed for automotive radars. It consists of two array forms, i.e., three-transmitter-four-receiver for the MRR scenario and one-transmitter-four-receiver for the LRR scenario, with one-dimensional and two-dimensional weighted series-fed antenna arrays, respectively. By utilizing the sparse array optimization method, the multi-mode antenna system can realize a high angular resolution performance. Through simulations and measurements, the angular resolution of 0.5° of the proposed antenna array is verified via the CSM algorithm and the stepped frequency transmitting waveform. As the angular resolution is greatly improved, it is preferable for automotive radars to recognize small road targets, e.g., pedestrians and non-motor vehicles, and estimate the contour and direction of the vehicles ahead.

References