CPW Center-Fed Single-Layer SIW Slot Antenna Array for Automotive Radars

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Abstract—A compact co-planar waveguide (CPW) center-fed substrate-integrated-waveguide (SIW) slot antenna array is proposed to achieve narrow H-plane beamwidth and low sidelobe levels for automotive radars such as Across-the-Road (ATR) Radar. The antenna consists of an array of SIW slot elements, a CPW-SIW transition and a CPW power divider. A new type of CPW-SIW transition is proposed to minimize the blockage aperture of the slot antenna array for suppressing the sidelobe levels of the array. A parallel center feeding configuration is applied to avoid beam squinting. The antenna prototype printed onto a single-layer Rogers 5880 is with $32 \times 4$ slot elements and an overall size of $195 \text{ mm} \times 40 \text{ mm} \times 0.79 \text{ mm}$. Measured results show that the proposed antenna exhibits gain of > 22.8 dBi, efficiency of > 67%, return loss of > 10 dB, sidelobe level of < -21 dB, and a fixed boresight beam of < 4.6° in H-planes over 24.05 – 24.25 GHz, suitable for ATR radar.

Index Terms—Co-planar waveguide (CPW), substrate integrated waveguide (SIW), slot antenna array, sidelobe, fixed beam, single-layer, Across-the-Road (ATR) Radar, automotive radar.

I. INTRODUCTION

A ntenna is one of the key components in automotive radar applications [1]–[3]. Across-the-Road (ATR) radar, a sensor speeding the speed of moving vehicles, is such an example. The specifications of a K-band ATR radar in U. S. are given in [4]. The assigned frequency band is 24.05 GHz to 24.25 GHz, or 0.8% operating bandwidth. The critical electrical requirements for ATRs include half power beamwidth (HPBW), sidelobe levels (SLLs), and beam direction. Specifically, the HPBWs in the horizontal and vertical planes are required to be less than 6° and 20°, respectively. The SLLs in the horizontal plane should be less than the -20 dB. In addition, the primary radar beam should be centered horizontally and vertically without beam squinting across the bandwidth.

Microstrip antennas have well-known advantages such as low cost, low profile, and easy to integrate with active circuits. Microstrip antenna arrays have thus been reported to apply in automotive radars with good electrical performance [5]–[8]. The main concerns of the microstrip antennas come from the undesired radiation and surface wave coupling/loss. To alleviate the surface wave mode, single-layer microstrip antennas usually have to be printed onto thin substrates [8]. With a thinner substrate, however, the microstrip antenna has more severe conductor and dielectric losses, a narrower bandwidth, and less mechanical robustness [8]–[9].

Alternatively, substrate integrated waveguide (SIW) [10]–[12] or post-wall waveguide [13] fed antennas feature the same advantages of microstrip antennas because they can be fabricated with the same planar printed technology. Furthermore, as a waveguide-like structure, SIW does not suffer from the unintentional radiation and surface wave loss, which alleviates the limitation of the thin substrate. These merits offer the SIW antennas the possibility of achieving high efficiency. A SIW slot antenna with SLLs as low as -30 dB in both E- and H-planes have been achieved [10]. However, the end-fed antenna array suffers from beam squinting against operating frequency, which does not meet the center beam requirement of the ATR radar.

To achieve the required unchanged maximum radiation direction across the operating bandwidth, a center-fed antenna array is preferred [13]–[15]. However, it is difficult to realize a conventional single-layer center-fed waveguide-based slot antenna array to achieve low SLLs in the H-plane because of the large aperture blockage (slot-free area) in the center portion of the antenna array. Some studies have reported to reduce the aperture blockage effect and lower the SLLs [13]–[15]. In [14], the SLL of -9.5 dB associated with the aperture blockage is improved to -14.7 dB by applying a genetic algorithm to control the slot excitation distribution. In [15], the first SLL is reduced from -10 dB to -13 dB using the $E$- to $H$-plane cross-junction power dividers. A post-wall waveguide feeding network for a center-fed antenna array has been reported [13]. The large blockage area results in a high SLL of -7.8 dB but is suppressed to -11.1 dB using a tapered amplitude distribution. Generally, the width of a post-wall waveguide or SIW is wider than that of a microstrip or co-planar waveguide (CPW), which may be undesired in the design of antenna feeding network.

In this paper, a CPW center-fed SIW slot array antenna is proposed to achieve a low SLL, narrow and fixed beam in the $H$-plane for the 24-GHz ATR radar using normal low-cost single-layer printing circuit board (PCB) process. We first present the design of the slot array and investigate the effects of the blockage area on the SLLs of the antenna. With the motivation of reducing the blockage area for low SLLs, we employ a compact, parallel CPW center feeding structure to avoid the undesired beam squinting with frequency. Finally, the measured results of the antenna array are discussed and compared with the simulation using CST Microwave Studio [16].
II. ANTENNA STRUCTURE AND DESIGN

A. Antenna Configuration

Fig. 1(a) shows the top view of the SIW slot array antenna. There are totally 4 × 32 slots on the broadwall of the SIWs. In each row, two 16-element linear arrays are placed end to end with a distance of $d_1$ between the two starting slots. The central portion of the antenna is occupied by the feeding structure and is thus slot free, which is the bottleneck of sidelobe suppression. The effects of $d_1$ on radiation pattern will be discussed in next subsection. The spacing between the adjacent slots in the $H$-plane, $d_h$, is $\lambda_g/2$ ($\lambda_g$ is the guided-wavelength of the SIW). And the spacing between the adjacent slots in the $E$-plane, $d_e$, is $\lambda_w$ ($\lambda_w$ is the guided-wavelength of the CPW) so that the adjacent linear arrays are fed in phase. The detailed geometrical dimensions of the antenna are tabulated in Table I, in which $x_{i,j}$ for $i \in \{1, \ldots, 16\}$ and $l_{i,j}$ for $i \in \{1, \ldots, 16\}$ are the offsets and lengths of the slots.

Fig. 1(b) shows the bottom view of the overall slot array antenna. The input of the array is connected to an external commercial available mini-SMP connector. The compact CPW feeding network is located in the center of the antenna with an eight-way parallel feeding configuration.

The antenna is printed onto a single-layer PCB of Rogers 5880 with $\varepsilon_r = 2.2$ and $\tan \delta = 0.0009$ at 10 GHz. The thickness of the substrate is $t = 0.79$ mm. The conductor used for metallization is copper with a conductivity of $5.8 \times 10^7$ S/m and a thickness of 0.02 mm.

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B. Design and Analysis of Slot Array

The design of the slot array follows the method in [10] and is briefly described here. First, the parameter is extracted for a single slot on the SIW with various offsets. Resonant length, resonant conductance, and admittance are obtained for the array synthesis. Next, Elliott’s iterative procedure for a waveguide-fed slot array [17], including all mutual couplings, is applied for the SIW-fed linear array to calculate initial slot parameters for a targeted amplitude distribution. Further fine tuning by electromagnetic (EM) simulation finalizes the slot parameters for desired SLLs. In order to achieve the requirement of -20 dB SLL in the $H$-plane, -26 dB Taylor distribution is chosen. To achieve the $<6^\circ$ HPBW in the $H$-plane, 32 slots are used.

In the simulation of the 32-element linear array as shown in Fig. 2, two ports excite each half of the slot arrays

Fig. 1. Top and bottom view of the CPW fed SIW slot array antenna.
The spacing between slots 16 and 17, \( d_1 \), is first set to be \( \lambda_g/2 = 5.8 \) mm. In this case, there is no blockage because \( d_1 \) equals to the spacing of all other adjacent slots, \( d_6 \). The distance between the ends of slots 16 and 17 is \( d_e = 1 \) mm. The geometric parameters of the slot array are listed in Table I. The offsets and lengths of the slots are also shown in Fig. 3. As the slot number decreases, the lengths become smaller and slots are located closer to the center of SIW so that the radiation from the “edge” slots weakens. The offsets of slots 1–4 are set a little larger than that of slot 5 for a 2-dB better SLL.

Fig. 2. Simulation model of the two 1× 16 linear arrays located end to end with different \( d_1 \).

Fig. 3. Slot offsets and lengths of the antenna array.

Fig. 4 compared the simulated amplitude distribution and the theoretical -26 dB Taylor distribution of the slots. The simulated result including all mutual coupling agrees reasonably with the theoretical one. The simulated \( H \)-plane radiation pattern is shown in Fig. 5 (\( d_1 = 5.8 \) mm). The -26 dB peak sidelobe locates at the first one. The SLLs decrease as |\( \theta \) | increases. In this case, unfortunately, \( d_1 \) of 1 mm is too small to accommodate a feeding network and necessary space between the slot and the feeding network to avoid unwanted EM coupling which affects the SLLs [15].

Therefore, there is a tradeoff between the blockage area and the requirement of realizing the feeding network. Fig. 5 also shows the \( H \)-plane radiation patterns of the linear array with different \( d_1 \) at 24 GHz. As \( d_1 \) increases from \( \lambda_g/2 \) (5.8 mm) to \( \lambda_g \) (11.6 mm), the innermost SLL changes slightly but the grating lobes in the range of \( 30^\circ < |\theta| < 75^\circ \) increase due to the enlarged blockage area. The peak SLL is -24.5 dB. When \( d_1 \) becomes 16 mm, the peak and first SLL degrades to -19 dB. To meet the requirement of -20-dB SLL, the case of -24.5-dB SLL is selected with 4.5 mm margin considering fabrication tolerance. This case with \( d_1 \) of 11.6 mm results in \( d_1 = 6.84 \) mm, possible to accommodate a compact feeding network and the space to avoid undesired coupling.

Fig. 4. Amplitude distribution of the linear array.

Fig. 5. \( H \)-plane radiation patterns of the linear arrays with different \( d_1 \) at 24 GHz.

After aligning multiple linear arrays side-by-side to form the SIW planar array, the \( H \)-plane SLL of the whole planar array is almost unchanged. The reason is that SIW is a low-profile waveguide with about 8:1 width-to-height ratio so the \( TE_{20} \) mode internal mutual couplings inside the SIW are dominant among all kinds of internal and external mutual couplings [10]. After forming the SIW planar array, only external mutual couplings among slots at different branches of SIW are introduced. The effects of the additional external mutual couplings are much smaller than those mutual couplings caused by the \( TE_{20} \) mode which have already been included in the SIW linear array design.

It is known that direct tuning the hundreds of parameter of the SIW planar array in EM simulation is very time-consuming and inefficient. In the above way, one can significantly simplify the EM design process by only optimizing one row of the planar array with much less parameters.
C. Feeding Network Design

To achieve relatively smaller blockage for low SLLs, CPW, instead of a wide SIW or post-wall waveguide [15], is utilized to configure the power divider to feed the array antenna.

The first step is to select the most compact CPW-SIW transition. Several CPW-SIW transition designs have been reported and four typical ones are shown in Fig. 6 as the candidates for the compact feeding structure of this design [18]–[21]. The transition based on a dipole slot in Fig. 6(a) [18] and transition with a current probe in Fig. 6(b) use a λg/4 GCPW stub for impedance matching [19]. Another transition in Fig. 6(c) has a >λg/4 cavity resonator under the dipole slot [20]. Because all the λg/4 structures are inside the SIW, it is impossible to move them from the limited H-plane space (x-direction) to E-plane space (y-direction).

The transition depicted in Fig. 6(d) [21] is preferred wherein the dipole slot is next to the short-circuit end of the SIW so that the most compact configuration is realized. In that design, however, a λg/4 CPW impedance transformer is next to the SIW which cannot be applied directly to form an H-plane (x-direction) compact design. The impedance transformer outside the SIW can be removed along x-direction and the impedance matching network can be designed along y-direction because of more space there. The removal of the impedance transformer at each stage leads to an inter-stage impedance other than 50 Ω in the feeding network. In this way, d1 = 11.6 mm is achieved as shown in Fig. 7. Distance d1 = 2 mm is set to avoid the unwanted EM coupling between the transition and the slots which affects the slot excitation and SLLs [15]. Width of w2 = 1.9 mm on the other side of the widest CPW width used in the power divider. If the two λg/4 CPW impedance transformers were positioned along the y-direction [21], d1 would increase to 16 mm. The radiation pattern of this array configuration is shown in Fig. 5, where d1 = 16 mm leads to a higher SLL up to -19 dB.

![Fig. 6. Four types of CPW-SIW transition in [18]–[21].](image)

![Fig. 7. First stage of the feeding network.](image)

Design of the feeding network is illustrated in Fig. 8. The E-plane spacing (y-direction) between the adjacent SIW is set to be the guided wavelength of CPW at 24 GHz, λg, so that the phase of S21 equals to the phase of S31. Because of the structure symmetry, this configuration guarantees the phase balance at the eight outputs.

The input impedance of the half-wave slot dipole and the CPW with a length of l0 determine Z1 as shown in Fig. 8. If the characteristic impedance of the CPW is low, Z1 is small and Z2 (roughly Z0/4) is thus even smaller. This small Z2 requires an impedance transformer with a width exceeding the pre-assigned w2 of 1.9 mm. Here, the characteristic impedance of the CPW is set to be 83 Ω. With the dimensions in Fig. 9 (a), the simulated Z1 is (211 - j30) Ω. The imaginary part of Z1 is not necessarily zero and can be canceled out at the next stage.

The simulated Z2 = (50 + j37) Ω is not exactly Z0/4 because of the effects of the junctions, taken into account only in simulation. To match Z2 to 100 Ω, a 46 Ω CPW line with an electrical length of 30° is needed. Because the width of the 46 Ω CPW line is 1.5 mm, it leads to a wide w3 greater than pre-assigned 1.9 mm. Instead, we used a 55-Ω CPW line, with an electrical length of 38°, which is only 0.8 mm wide, for the impedance transformer. Although in this case Z2 is matched to Z3 = (110 - j4) Ω, the overall impedance matching is still acceptable as shown later. The rest of the feeding network includes two parallel 100 Ω CPW lines connecting to the 50Ω input line. Fig. 9(b) shows the detailed geometry of the rest of the feeding network.
The simulated amplitude of the S-parameters of the feeding network is shown in Fig. 10. Only four output ports, Ports 2–5 are shown because of the symmetrical structure. At 24 GHz, \(|S_{21}|\), \(|S_{31}|\), \(|S_{41}|\), and \(|S_{51}|\) are -9.64 dB, and -9.65 dB, respectively. In Fig. 11, the phase of \(S_{21}\), \(S_{31}\), \(S_{41}\), and \(S_{51}\) are 173°, 171°, 171°, and 169°, respectively. Therefore, good impedance matching and amplitude/phase balance are achieved by the proposed feeding network.

III. EXPERIMENTAL RESULTS

The antenna was prototyped and measured in a full anechoic chamber. The antenna prototype shown in Fig. 12 is with an overall size of 195 mm × 40 mm × 0.79 mm.

Fig. 8. Input impedances at various reference planes of the feeding network.

Fig. 9. Detailed geometry of the feeding network (unit: mm). (a) CPW-SIW transition, \(s = 0.2, g = 0.1, w_c = 1, l_t = 0.5, l_s = 4.3, w_{SIW} = 6.2, l_c = 0.95\), (b) impedance transformer part. \(s_1 = 1.2, s_2 = 0.3, s_3 = 0.8, g_2 = 0.25, t_1 = 2.9, t_2 = 1.25, p_s = 0.8, d_s = 0.4\).

The simulated amplitude of the S-parameters of the feeding network is shown in Fig. 10. Only four output ports, Ports 2–5 are shown because of the symmetrical structure. At 24 GHz, \(|S_{11}|\) is -18 dB, \(|S_{21}|\), \(|S_{31}|\), \(|S_{41}|\), and \(|S_{51}|\) are -9.5 dB, -9.53 dB, -9.64 dB, and -9.65 dB, respectively. In Fig. 11, the phase of \(S_{21}\), \(S_{31}\), \(S_{41}\), and \(S_{51}\) are 173°, 171°, 171°, and 169°, respectively. Therefore, good impedance matching and amplitude/phase balance are achieved by the proposed feeding network.

Fig. 10. Simulated amplitude of S-parameters of the CPW feeding network.

Fig. 11. Simulated phase of S-parameters of the CPW feeding network.

Fig. 12. Photo of the antenna array: (a) top view, (b) bottom view.
Fig. 13 shows the measured and simulated $|S_{11}|$ of the antenna array prototype. Reasonable agreement has been achieved between the measurement and simulation. The simulated $|S_{11}|$ is less than -10 dB in 23.8–24.2 GHz and the measured $|S_{11}|$ is less than -10 dB in 23.84–24.25 GHz.

Fig. 14 shows the measured and simulated boresight gain of the antenna array. In the measurement, a mini-SMP adapter was used [22]. In Fig. 14, the measured gain includes the insertion loss of such an adapter. Compared to the original simulated gain, the modified simulated gain shifting about 0.1 GHz upwards agrees better with the measured one. The difference between measured and shifted simulated gain attributes to the actual antenna loss which may be caused by lossy dielectric and higher than that in simulation. In 23.86–24.12 GHz, the measured gain is higher than 23 dBi and the maximum gain is 24 dBi at 23.92 GHz.

Fig. 15 shows the measured and simulated SLLs of the antenna array. In 23.9–24.3 GHz, the simulated SLL is lower than -22.9 dB and the lowest SLL is -25.4 dB at 23.9 GHz. The measured SLL is lower than -21 dB in 24.05–24.4 GHz. An upward frequency shift is also observed between measurement and simulation.

Compared to the simulated -24.5-dB SLL of the linear array at 24 GHz in Fig. 5, the simulated SLL of the planar array without any tuning of the slots is -24.4 dB. The SLL of the planar array is almost unchanged compared to the linear array.

The antenna efficiency takes both radiation efficiency and mismatch loss into account [23]. The simulated efficiency is higher than 70% in 23.8–24.3 GHz. Because of the agreement between the measured and simulated beamwidths, the simulated directivity (shifting 0.1 GHz upwards) and measured gain (including the insertion loss of the mini-SMP adapter) are used for the calculation of measured efficiency. The measured efficiency is higher than 67% within the range of
23.9–24.4 GHz. The high antenna efficiency attributes to the compact CPW feeding network with minimizes the path loss and the low-loss property of SIW. It is observed that the measured efficiency is shifted upwards compared to simulation as well.

![Efficiency vs Frequency](image)

From Figs. 14–17, it is observed that all the measured responses shifted upwards 0.1 GHz compared with the simulation. Therefore, in the comparison between the measured and simulated radiation patterns, we consider this frequency shift. Figs. 18–19 show the comparison of the measured and simulated radiation patterns of the antenna array in the H- and E-planes, respectively, with a +0.1-GHz shift. In addition, the main beam of the antenna keeps at boresight without any squinting as expected.

For the H-plane radiation patterns over the band from 24.05 GHz to 24.25 GHz in Fig. 18, the measured beamwidths agree well with the simulated ones. The measured inner-most SLLs increase because they are sensitive to fabrication tolerance and errors.

In Fig. 19, the measured beamwidths and peak SLLs in the E-plane are also very close to the simulated ones. However, the first sidelobes are not symmetrical in the measurement. Because of the E-plane symmetry in antenna configuration, the radiation pattern should be symmetrical as the simulation shows. Thus, the possible asymmetry in measurement and/or fabrication tolerance may lead to the asymmetry of the measured results.

![H-plane Radiation Patterns](image)

**Fig. 18.** Measured and simulated H-plane radiation patterns of the antenna array. (a) Measurement at 24.05 GHz and simulation at 23.95 GHz, (b) measurement at 24.15 GHz and simulation at 24.05 GHz, (c) measurement at 24.25 GHz and simulation at 24.15 GHz.
The measured results shown in Figs. 13-19 suggest that, the workable frequency range of the antenna is 24.05–24.25 GHz, the same as the officially assigned band of the ATR radar [4]. In this band, all the requirements of the ATR radar on HPBWs, SLLs, beam directions are well met. Furthermore, the measured gain, efficiency, and return loss are higher than 22.8 dBi, 67%, and 10 dB, respectively.

IV. CONCLUSION

In automotive radars, it is a required for an antenna array to achieve low SLL and narrow H-plane HPBW without any beam squinting. The proposed CPW center-fed planar SIW slot array antenna has been validated to be with smaller blockage. The proposed antenna has experimentally demonstrated high gain greater than 22.8 dBi, HPBW narrower than 4.6°, low SLL less than -21 dB and stable boresight radiation. The compact antenna printed onto a single-layer PCB has been with a simple structure and low fabrication cost, which is suitable for ATR radars operating at 24 GHz.

REFERENCES


