Hybrid thin film multilayer antenna for automotive radar at 77 GHz

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Abstract—A hybrid thin film multilayer antenna for automotive radar is presented in this work. A 2×8 aperture coupled stacked patch antenna array is realized on a single layer printed circuit board (PCB) using a novel thin film based approach. Using a compact 180° phase difference power divider, inter-element spacing in a 2×2 sub-array is reduced. Measurement results show a 19% (67.9–82.5 GHz) impedance bandwidth and a wideband broadside radiation pattern, with a maximum gain of 15.4 dBi realized gain at 72 GHz. The antenna can be employed in mid-range automotive radar applications.

Index Terms—Aperture-coupled antenna, millimeter-wave antennas, microstrip antennas, grounded coplanar waveguide, bi-phase power divider

I. INTRODUCTION

Millimeter waves are being used extensively for commercial applications. Recent frequency allocation of the 76–81 GHz band for automotive radar sensor applications [1] has allowed development for such sensors for driver assistance systems. An important component of radar sensors is the antenna. It determines the most important range and field of view properties of the radar sensor. The antenna design is hence a very crucial part of the sensor development.

For many commercially available sensors, single layer printed circuit board (PCB) antennas such as the microstrip patch antenna or the substrate integrated waveguide antenna are used in antenna arrays. For small substrate thicknesses, these antennas suffer from narrow impedance bandwidth [2]. Due to single sided feeds and linear arrays, such antennas also have relatively narrow radiation pattern bandwidths. Recent multilayer antenna designs employing wideband antenna elements such as the grid array antenna [3], [4], [5], when used as linear arrays fed in the center, have larger impedance bandwidths as well as larger radiation pattern bandwidths. Such designs rely on multilayer PCB and hence are more expensive and complex to manufacture than single layer PCB designs.

Recently, a hybrid approach to designing multilayer antennas was demonstrated by the authors [6]. Instead of employing a multilayer PCB, this approach uses a single layer PCB and multilayer thin films that house the complete antenna element. These thin films are attached at the radiating positions on the PCB. In addition to being simpler to manufacture, this

Figure 1: Layer construction of antenna

approach is also more flexible in terms of combination of antenna elements and feed network, as was shown in [6] where the same antenna element was integrated with three different feed networks.

This work presents a novel multilayer grounded coplanar waveguide (GCPW) fed aperture coupled stacked patch (ACSP) antenna using the explained hybrid approach. The GCPW feed network is realized on the single layer PCB, whereas the stacked aperture coupled patch is realized on a multilayer thin film. This paper is organized as follows: Section II describes the design and attachment process for the antenna. Section III describes the array design where a compact power divider is described. Measurement results are provided in Section IV. The paper concludes with Section V.

II. ANTENNA DESIGN

A multilayer low loss RF qualified thin film houses the complete antenna element. It is attached to the single layer PCB using a non-conducting epoxy based adhesive [7]. The thin film consists of two substrate layers, Rogers Ultralam liquid crystal polymer (LCP) with $\epsilon_{r,TF1} = 3.00$ and $h = 100 \mu\text{m}$ and Dupont Pyralux AP and DuPont LF adhesive stackup with $\epsilon_{r,TF2} = 3.40$ and $h = 150 \mu\text{m}$. Hence, the total thin film thickness, excluding metal layer thickness is around $250 \mu\text{m}$. The single layer PCB consists of a Rogers RO3003 substrate with $\epsilon_{r,PCB} = 3.00$ and $h = 127 \mu\text{m}$.

Figure 2: Single stacked patch antenna element (a) Patch 1 (b) Patch 2 and (c) PCB top layer

A. Element design

The radiating element is a stacked rectangular patch structure. Each substrate layer of the thin film has a rectangular patch element etched on it. A transverse slot etched on the top metal layer of the PCB couples the energy to the thin film. The feed network is also etched on the PCB. The different layers of the stacked patch single element are shown in Fig. 2.

The feed network terminates in a microstrip (MS) to grounded coplanar waveguide (GCPW) transition which has the transverse slot at the end of it. The MS line with width iw transforms to a GCPW with transformer length tl and width tw . The GCPW center conductor has a width lw . The center conductor of the GCPW is extended by length ls to improve the impedance match to the slot. The slot at the end of the GCPW has length sl and width g . A similar feeding method based on coplanar waveguide (CPW) was investigated in [8]. The main difference here is the use of a ground plane below the CPW to shield the feed network, as well as the use of fencing vias with diameter vd around the excitation slot to reduce surface waves and hence increase the efficiency of the antenna.

The thin film is placed on the PCB such that the excitation slot is aligned with the center of the patches. An originally rectangular thin film with width fw and length fl is used to house the stacked patches. The thin film is altered by removing triangular portions defined by angle α and length fs in order to minimize the effect of the film on the feed network on the PCB.

The dimensions of the stacked patches are $wp1$ and $lp1$

TABLE I: Design parameters for GCPW ACSP antenna. All length dimensions are in mm

(a) PCB parameters							
vd	sl	ls	g	lw	tl	tw	iw
0.20	1.05	0.075	0.075	0.50	0.25	0.75	0.10

(b) Thin film parameters							
fw	fl	$wp1$	$lp1$	$wp2$	$lp2$	fs	α
2.1	2.1	0.60	0.80	1.30	0.8	0.71	45°

for Patch 1 and $wp2$ and $lp2$ for Patch 2 respectively. The length and width of the patch elements, along with that of the excitation slot was optimized to result in two impedance resonances which enable a wideband impedance match. The coupling effect of each resonance was adjusted to achieve maximum impedance bandwidth. The final values of all design parameters are listed in Table I.

Figure 3: Reflection coefficient and gain of single stacked patch element

Simulated reflection coefficient and gain results for a single stacked patch element are shown in Fig. 3. In the reflection coefficient result, two resonance dips can be seen, due to the coupled resonances of the stacked patches. A wideband impedance match of 15% (73.4–85.4 GHz) is achieved. The simulated gain of the single element is a maximum of 5.5 dBi at 75 GHz and remains within 1 dB of the maximum from 73–85 GHz. Farfield results are not shown, but they show an expected broadside beam in the E- and H-Planes. A beam squint of upto 6° is however observed in the E-Plane over the operating frequency range. This can be attributed to a slightly asymmetric electric field on the radiating edges of the patch due to the asymmetric ground plane on the PCB top layer below the patch.

B. Element placement

As mentioned earlier, the thin film needs to be placed on the PCB such that the excitation slot is aligned to the center of the patches. This requires an accurate thin film placement mechanism. It is achieved using a die-bonder [9]. An epoxy based non-conductive paste adhesive with a thickness of about 10 μm is first deposited by the integrated dispenser of the

die bonder at the position where the thin film will be placed. Additional structures to assist the pick and place head and integrated camera of the die bonder are etched on the bottom substrate layer of the thin film, as well on the top layer of the PCB. A positioning accuracy of $\pm 7 \mu\text{m}$ in the x and y axis can be achieved using this process, which is within the allowed maximum misalignment for optimum antenna operation.

III. ARRAY DESIGN

Figure 4: Microstrip-Slotline compact power divider top (a) and bottom (b) view

A 2×8 array was designed using the described radiating element. At the design frequency of 77 GHz the free space wavelength is $\lambda_0 = 3.90 \text{ mm}$ and the wavelength in the PCB substrate is $\lambda_g = 2.24 \text{ mm}$. For the E-Plane radiation pattern, the antenna elements were placed in opposite direction. For constructive interference in the broadside direction, a 180° phase difference is required between these two elements. This is conventionally achieved through a $\lambda_g/2$ length extension to one of the feed lines. In the compact design presented however, such a line extension would result in a large element spacing of about λ_0 and hence the introduction of grating lobes in the E-Plane pattern.

In order to maintain a compact inter-element distance, a novel power divider based on a microstrip to slotline transition is designed to feed a 2×2 sub-array. It is based on the bi-phase divider shown in [10] which was designed for lower frequencies. Essentially, the power divider consists of a microstrip to slotline transition, followed by a length of slotline, which again transitions back to microstrip. After the second transition, the opposite phase signals of the slotline couple to the microstrip lines going in opposite directions to each other and hence the output lines have 180° phase. The designed power divider is shown in Fig. 4. It operates over a wide frequency range. The simulated results of the power divider are shown in Fig. 4. Since it is a symmetrical design, only results for output ports on one side of the power divider are shown. It can be seen that a wideband equal power division is achieved for both output ports, with low reflection coefficient at the input. In addition, a constant 180° phase difference is achieved between the output ports while keeping the power divider compact. By using this power divider, element placement in both E and H-Planes of $de = 2.52 \text{ mm}$ ($0.65\lambda_0$) and $dh = 2.60 \text{ mm}$ ($0.67\lambda_0$) respectively is achieved for the antenna array.

Figure 5: S-Parameters magnitude (a) and phase (b) of power divider

Figure 6: Top view of 1×4 power divider for array

TABLE II: Design parameters for power divider. All length dimensions are in mm

(a) 2×2 power divider

<i>cd</i>	<i>sg</i>	<i>sd</i>
0.60	0.10	0.66

(b) 1×4 power divider

<i>wi</i>	<i>wti1</i>	<i>lti1</i>	<i>wti2</i>	<i>lto2</i>	<i>wto2</i>	<i>lti2</i>
0.10	0.27	0.73	0.35	0.65	0.20	0.63

Additionally, a conventional corporate 1×4 microstrip power divider is designed to feed each 2×2 sub-array. Consisting of quarter-wave transformers with different output width lines, it does equal power division in the first stage followed by unequal power division in the second stage, providing more power to inner ports. Unequal power division is done to achieve amplitude tapering over the array, thereby reducing

side-lobes in the E-Plane. The 1×4 power divider design is shown in Fig. 6. The design parameters for both power dividers are listed in Table II.

IV. MEASUREMENT RESULTS

Figure 7: Manufactured antenna array for pattern measurement complete (a) view and zoomed in microscope view (b) of 2×2 sub-array

The described 2×8 array was manufactured. Two different PCB designs were manufactured. In one design, the array was connected to a GSG-probe transition to enable S-Parameter measurements using a wafer-prober based setup. In the second design, the array was connected to a microstrip to waveguide transition to enable antenna pattern measurement using a waveguide based measurement setup. This manufactured design is shown in Fig. 7. The waveguide transition has been designed and characterised in [11].

The simulated and measured results for the reflection coefficient and realized gain can be seen in Fig. 8. Also shown are the results of the simulation model during the design process (**Sim. Refl.** and **Sim. Gain**), as well as the adjusted simulated results (**Sim. adj. Refl.** and **Sim. adj. Gain**) after observing measured results. It was observed in the measured reflection coefficient results and realized gain results, as well as farfield results (shown in Fig. 9) that the operating frequency of the antenna is approximately 5 GHz lower than

Figure 8: Simulated and measured reflection coefficient and realized gain results for GCPW ACP antenna array

Figure 9: Simulated and measured farfield radiation pattern results at (a) 71 GHz (b) 73 GHz and (c) 76 GHz in the E-Plane and H-Plane of the 2×8 antenna array

the designed operating frequency of 77 GHz. It is known from literature that a shift between simulated and measured frequency performance often results due to a discrepancy in substrate dielectric constant values. As mentioned in Section II, the thin film substrate TF Sub2 consisting of a Dupont Pyralux AP and DuPont LF adhesive stackup was modelled initially as a homogeneous substrate with $\epsilon_{r,TF2} = 3.40$. However, the dielectric constant value provided by the manufacturer in the data sheet is only measured up to 20 GHz. Similarly, the epoxy adhesive was modelled initially as an air gap with thickness of 10 μm , since its dielectric constant value had not been measured at high frequencies [7] either, and it had minimal thickness compared to the other substrate thicknesses.

After modelling the thin film substrate TF Sub2 as a homogeneous dielectric with $\epsilon_{r,TF1} = 3.70$ and the epoxy adhesive as a homogeneous substrate with $\epsilon_{r,adh} = 3.00$, the model was simulated again. It can be seen that a better agreement between the measured and simulated reflection coefficient results is achieved with the adjusted simulation model. Two resonance dips, as expected from the single element simulation, can be seen in the measured results. These enable a wideband impedance match. An $|S_{11}| < -10\text{ dB}$ impedance matching bandwidth of about 19% (67.9–82.5 GHz) is obtained.

In terms of realized gain, accounting for the losses due to the waveguide transition (0.4 dB) and those of the microstrip feed line to the antenna (1.5 dB), a maximum measured realized gain of about 15.4 dBi at 72 GHz in the usable frequency region of 71–76 GHz is measured. The realized gain remains within 3 dB of the maximum throughout the measured frequency range of 71–80 GHz.

The measured and simulated farfield radiation patterns are shown in Fig. 9. As explained previously, the operating frequency of the designed antenna is appr. 5 GHz lower than the designed frequency due to dielectric constant modelling errors. Hence, results are shown for 71, 73 and 76 GHz. E-Plane and H-Plane co-polarized simulated and measured results are shown. A broadside radiating main lobe is observed in the H-Plane, and a 3 dB beamwidth of 11° is measured. Good side-lobe suppression is observed at 71 and 73 GHz, with a maximum side-lobe level of -12 dB at 73 GHz. The side-lobe reduces to -7 dB at 76 GHz. Lower side-lobes are seen in the simulated results, and the discrepancy is expected to be due to slight tolerances in manufacturing, which can have a large influence on side-lobe suppression. In the E-Plane, there is a squint in the negative direction of about 10° at 71 GHz, which reduces to 3° at 76 GHz. A 3 dB beamwidth of 44° is measured at 71 GHz, which reduces to 36° at 76 GHz. The beam squint and change in E-Plane pattern is due to the asymmetric ground plane below the thin film patch element, and is also seen in the simulated results.

V. CONCLUSION

A novel multilayer thin film antenna pasted on a single layer PCB was presented. A wideband antenna performance is achieved through the use of an aperture coupled stacked patch antenna, which is completely etched on the multilayer thin film. A single layer PCB is used with microstrip power distribution terminated in a GCPW fed excitation slot. To enable close placement of antenna elements for broadside radiation, a compact wideband bi-phase power divider was implemented. Measurement results of a 2×8 array show a shift of 5 GHz to lower frequencies due to a discrepancy in modelled dielectric constant values. Nonetheless, a wideband impedance match of about 19% (67.9–82.5 GHz) with a broadside radiation pattern and a maximum realized gain of 15.4 dBi at 72 GHz in the usable frequency range of 71–76 GHz is measured. A wide beam in the E-Plane as well a narrow beam in the H-Plane is achieved. This antenna can be used in mid-range automotive radar applications.

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