Fully micromachined W-band rectangular waveguide to grounded coplanar waveguide transition


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Abstract: This study presents a novel grounded coplanar waveguide (CPW) to fundamental rectangular waveguide mode transition for W-band applications. The designed waveguide-to-CPW transition is optimised to achieve low-loss. The maximum insertion loss of the waveguide-to-CPW back-to-back transition is $-1.8$ dB in a frequency range of 91.2–113.2 GHz, corresponding to a large 22% bandwidth. Since the structure benefits from a simple topology and in order to demonstrate the claims, it is fabricated in-house. The measurement results are presented and are in very good agreement with the simulations.

1 Introduction

New frequency resources are being exploited because of the recent rapid developments in various wireless communication service areas and the resulting exhaustion of the microwave frequency band for commercial use. For this reason, many researchers have made great efforts on highly integrated millimetre-wave monolithic integrated circuits (MIMIC) for the purpose of W-band wireless communication systems. A significant amount of research has been performed on coplanar waveguides (CPW), as one of the key MIMIC circuit features, owing to several advantages over other competing transmission lines [1, 2]. The grounds and the signal are on the same side of the substrate, eliminating the need for via holes or a wraparound process [3, 4]. Also, the characteristic impedance is determined by the ratio of the width of the centre strip conductor to the distance between the two semi-infinite ground planes [5], enabling compact design.

On the other hand, rectangular waveguides are widely used at higher frequencies for their low-loss characteristics in applications such as high-Q filters, resonators and antenna feed networks. Therefore for millimetre- and submillimetre-wave applications in which active and passive components are integrated, waveguide structures are often to be combined with CPW or microstrip lines [6, 7]. For those applications, low-loss transitions are required, as shown in Fig. 1.

There are several types of transitions from rectangular waveguide to printed lines such as microstrip, CPW and grounded CPW (GCPW) [8]. However, the concept of converting the transverse electric (TE) or transverse magnetic (TM) waves of a rectangular waveguide to the quasi-TEM modes of a planar line is the same, as will be illustrated in Section 2. Probe feeding [6] from planar line to waveguide demonstrates one of the best performances compared to other types, but it suffers from manufacturing complexity, especially in higher bands owing to the fact that we need to fabricate a very tiny and sensitive extra mechanical part such as a millimetre waveguide circuit [6, 9] and/or a metal post [10]. There is another type of transition, namely the slot-coupled transition that can be fabricated through a micromachining technique, which simplifies the fabrication and the assembly.

For an optimal design of a transition, the requirements are a minimum return loss at the input of the waveguide and a minimum radiation from the coupling slots. This paper presents a novel slot-coupled wideband transition structure. We have started with a simple structure that is the end-wall connection between a metal waveguide and a CPW line with a single slot in the ground. Several waveguide transitions with one slot in the ground have already been presented [11, 12], but they suffer from a narrow bandwidth as a result of the high-impedance difference between the waveguide and the strip, proportional to the difference between the dielectric constant of the substrate and the material in the waveguide [13].

In this work, we propose a very simple and easy-to-fabricate low-loss transition from a grounded coplanar waveguide (GCPW) to a rectangular waveguide operational in the 91.2–113.2 GHz frequency band, corresponding to a large 22% bandwidth. It consists of two transitions, one from the rectangular waveguide to the slots in the ground of the CPW, and the other one from these slots to the CPW itself. It is shown in the paper that the key elements, which enable the wide bandwidth performance are the use of a double slot, the via placement and the matching circuit on the printed side, which are elements that are normally not used in such a configuration.

In Section 2, a brief overview of important end-wall slot-coupled transition structures is presented. Section 3 discusses the proposed double-slot waveguide-to-GCPW transition design and presents the primary simulation results. Since the primary structure had many vias to block the parallel plate leakage and these vias are difficult to implement, Section 4
2 End-wall transitions in the literature

As mentioned earlier, a minority of the CPW or GCPW transitions use slot coupling [14]. In this section, different transitions from the literature are reviewed in detail, explaining how to achieve state-of-the-art wideband behaviour.

We start with a simple structure that is the end-wall connection between a metal waveguide and a microstrip line with a single slot in the ground as depicted in Fig. 2a. This structure uses magnetic coupling as the magnetic field is more involved in the coupling. There are other types of the transition which use electrical coupling, such as some probe feeding structures [10]. A slot-fed microstrip antenna has been used to couple a cavity resonator to a microstrip line by Scheck [15]. Using this idea, many microstrip-to-waveguide transitions have been presented. This kind of transitions, with one slot in the ground, are narrowband as a result of the high-impedance difference between the waveguide and the microstrip, proportional to the difference between the dielectric constant of the substrate and the material in the waveguide [13]. Although coupling aperture enlargement increases the coupling level, it decreases the coupling efficiency because of the increase in radiation loss. However, if the signal level received in the microstrip ports is maximised, the radiation loss from the slot will be minimised proportionally as well. Pozar [16] pointed out that the end-wall connection of the guide causes strong reflection which prevents us from matching the waveguide input in Fig. 2a, even for a good coupling. Also, a microstrip-line-fed resonant slot has very high series impedance. To reduce this impedance one can offset the feed line from the centre of the slot or use matching transformers as in [17]. However, about half of the input power will be radiated at the microstrip side of the structure.

[16]. Moreover, offsetting the slot from the centre increases the cross-polarisation if used inside an antenna. A successful narrowband design, similar to Fig. 2a, has been reported at 140 GHz [18]. However, the simulated −10 dB return loss and 1 dB insertion loss bandwidth are about 2%, which is too small for most applications. Having said so, this transition design is highly dependent on the frequency, the relative thickness of the substrate and its dielectric constant.

Villegas and Stones have modified the mentioned structure using a bulky metal cavity enclosure and a broadband matching stub. Then, using a broadband microstrip end at one side, a bandwidth of 10% is achieved. Clearly, the closed bulky structure that cannot be micromachined complicates the fabrication [13]. Using a piece of dielectric in the waveguide part, as shown in Fig. 2b, acting as a matching transformer, yields an improved return loss [13, 16, 19]. This method eliminates the need for a specially formed waveguide junction. It is of course true that the practical implementation is much more difficult and more expensive than that of Fig. 2a. Another similar configuration has been employed by Wang et al. [20] in a multi-layered LTCC. In Grabherr’s work [19], a bandwidth of more than 10% for a −15 dB return loss and a 0.3 dB transition loss has been claimed, whereas Hyvonen reported a 16% for 20 dB bandwidth. Wang achieved 8.5% for $S_{11}$ less than −15 dB for a single-ended transition.

The highest-bandwidth design of an end-wall transition has been proposed by Davidovitz [12], a tapered transition as depicted in Fig. 2c. The key idea in this transition is the tapered section of the waveguide feeding the aperture in the microstrip ground. This taper matches the high impedance of the standard guide and the relatively low impedance of the reduced-height waveguide. Thus, the problem of the high series impedance in the microstrip is solved and the operation in the whole waveguide band is improved. In splitter/combiner configurations, each microstrip port receives −3.7 dB of power in a full-wave simulation that we performed for this structure. Disadvantages of the structure are that the transition needs expensive high-precision fabrication and holds large dimensions. To be more specific, the transition of [12] has a 15.2 cm tapering length in the X-band. Including additional transition pieces, the total length is more than 20 cm, and which is more than six wavelengths. Simulation shows that the performance would also be acceptable for smaller tapering lengths.

3 Proposed double-slot waveguide-to-GCPW transition design

To design a highly efficient low-loss and wide-bandwidth transition between a rectangular waveguide and a printed
line, the coupling between both ports should be maximised. This automatically improves the return loss at both sides and at the same time minimises the radiation and substrate and metallic losses.

Fig. 3  Proposed structure for the GCPW to waveguide transition

a Top view  
b Bottom view  
c Cross-section cut along the Y–Z plane

The structure shown in Fig. 3 is a new double-slot transition from rectangular hollow waveguide to a grounded CPW. Double-slot coupling highly increases the performance especially with respect to bandwidth [8] and eliminates the
need for matching inside the metallic waveguide part [13]. In this work, the slots are also used for improving the coupling to the CPW line. In [8], the design of the microstrip line to waveguide transition includes two matching stubs on top of the slots, which also block the radiation from the resonant slots. In our case this is not sufficient for the CPW, whereas the problems of radiation and parallel plate leakage are even more severe. Simulations show that the newly presented transition configuration shown in Fig. 3 can increase the performance to a satisfactory level [21]. The structure above the ground looks like a GCPW line with a square cut on top of the coupling part. Two matching stubs, which are shortened to each other at their ends, are located on top of the slots. Then, to reduce the effect of leakage, one set of metalised vias shorts the bottom ground to the GCPW’s top ground. The waveguide aimed for the targeted frequency band is a rectangular WR 8 (2.032 mm × 1.016 mm) waveguide operating from 90 to 140 GHz.

The final design using the above structure has been optimised with CST. It secures a coupling better than −1.5 dB over a bandwidth of 8.6 GHz (i.e. 9% of the band) and better than −1.8 dB over 22 GHz (i.e. 22% of the band), as shown in Fig. 4a. This has been designed using the commercially available Rogers 5880 substrate. The return loss is better than −10 dB between 89.9 and 118 GHz for the waveguide port and the GCPW port, respectively, as shown in Figs. 4b and c. The rest of the power that is not coupled to the GCPW is either reflected, radiated or coupled to other modes of the PCB board. The even-mode coupling level, as shown in Fig. 5a, is ignorable, but the coupling to the parallel-plate mode is quite significant (−12 dB), see Fig. 5b. CST 2009 simulations are validated using Ansoft HFSS 11. There is a slight difference between the results of the softwares. As thoroughly discussed in [22], this can be expected. One of the reasons is the use of different analysis methods (finite integral technique in CST against finite elements in HFSS). The important dimensions of the transition structure are shown in Table 1. The value for the slot length is 1.61 mm which is half of the wavelength at the lower band frequency. Also, the slots are positioned at the edge of the waveguide, 0.88 mm away from each other.

The −1.2 dB coupling is equal to approximately 75% of power. Subtracting the reflection, absorption and coupling to other modes, the remaining 7 to 8% will be radiated.

This power can be radiated from the GCPW slot or from the two slots in the centre of the structure.

Another important issue of concern is the sensitivity to any waveguide-PCB misalignment. Fig. 6 shows the sensitivity of the coupling to miss-alignment in the X direction. It shows that even a 60 µm shift can cause 1 dB loss in coupling.

During the fabrication process, it became evident that the fabrication of 50 µm vias causes many problems. Not only

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Name</th>
<th>Value, mm</th>
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<tr>
<td>ground slot’s length</td>
<td>GSL</td>
<td>1.61</td>
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<tr>
<td>ground slot’s width</td>
<td>GSW</td>
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<td>ground slot X position</td>
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<tr>
<td>line width</td>
<td>LW</td>
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<td>CPW gap</td>
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<tr>
<td>stub’s length</td>
<td>SL</td>
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</tr>
<tr>
<td>stubs’s width</td>
<td>SW</td>
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<tr>
<td>stub’s X position</td>
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<td>width of stub short line</td>
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<tr>
<td>gap size between ground and transition (in the x-direction)</td>
<td>GX</td>
<td>0.1</td>
</tr>
<tr>
<td>gap size between ground and transition (in the y-direction)</td>
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</tr>
<tr>
<td>distance of the short from the transition</td>
<td>DS</td>
<td>0.77</td>
</tr>
</tbody>
</table>

Fig. 6 Sensitivity of the coupling to waveguide-PCB misalignment (simulated by CST microwave studio)

Fig. 5 Coupling

a From the waveguide port to the GCPW’s even mode
b To the parallel-plate mode (simulated by CST microwave studio)
it is a time-consuming process, the pressure required for drilling each hole on the flexible substrate ruins the flatness of the substrate and consequently creates several difficulties in the rest of the fabrication process. So the new goal is to see if the vias can be removed or if their number and placement can be optimised.

Fig. 7 shows that eliminating the vias will reduce the coupling with about 2 dB on average and that the bandwidth will also be decreased. Fig. 8 shows the distribution of the field inside the substrate with and without the presence of vias. As can be expected, the mode conversion from waveguide to microstrip line excites the parallel plate mode in the substrate. This problem is more severe especially when we are using a GCPW line on the substrate. On top of this, the radiation is increased noticeably in the absence of vias, which we did not primarily expect. The increase can be explained by the fact that it shorts the upper and lower ground of the GCPW. Therefore in order not to degrade the performance but at the same time reduce the difficulties of the via fabrication, the next option is to optimise their number, their size and their placement.

4 Via optimisation

Using wider vias basically improves the performance because it provides a shorter via-to-via gap and therefore allows less leakage. It also allows us to use less vias, which is cheaper and easier to implement. To do that, the E-field distribution on the GCPW has been verified to see which vias are more critical in the surface distribution. The result is illustrated in Fig. 9. From this figure, it is obvious that there is a loss of power at the shorted line and one via should be placed there. Moreover, the importance of the

Fig. 7 Coupling level between ports when vias are removed using two different softwares

Fig. 8 Radiation from the structure

a Without vias
b With vias (simulated by CST microwave studio)
vias around the left and right parts of the GCPW line is higher than the vias around the matching structure on top of the slots as the amplitude of the E-field at the via points is much higher.

Thus, the conclusion is that we can reduce the number of 100 μm vias in all three parts of the circuit. The number of vias in the probe side of the GCPW line is decreased from 14 to 8. The more important effect which is observed is that all 26 vias around the centre part can be removed as they do not contribute to exciting the undesired modes. On the shorted part of the line, we end up with a more efficient arrangement of 5 vias instead of 8. The one which is placed at the end is effectively contributing to making a real GCPW short although before the lower ground was not shorted directly to the upper part. Moreover, the diameter of the vias is increased to 200 μm for easier fabrication and better grounding performance.

The end part of the CPW port is tapered smoothly to meet the size necessary for the GSG 200 μm probes. Simulations show that this is not affecting the results, but it increases the sensitivity of the structure to the fabrication process because of the extremely thin gap between the line and the grounds which is 20 μm. Since the thickness of the copper and the width of the gap are in the same order, the effect of the chamfering caused by under-etching in practice has to be modelled carefully. This has been added in simulations to ensure that the 16 μm chamfering does not degrade the results as illustrated in Fig. 10a at the end part of the GCPW port. The final simulated insertion loss and return loss of the transition after the via optimisation are shown in Figs. 11a and b, respectively. The insertion loss is even slightly improved as a result of the via placed in the short-circuit part and the removed vias do not degrade the performance. The return loss only degrades 1 dB, which is not significant.
The above structure has been fabricated in house using a Rogers 5880 substrate. Out of the design two different files are generated: one for the drilling of the different-sized holes (dxf file) and one file for generating the masks (GDSII file) for etching. The holes are drilled by means of a CNC-controlled tool. The challenge here is to generate holes with very little or no burr. The drilling has been done at the highest possible speed, 30 K rpm. After the drilling and cleaning step a photo sensitive chemical layer is sprayed on both sides. The masks are generated with the laser pattern generator LW405 from Microtech. Both masks are aligned to the drilled holes with a MJB55 aligner from Karl Suss. After development the Cu layer is etched by means of a spray etcher. The last step is the filling of the vias by means of CW2000 Nickel-filled conductive polymer epoxy as shown in Fig. 10b.

To measure the performance of the structure, a back-to-back measurement set-up has been built using a pair of WR8 bends and an aluminium fixture structure shown in Fig. 10c connected to probes. Each of the two manufactured transitions is mounted on top of each WR8 bend which themselves are connected from the other side, as illustrated in Fig. 10d. A pair of Ground-Signal-Ground (GSG) probes is connected to both GCPW lines. A 8361-A Agilent PNA (Programmable Network Analyser) has been used to measure the S-parameters. The line-line-reflect-match (LLRM) calibration method is used in this network analyser. The calibration was done at the tips of the probes using a standard W band calibration substrate. The final measured results after the extraction of the return loss and insertion loss of the probes are shown in Figs. 12a and b, respectively. The figures show the results up to 110 GHz because of the frequency limit of the PNA.

The results show a good correspondence to the simulations around 100 GHz, and they are acceptable for the targeted application. However, the insertion loss has suffered more than 1.5 dB of loss. There are two main sources of errors, namely manufacturing and assembly errors, which are responsible for the difference. The manufacturing error is visible in the Fig. 10b. As can be seen, because of the small gap sizes, the copper layer has not been removed completely in many parts. These remaining parts can cause some unwanted resonances owing to the parasitic capacitors they introduce. The drop in the insertion loss around 103 GHz is probably because of this problem. However, the main reason of the difference in insertion loss is owing to the assembly error as can be seen in Fig. 6 for a very small misalignment. As indicated in Fig. 6, a tiny shift of 60 μm in the direction of the CPW port causes up to 1 dB extra loss.

![Fig. 12](https://example.com/fig12.png)

**Fig. 12** Final measured results

- a Measured $S_{11}$ of the fabricated transition
- b Measured $S_{12}$ of the fabricated transition
5 Conclusion

A new transition from CPW to rectangular hollow waveguide for W-band applications has been presented. The transition has been designed to operate in the frequency range of 91–113 GHz. The return loss and radiation loss have been minimised to reduce the coupling insertion loss of the transition within the required band. The maximum measured insertion loss of the waveguide-to-CPW back-to-back transition is −1.8 dB in the frequency range of 91.2–113.2 GHz, corresponding to a 22% bandwidth, which is a huge improvement compared to previously reported works.

6 Acknowledgments

This project was funded in part by the GOA and Center of Excellence projects of K.U. Leuven. The authors thank Michel De Cooman for fabricating the transition.

7 References

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