A Wideband Differential Microstrip-to-Waveguide Transition for Multilayer PCBs at 120 GHz

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Abstract—A robust and wideband differential microstrip line-to-WR6-waveguide transition for mixed multilayer PCBs at D-band is presented. The PCB stack is composed of three standard RF core materials with standard thicknesses. Compared to other inline transitions above 90 GHz, which are designed for single substrates, it can be applied to mixed multilayer circuits. The tapered transition is only 8.0 mm wide and 2.5 mm long, has an insertion loss between 1.2 dB and 1.8 dB, and a return loss higher than 10 dB for 30 GHz bandwidth around the center frequency of 120 GHz.

Index Terms—D-Band, waveguide, transition, differential microstrip lines, millimeter-wave (mmW)

I. INTRODUCTION

WAVEGUIDE transitions are a versatile tool for all RF areas ranging from scattering parameter measurements of passive devices to the characterization of active components. Regardless that many RF frontends are implemented on chip nowadays, the radiating elements are still often realized on PCBs for better radiating performance [1]. In addition, PCB-to-waveguide transitions are robust, easy to assemble, and enable modular designs. Differential microstrip lines (DMSL) are used, because in comparison to single-ended microstrip lines, they are more robust to discontinuities on the ground plane and can suppress external interferences.

A vertical DMSL-to-waveguide transition for 60 GHz is presented in [2]. It uses a triangular patch with a quarter-wave cavity resonator exciting the TE_{10} mode into the E-plane of a WR15 waveguide. Within 33% relative bandwidth for a return loss (RL) higher than 15 dB, an insertion loss (IL) between 1.6 dB and 2.3 dB is achieved. In [3] another vertical transition for a center frequency of 96 GHz is published. A differential microstrip patch together with a shortened parasitic patch inside a WR10 waveguide is used to feed the fundamental mode into the H-plane of the waveguide. The advantage is that no backside quarter-wave resonator is needed, but feeding the H-plane leads to a less broadband design. With a minimum RL of 10 dB, a relative bandwidth of 15 % with a maximum IL of 0.5 dB is realized.

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Inline transitions with center frequencies of 70.5 GHz and 92.5 GHz are reported in [4] and [5], respectively. The unilateral finline design in [4] covers 52% relative bandwidth with respect to 15 dB RL with an IL between 0.3 dB and 1.0 dB. It uses Rogers RO3003 substrate with a thickness of 127 μm and is 14 mm long. Tapered transitions on an ultra-thin 25 μm RO3850 Ultralam substrate are used in [5]. For measurements the transition with a length of 2.45 mm is completely inserted in a WR10 waveguide and achieves an IL lower than 0.9 dB over the full W-band.

In this work, a compact transition for mixed multilayer PCBs with a center frequency of 120 GHz covering at least 20 GHz bandwidth is presented. Because PCB manufacturing tolerances are a severe problem for RF designs above 100 GHz, patch antenna-based approaches are unfavorable due to expected frequency shifts. Hence, a tapered transition comparable to [5] is designed but using mixed multilayer PCBs with a height of 1 mm consisting only of standard RF substrates with standard thicknesses. The multilayer PCB with the used core materials has the advantage that the transition can be used in complex circuits with the need of multiple routing layers, it can be produced with standard PCB manufacturing processes, and it offers mechanical stability. As a consequence of the relative thick PCB, special attention has to be paid to not excite higher order modes.

II. DESIGN AND FABRICATION

The designed inline transition from DMSL to waveguide is built for a mixed multilayer PCB consisting of three

Fig. 1. Multilayer PCB, showing the used substrate materials, bonding films, and the available metal layers L1 to L6. The intersection between the PCB and the WR6 waveguide is also highlighted.

Fig. 2. Top view on metal layers, (a) L1: RF signal layer with taper structure, (b) L2: RF ground layer with matching structure, and (c) the shape of the remaining layers L3 to L6 plus the fully metallized FR4 carrier. All layers are symmetric and can be mirrored at the symmetry line (---).

substrates with thicknesses of 127 μm , 254 μm , and 500 μm as depicted in Fig. 1. All metal layers are 17 μm copper foil. In the following, the single metal layers and their function are described.

The metal layers L1 and L2 with RO3003 core material are used for 122 GHz signal routing, and the most important part of the transition is designed on these layers. RO5880 core material with a height of 254 μm and lower ϵ_r compared to RO3003 can be used to implement radiating elements on the metal layers L3 and L4. These layers are also appropriate for lower frequency signal routing. An edge metalized FR4 substrate surrounded by the metal layers L5 and L6 acts as a carrier in this work. The edge metalization is only important at the interconnection to the waveguide to guarantee a good electric contact in order to not disturb the surface currents of the waveguide. These layers could also be used for digital signals or power planes in future designs. For radiating elements on L4, the carrier could be milled out to realize a cavity resonator.

The parameters of the realized structures of the different metal layers L1 to L6 are depicted in Fig. 2, and the final parameter values are summarized in Table I. All layers are symmetric and can be mirrored at the blue dash-dotted line.

A. Transition Design

The differential mode of a DMSL (segment I) shall be transformed in the overlapping area between the PCB and the waveguide (segment II) to the TE_{10} mode of a WR6 rectangular waveguide (segment III). For this reason, the waveguide is centered along its width a_{wg} to half the height of the RO3003 substrate as depicted in Fig. 1.

On the top metal layer L1 (see Fig. 2(a)) the DMSL is parameterized by the width w_{DMSL} of a single transmission line and the gap g between the two lines for a differential impedance of 100 Ω . To transform the differential DMSL mode to the waveguide mode, g is exponentially widened to the waveguide height b_{wg} over the taper length t_{L1} .

Matching is achieved on the RF ground layer L2 as shown in Fig. 2(b). The ground plane is exponentially tapered to zero over t_{L2} . In addition, an inset i_{L2} and an offset o_{L2} with respect to the waveguide walls are introduced for matching.

In order not to alter the fundamental waveguide mode the remaining metal layers L3 and L4 as well as the complete edge metalized FR4 carrier (with L5 and L6) are etched and milled

TABLE I
DESIGN PARAMETER VALUES OF THE OPTIMIZED TRANSITION.

Parameter Value [mm]	g	w_{DMSL}	b_{DMSL}	d_{via}	p_{via}
	0.100	0.200	2.350	0.300	0.450
Parameter Value [mm]	b_{via}	r_{tool}	a_{wg}	b_{wg}	z_{wg}
	1.325	0.500	1.651	0.825	1.357
Parameter Value [mm]	t_{L1}	t_{L2}	i_{L2}	o_{L2}	
	1.720	1.470	0.137	0.200

Fig. 3. Pictures of the fabricated transition, (a) top metal layer L1, (b) milled out and edge-metalized FR4 carrier, and (c) top view with attached waveguide.

out in the region where the waveguide is overlapping with the PCB (segment II). The region is specified by the overlapping length z_{wg} and the height of the waveguide b_{wg} as depicted in Fig. 2(c). The size of the corners is restricted by the smallest milling tool available with tool radius r_{tool} .

The top and bottom view as well as the assembled fabricated transition are depicted in Fig. 3. The shape of the taper on L1 and the single core materials of the multilayer PCB can clearly be seen in Fig. 3(a). On the bottom side, Fig. 3(b) shows the milled out carrier and the adjusting pins for proper alignment of the PCB to the waveguide. In Fig. 3(c) the influence of fabrication tolerances for the holes are visible leading to a slight misalignment of the PCB in the waveguide. The transition including the holes for the adjusting pins is 8.0 mm wide and 2.5 mm long.

III. MEASUREMENT RESULTS

In order to characterize the transition, two back-to-back (B2B) structures with different line lengths l_1 and l_2 with 16 mm and 16.632 mm, respectively, are fabricated. Measurements are performed with a vector network analyzer and

Fig. 4. Measurement results of the de-embedded transition. (a) depicts the transmission coefficient in comparison to the simulated ideal transition and with air gaps of 20 μm and 60 μm between the PCB and the waveguide along the z -axis (cf. Fig. 2). (b) shows the reflection coefficient in comparison to the simulated ideal transition.

WR8 frequency converters in combination with WR8-to-WR6 waveguide tapers. The evaluated frequency range is 100 GHz to 140 GHz. A de-embedding to the scattering parameters of a single transition S_{xx} from the measured scattering parameters of the B2B structures $S_{B2B1/2}$ is performed by applying Mason's rule to the resulting signal flow graph [6]. The derived equations are unambiguously solvable if the line length difference is smaller than half of the guided wavelength of the highest frequency of interest, which is valid up to about 145 GHz in this case.

For de-embedding of a single transition, the effective relative dielectric constant $\varepsilon_{r\text{eff}}$ as well as the losses of the differential mode are needed. $\varepsilon_{r\text{eff}} = 2.255$ can directly be extracted based on the known line difference between $S_{B2B1/2}$ and $S_{B2B2/2}$. In order to evaluate the losses, the dielectric loss tangent $\tan \delta$ of the RO3003 substrate is measured with a time-domain spectroscopy system at 150 GHz and amounts to 0.005. The actual losses used for de-embedding are then derived from a 3D electromagnetic simulation software using $\tan \delta$ and result in 0.13 $\frac{\text{dB}}{\text{mm}}$. This measurement technique does not include all losses, e.g. the rough surface between the copper cladding and the substrate, and therefore serves as a conservative lower boundary for the loss estimation.

Measurement results for the de-embedded transmission coefficient S_{21} are depicted in Fig. 4(a). In the frequency range from 105 GHz to 135 GHz $|S_{21}|$ is between -1.2 dB and -1.8 dB. There is a discrepancy between simulations and measurements of around 1 dB, which is due to the conservative estimation of the losses and a small air gap (50 μm) between the PCB and the waveguide. The air gap is caused by misalignments of the drill holes for the adjustment pins as a result of fabrication tolerances and interrupts the current flow from the waveguide walls to the PCB. Simulations with introduced air gaps of 20 μm and 60 μm along the z -axis (cf. Fig. 2) show the impact on S_{21} .

Results for the reflection coefficients are depicted in Fig. 4(b). $|S_{11}|$ denotes the connection to the waveguide and $|S_{22}|$ the connection to the DMSL. Besides a slight frequency shift of around 5 GHz, the overall trend of the simulations

TABLE II
COMPARISON OF DIFFERENT DMSL-TO-WG TRANSITIONS. FOR DIFFERENT REALIZATIONS, THE CENTER FREQUENCY f_c , THE ABSOLUTE AND RELATIVE BANDWIDTH $BW_{\text{abs/rel}}$, AND THE MAXIMUM INSERTION LOSS IL_{max} AND THE MINIMUM RETURN LOSS RL_{min} ARE LISTED.

	[2]	[3]	[7]	[4]	[5]	this work
Substrate	Mixed	TLE-95	RO4350B	RO3003	RO3850	Mixed
Height [μm]	1450	127	101	127	25	1000
f_c [GHz]	60	96	92	70.5	92.5	120
Band	V	W	W	V/E	W	D
BW_{rel} [%]	33	15	38	52	38	25
BW_{abs} [GHz]	20	14.5	35	37	35	30
IL_{max} [dB]	2.3	0.5	1.1	1	0.9	1.8
RL_{min} [dB]	15	10	13	15	12	10
Length [mm]	3.77*	3.11*	2.54*	14.00	2.45	2.50
Alignment	vertical	vertical	vertical	inline	inline	inline

*neglecting 19 mm outer diameter of WR15/10 round waveguide flange

and measurements are in good agreement. An RL better than 10 dB is achieved in the frequency range from 105 GHz to 135 GHz.

Table II gives an overview of the presented transition compared to state-of-the-art DMSL-to-waveguide transitions. Even though a relatively thick mixed multilayer PCB of 1 mm is used at D-band, a compact and robust transition with comparable performance is achieved. Despite the high center frequency 25 % relative bandwidth is covered with only a slight increase of the IL .

IV. CONCLUSION

An exponentially tapered transition from DMSL to waveguide for a three core mixed multilayer PCBs using industry standard RF materials and thicknesses at 120 GHz has been designed. Design challenges arising from the relatively thick substrate stack of 1 mm are solved by a combination of a milled carrier and edge metalization in the overlapping area of the PCB and waveguide. Over a bandwidth of 30 GHz an IL between 1.2 dB and 1.8 dB with an RL higher than 10 dB has been achieved.

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