Numerical Investigation of Vertical Contactless Transitions for Multilayer RF Circuits

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Abstract: **This paper presents some concepts for contactless transitions between different planar transmission lines. These types of transitions are important for use in e.g. multi layer designs. Another important reason to avoid galvanic connections at RF is that they may generate *passive intermodulation products* (PIM), which can be a serious problem in multifrequency systems. The investigated transitions operate in the extended GSM 900 band (880-960 MHz).**

I. MICROSTRIP-TO-STRIPLINE TRANSITION

The investigated structure provides a transition between a microstrip and a stripline transmission line. The characteristic impedance of both lines is 50 Ω. The stripline structure consists of three layers of different dielectrics but with the same permittivity: TLC-30, manufactured by Taconic, and Kapton® manufactured by Du-Pont (Fig. 1).

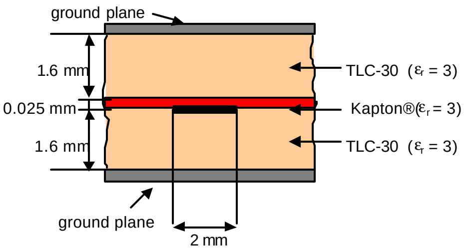

Fig. 1. Stripline dimensions.

The Kapton® layer is centered between the two ground planes and the stripline is etched on the lower TLC-30 layer.
The microstrip is designed as a matching 50 line.
The coupled section of the transition is shown in Fig. 2. The overlapping length of the two conductors has been adjusted numerically at 920 MHz, center frequency of the 900 MHz GSM band. The coupling is maximum when the overlapping is equal to a quarter of the guide wavelength, 46 mm in our case. The stripline conductor ends exactly at the discontinuity air-TLC-30.

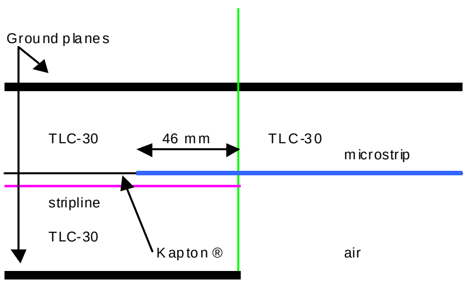

Fig. 2. Coupled section, side view

The microstrip width varies at the discontinuity air-TLC-30 with a step from 4.08 to 2 mm and is equal to 2 mm in the coupled section.
The measured and calculated results are compared in Figs. 3 and 4.

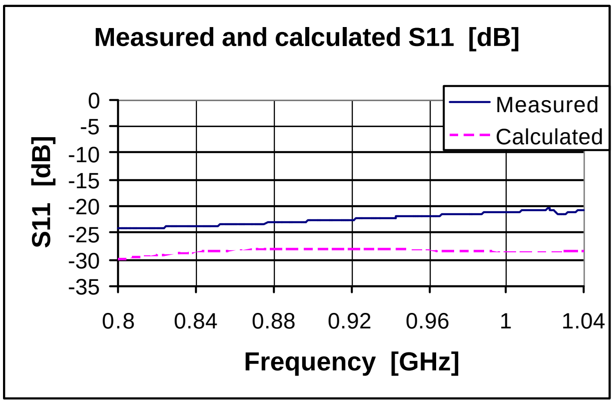

Fig. 3. Measured and calculated S_{11} [dB].

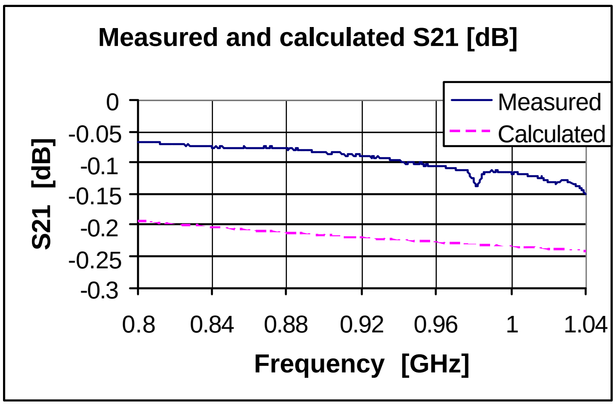

Fig. 4. Measured and calculated S_{21} [dB].

The transition is better in reality than in the simulations, with S_{21} better than –0.15 dB over the whole range of frequency. The radiated power doesn't exceed 3% measured, compared to 5.2% calculated. The transition behavior is broadband.

II. MICROSTRIP-TO-MICROSTRIP TRANSITION THROUGH A RECTANGULAR SLOT

The transition is realized between two parallel microstrip lines on parallel substrates coupled through a rectangular slot in the common ground plane. The designed transition is between two 50Ω microstrips and employs two identical TLC-30 substrates 1.6 mm thick, as shown in Fig. 5.

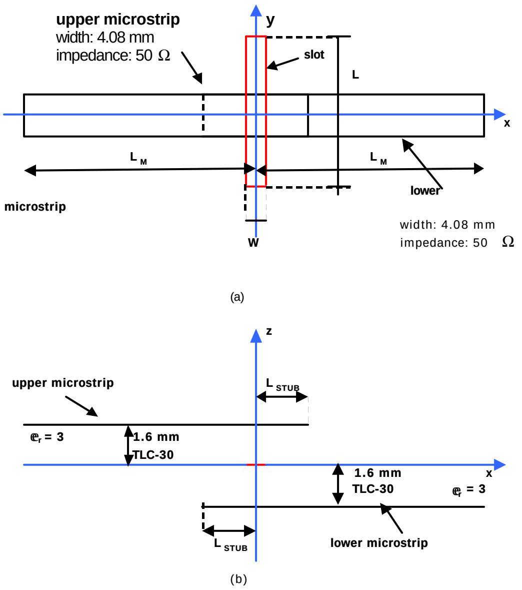

Fig. 5. Transition structure and dimensions; (a) top view. (b) side view.

The design parameters are the length of the tuning stubs and the length and width of the coupling slot. The designed circuit is symmetrical (equal substrates) and with homogeneous dielectrics, therefore the open-circuited stubs have equal length.
If we consider a single microstrip line fed by a radiating slot printed in the ground plane, the following equivalent circuit can be derived [7]:

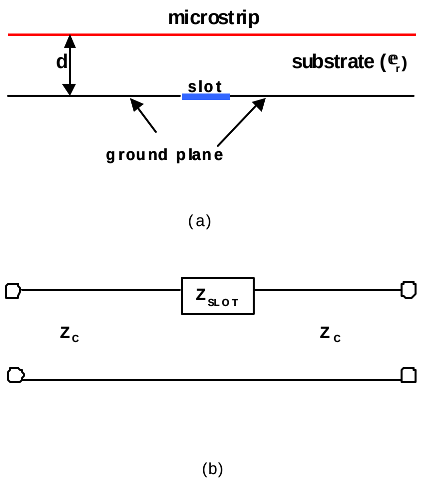

Fig. 6. (a) slot-fed microstrip, side view; (b) equivalent circuit of (a).

The slot discontinuity appears as a series impedance, $Z_{SLOT} = R + jX$, to the microstrip line and its value can be evaluated numerically [7]. The numerical computation involves a proper representation of the fields in the slot by employing appropriate Green functions. The complete analytical analysis is reported in [7]. When $Z_{SLOT} = R$, the slot is said to be resonant. The first step in our design has been to determine the slot's resonance length at 920 MHz, center frequency of the extended GSM 900 band. The length of the open-circuited tuning stubs has been fixed to $\lambda_g/4$, where λ_g is the guide wavelength at 920 MHz. With this length, the open-circuited stubs are equivalent to a short circuit in the plane at x=0, Fig. 5 and provide the excitation to the slot. At 920 MHz the computed propagation constant, equal for both the feed and coupled microstrip, is β=29.883 rad/m, from which the guide wavelength is found to be $\lambda_g = 210.3$ mm $\rightarrow \lambda_g/4 = 52.6$ mm. In this simulation the stubs' length has been set to 53 mm. The simulations have been performed with the slot width W equal to 3 mm. The slot length L has been gradually increased from a starting value of 40 mm, until a minimum in S_{11} [dB] has been encountered. The slot resonance length at 920 MHz has been found to be 130 mm. From a series of simulations it has been noticed that the power radiated by a slot increases rapidly with the slot length. It is therefore essential, in the transition design, to keep the slot as short as possible, in order to reduce the transmission loss. The lower limit to the slot length is set by the coupling, which turns out unsatisfactory if the slot is too short. In our case the coupling was to low with a slot shorter than 50 mm.
The transition has been simulated between 800 MHz and 1.04 GHz with the following slot lengths:

- L = 130 mm (resonant slot at 920 MHz)
- L = 80 mm
- L = 60 mm

The stubs are shorter than a quarter wavelength for the slots below resonance and have been adjusted numerically.
The advantage of employing a slot below resonance is the significant reduction of the radiated power, as shown in Fig. 7.

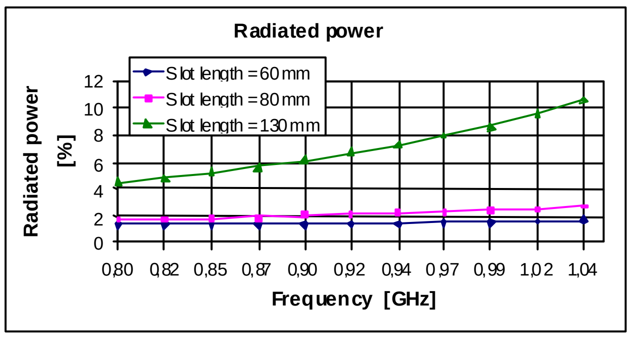

Fig. 7. Comparison of the percentage of radiated power as a function of frequency.

The resonant slot radiates a high amount of power with a maximum value exceeding 10% at 1.04 GHz, compared to the non-resonant slots, for which the transmission loss is less than 3%. In particular for a 60 mm slot the radiated power is less than 2% over the entire range of frequency. For all three slots the amount of radiated power tends to increase with frequency, but this increase is less evident as the slot becomes shorter. As a consequence, the coupling is significantly improved by using the 60 mm slot, which ensures S_{21} [dB] better than –0.064 dB (calculated) over the entire GSM band, compared to –0.102 dB for the 80 mm slot and –0.39 dB for the resonant slot.

A transition has been realized and measured with the following parameters, Fig. 5.

- Slot length $L = 60\ mm$
- Slot width $W = 3\ mm$
- Stubs' length $L_{STUB} = 35\ mm$

The measured and calculated results are shown in Figs. 8, 9 and 10.

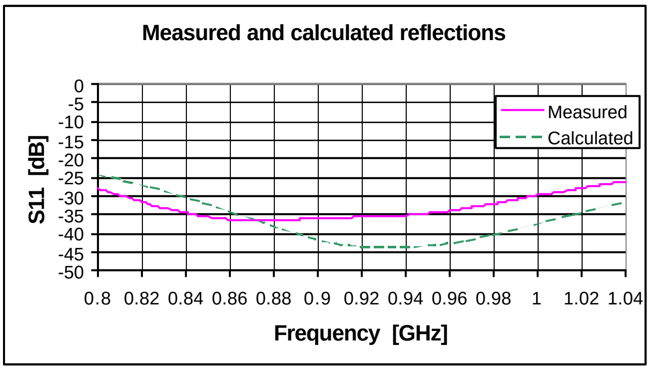

Fig. 8. Measured and calculated S_{11} [dB].

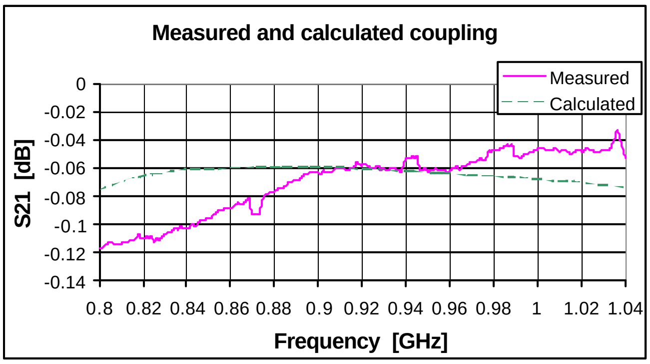

Fig. 9. Measured and calculated S_{21} [dB].

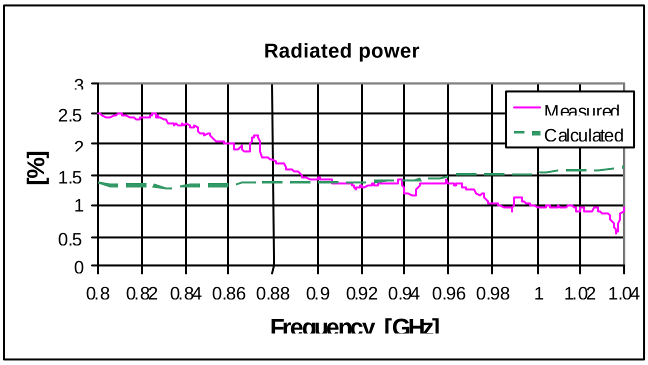

Fig. 10. Percentage of radiated power.

In the realized transition, the matching is better than –25 dB on the whole frequency range analyzed and better than –33 dB on the GSM band, Fig. 8. The coupling is higher than –0.12 dB (measured) compared to the minimum calculated value of –0.08 dB. The maximum radiated power is 2.5 %, compared to 1.6% calculated. For frequencies higher than 900 MHz the radiated power is lower than 1.5%.

III. SUSPENDED-STRIPLINE-TO-MICROSTRIP TRANSITION

The novelty of this transition is to enhance the power transfer between the two transmission lines by inserting a dielectric plug in the vicinity of the coupling slot [5]. The function of the dielectric plug is to concentrate the field into the upper substrate of the suspended stripline and especially in the region surrounding the slot. The resulting field in the suspended stripline will have, therefore, an asymmetrical pattern, with more dense field lines in the upper substrate around the slot than in the bottom layer. As a consequence, significantly more power is coupled through the slot compared to the case in which the dielectric plug is not used. In order to perturb the stripline field pattern, the dielectric plug should have a high dielectric constant compared to the substrates' permittivities.
The coupling can be maximized, for a determined slot length and width, by adjusting the length of the open-circuited tuning stubs. Since the dielectric materials surrounding the slot have different permittivities, the stubs will not have the same length, but they need to be adjusted separately.
The transition is between the following transmission lines:

<u>Suspended stripline</u>: the characteristic impedance is $Z_{CS} = 75\Omega$. The structure consists of a suspended substrate of polyester 75 μm thick, on which the stripline is etched, centered between two Rohacell layers 2 mm thick, Fig. 11.

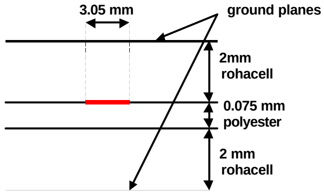

Fig. 11. Suspended stripline structure.

<u>Microstrip</u>. the chosen substrate is a TLC-30 layer 1.6 mm thick. The characteristic impedance is 75Ω. The microstrip structure is shown in Fig. 12.

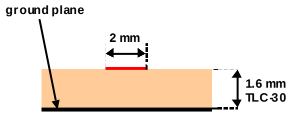

Fig. 12. Dimensions of the microstrip line.

The designed transition structure is shown in Figs. 13 and 14.

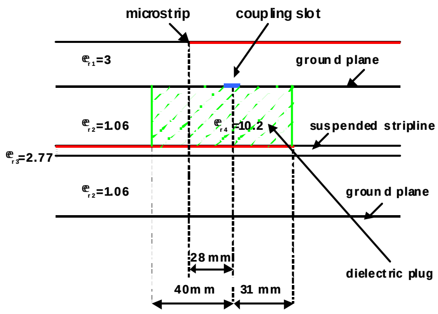

Fig. 13. Transition structure, side view.

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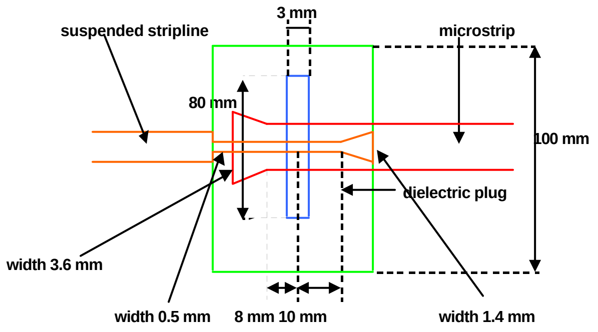

Fig. 14. Transition structure, top view.

The scattering parameters in the range of frequency between 800 MHz and 1.04 GHz are shown in Fig. 16.
The percentage of power lost due to radiation by the slot is shown in Fig. 16.

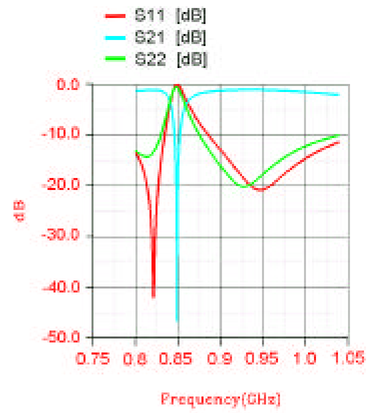

Fig. 15. S_{11} [dB], S_{21} [dB] and S_{22} [dB] in the range of frequency 800 MHz -1.04 GHz. Coupling: -1.48 dB at 880 MHz, -1.13 dB at 960 MHz, max -1.08 dB at 920 MHz.

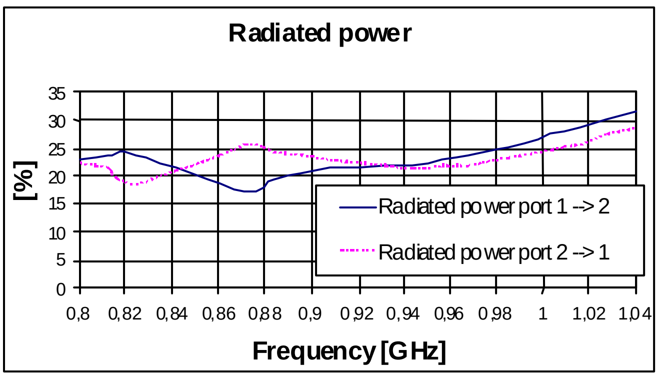

Fig. 16. Radiated power. Port 1: stripline. Port 2: microstrip.

The results above show that the transition is affected by a high loss due to power radiation by the slot. The radiated power exceeds 20% over the 900 MHz GSM band. It has been observed that the slot length could not be reduced further than 70 mm because of the low coupling with shorter slots. The issue of power radiation is also addressed in [5]. According to [5], the development of analytical tools is essential in order to optimize the transition design and minimize the slot radiation.
The transition behavior is also strongly influenced by the shape and size of the dielectric plug. In particular positioning the plug with respect to the coupling aperture has turned out as a critical point of the transition design. It has been noticed, however, that a plug placed asymmetrically with respect to the slot, as in Fig. 13 and 14, allows a higher power transfer between the two transmission lines than a centered plug. The coupling is also sensitive to the plug width. In fact, it has been noticed that a plug wider than the slot seems to enhance the coupling. The main design difficulty is combining the plug position with the adjustment of the tuning stubs in order to maximize the coupling. According to [5] and [7], the power sharing between the slot radiation and the coupled stripline mode can be controlled also by the shape and size of the coupling aperture and by the permittivity of the substrates.
In our design, tuning stubs with gradually-increasing width have been employed. It has been noticed that the coupling is increased by 0.15 dB in the whole range of frequency, compared to the case of straight stubs.
In Fig. 15 the transition behavior seems to be narrowband, with an impedance matching better than –13 dB over the GSM band The coupling curve however, is rather flat over the whole GSM band.

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