Measurement and Simulation of Coaxial to Microstrip Transitions' Radiation Properties and Substrate Influence

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Abstract. *A radiation and electro-magnetic (EM) field analysis of coaxial-to-microstrip transitions is presented. Radiation is quantified by simulation and measurement of a crosstalk between two Omni-Spectra's transitions using microstrip 'open' calibration standards at different positions. Simulation results are compared to the measured data and good agreement is reported on two different substrates. The evaluation method which is used to analyze quality of the transition and its radiation properties was already developed and verified on a grounded coplanar waveguide (CPWG) transmission line. Results can be used to estimate uncertainty budget of the calibrated measurement with respect to the measured radiation. Results on different substrates show interesting behaviour and can prove useful when choosing suitable substrate for specific test-fixture.*

Keywords

Microstrip, measurement standards, microwave measurement, microwave propagation, electromagnetic fields, radiation effects.

1. Introduction

These days it is common to use any commercially available coaxial-to-planar transitions during any stages of the microwave design process. These 'launchers' provide the transition between the transverse electro-magnetic (TEM) mode inside the coaxial line and the quasi-TEM (QTEM) mode on the planar transmission line. General reflective standards (usually 'open' or 'short') are used in most standard [1] or over-determined [2], [3] calibration/correction methods. When performing on-planar calibration one can then evaluate the error terms used to obtain the measured S-matrix of some device under test (DUT). This introduces an important assumption that the error model of the linear vector network analyzer (VNA) must preserve constant electrical parameters when measuring any type of DUT.

It has been shown previously [4], [5] that coaxial-to-microstrip transitions radiate part of the incident wave especially when subjected to large values of voltage standing wave ratio (VSWR). It was observed [4] that the leaked power is proportional to the phase of the reflected wave. This means that one will observe radiation maxima in case that the VSWR anti-node is transformed over the $\lambda_g/2$ transmission line near the transition edge. This would invalidate the basic assumption of constant error model of the VNA. To verify this hypothesis measurements with different distances of reflective calibration standard from the transition can be used.

Based on these observations it is important to evaluate the launcher's radiation under large VSWR condition. Previous attempts to quantify the radiation influence on the on-planar calibration consisted of using a calibration verification element (usually any highly-reflective calibration standard with offset length). However this is not reliable enough, because residual error terms [6] of imperfectly calibrated VNA can cover-up the errors due to radiation. New approach was proposed [7] which relies only on a precise calibration at the coaxial connector and then measuring a residual transmission coefficient S_{21} - a crosstalk component. This has been verified by measurements and simulations on a CPWG transmission line.

In this paper we will utilize the proposed method for direct measurement of the crosstalk component to estimate the radiation properties of Omni-Spectra coaxial-to-microstrip-transitions with different substrates and with different offsets of the 'open' calibration standards.

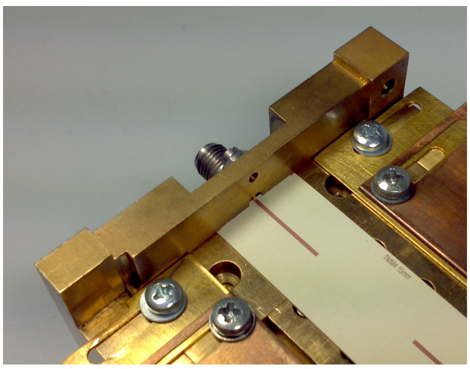


Fig. 1. Test fixture based on two SMA Omni-Spectra launchers with some DUT inside.

2. Coaxial-to-Microstrip Transition

The test-fixture used for the purpose of this paper was made in-house and it is fitted with Omni-Spectra SMA connectors/transitions [8] (see Fig. 1). These end-launchers were originally designed and built when there were no advanced 3D electromagnetic field simulators such as today and in order to understand the radiation phenomenon it is useful to simulate the set-up used during the measurement.

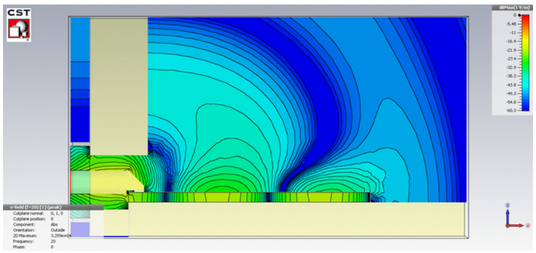


Fig. 2. Simulation results of Omni-Spectra transition with a microstrip 'open' calibration standard. Electric field distribution in shown in the longitudinal section of the model. This should provide the qualitative perspective of the radiation problem.

Based on the simulation results done in CST Microwave Studio and HFSS which are in correlation with the radiation theory [4], [5], [7] it was observed that substrate properties in combination with transmission line properties play a crucial role on the magnitude of radiated wave from the transition.

Fig. 2 shows a basic explanation of the radiation problem. It can be seen that the excited wave propagates along the transmission line and causes interferences with the propagating QTEM field.

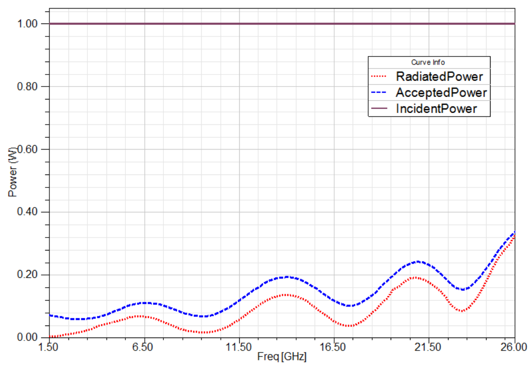


Fig. 3. Simulation results of the transition with a microstrip 'open' calibration standard. The maxima and minima of radiated power are obtained.

Deeper analysis of the transition's radiation is shown in Fig. 3 in terms of simulating the accepted power and radiated power from the structure. These terms are calculated as follows.

$$P_{accepted} = |a|^2 \left(1 - |S_{11}|^2\right), \tag{1}$$

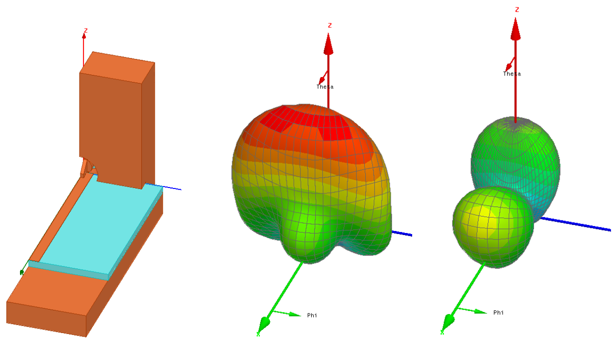
$$P_{radiated} = \Re \iint_S E \times H^* dS. \tag{2}$$

Equation (1) uses term a to define the incident wave with nominal power 1 W used for the simulations. Because the reflection coefficient S_{11} is very high we get very low accepted power. The radiated power is in correlation with the accepted power and Fig. 3 shows that radiation efficiency ranges from 0 to 0.9.

The frequency positions of maxima of radiated power (see Fig. 3) and radiation efficiency corresponds with a theoretical assumption that is based on the transformation of the microstrip 'open' end along the $\lambda_g/2$ (180°) transmission line near the transition edge which then forces the transition to radiate significantly into the free space.

Approximate n-th frequency position of radiated power maximum is a function of effective permittivity ε_{eff} and the length of the microstrip line d. Values can be calculated as

$$f_{rad} \approx n \frac{c}{2 \cdot d \sqrt{\varepsilon_{eff}}}. \tag{3}$$



(a) HFSS model (with symmetry) **(b)** 26 GHz **(c)** 20 GHz

Fig. 4. 3D radiation patterns of simulated transition with a microstrip 'open' calibration standard. Frequencies were chosen to compare between two maxima of radiated power (see Fig. 3). The linear scaling in both patterns is 0.1 V/m to 12 V/m.

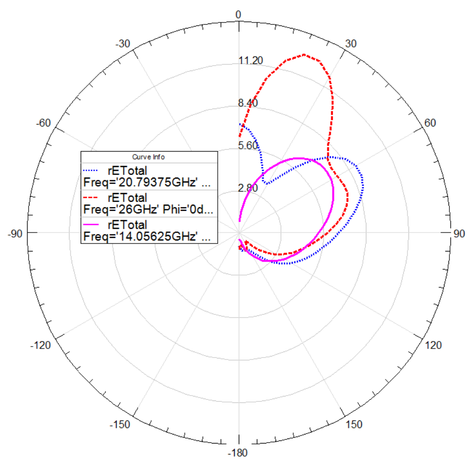


Fig. 5. Radiation patterns as a function of θ for a constant $\phi = 0°$ (X-Z plane) of simulated transition with a microstrip 'open' calibration standard.

The radiation patterns in Fig. 4 were simulated on two frequencies which represents maxima of radiated power. Similar analysis is shown in Fig. 5 for three frequency points: 14.05, 20.8 and 26 GHz. Frequencies were chosen to compare among three maxima of radiated power (see Fig. 3). The patterns show that shape and direction of the main lobe varies with frequency. Hence the radiation pattern properties do not follow any predictable behaviour except the magnitude of the main lobe which increases with frequency.

3. Proposed Approach

The proposed measurement approach is adapted and modified from the previous work [7]. The main advantage over other known methods for measuring the transition's radiation properties is mainly in avoiding the necessity of on-planar calibration. Rather then relying on precise microstrip calibration standards it is based on direct measurement of the crosstalk components, see the signal flow graph in Fig. 6.

By calibrating at the coaxial reference plane which can be done by any commercially available calibration kit we get the reference plane at the coaxial connectors. Coaxial calibration is more reproducible and more accurate than any calibration on the microstrip.

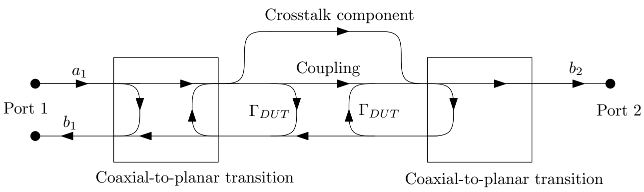


Fig. 6. Signal flow graph of the proposed evaluation method of quantifying the residual crosstalks between the transitions under the condition of large VSWR caused by highly reflective calibration standard ($|\Gamma_{DUT}| \approx 1$).

The measured wave ratio b_2/a_1 consists of the crosstalk between launchers combined with the intrinsic coupling between the reflective standards. These values are influenced by imperfections of the connectors which can be omitted for the first approximation. The crosstalk and the coupling will produce a superimposed wave quantity dependent on the electric length of the DUT.

The coupling between discontinuities can be omitted if the gap is long enough. For this purpose we designed and fabricated test-boards with gap dimensions 20 and 30 mm. The overall length of fabricated test-boards is 50 mm.

4. Measurement and Comparison

To verify the proposed approach a measurement setup was prepared to eliminate all potential systematic errors. Four sets of measurements were done on the Agilent E8364A VNA in the frequency range 1 - 26 GHz with a 100 Hz IF bandwidth. VNA was calibrated using 2.92 mm coaxial calibration kit and Unknown-Thru [9] calibration method implemented in the VNA firmware. The connection repeatability of the calibrated measurement (average squared magnitude error) was better than -80 dB in the whole frequency range (see Fig. 7).

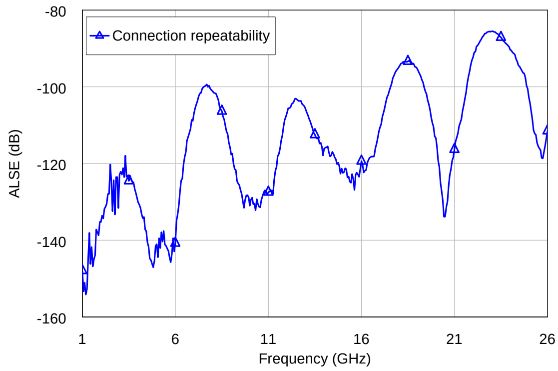


Fig. 7. Connection repeatability of the crosstalk measurement presented using average squared magnitude of the difference between each element of two S-parameter matrices. See (4).

$$ALSE_{dB} = 20\log\frac{\sum_{i=1}^{N}\sum_{j=1}^{N}\left(\left|S_{ijA}-S_{ijB}\right|\right)^2}{N^2}. \qquad (4)$$

The intrinsic coupling between discontinuities cannot be easily measured without influencing the measurement results with another systematic error. Therefore we have separated the discontinuities apart so that the coupling can be neglected in these measurements. DUTs with different lengths of microstrip line (10 and 15 mm) were fabricated to see results for two different phases of the reflected wave propagating along the microstrip. See Fig. 8 for a picture of the TMM4 'open' reflection standard with 15 mm microstrip line. Overall length of the test-boards is 50 mm which gives a distance between open ends of 30, 20 mm respectively.

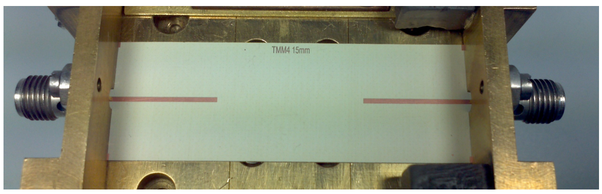
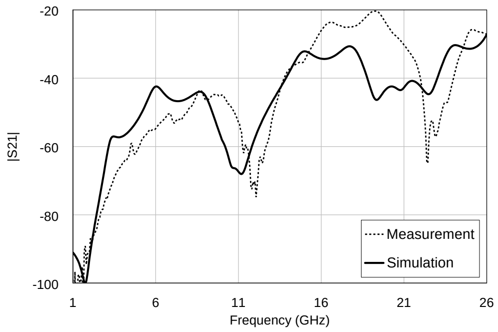


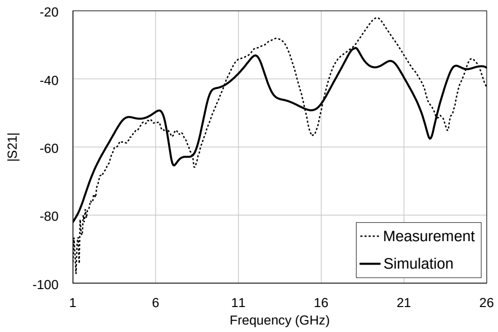
Fig. 8. 15 mm long microstrip lines on a Rogers TMM4 substrate with 'open' calibration standards. DUT is mounted inside the test-fixture.

The measurements were performed on two substrates: 0.508 mm thick Rogers RO3003 with $\varepsilon = 3.0$ [10] and 0.381 mm thick Rogers TMM4 with $\varepsilon = 4.5$ [11]. Two different substrates were chosen in order to check whether it will have some influence on the launchers' radiation properties. The microstrip transmission lines were designed with 50 Ω characteristic impedance on each substrate. Because of fulfilled reciprocity condition ($S_{21} = S_{12}$) only values of S_{21} are shown in the following section.

In the following figures (Fig. 9 - Fig. 10) we can see the measured and simulated data of transmission coefficient $|S_{21}|$ for both substrates with various lengths of the microstrip transmission lines. The measured and simulated transmission proves that for different phases of reflected wave from the open-ended microstrip line we get different radiation properties of the transition under test. This validates the concept introduced in [4].



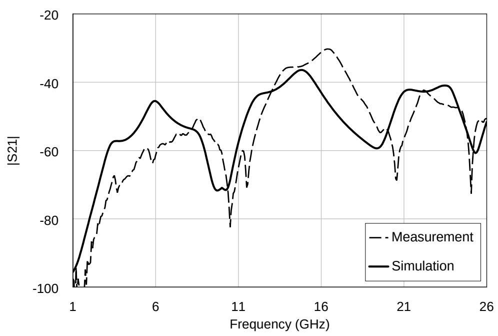
(a) 10 mm long open-ended microstrip lines. Distance between the open ends is 30 mm.



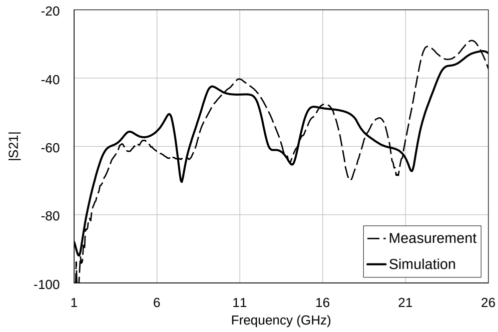
(b) 15 mm long open-ended microstrip lines. Distance between the open ends is 20 mm.

Fig. 9. Comparison of $|S_{21}|$ between simulated and measured data of DUT fabricated on a 0.508 mm thick Rogers RO3003 substrate.

Finally the difference between measurements on both substrates can be seen in Fig. 11. Both measurements are comparable in terms of the frequency characteristics (i.e. peaks in the transmission coefficient). However the substrates have different relative permittivity ε_r and therefore different wavelength λ_g which causes different interferences and superposition of propagating waves.



(a) 10 mm long open-ended microstrip lines. Distance between the open ends is 30 mm.



(b) 15 mm long open-ended microstrip lines. Distance between the open ends is 20 mm.

Fig. 10. Comparison of $|S_{21}|$ between simulated and measured data of DUT fabricated on a 0.381 mm thick Rogers TMM4 substrate.

We can observe that radiated power from the transition increases on thicker substrates with lower ε_r. This should be investigated in order to find a meaningful explanation for this behavior.

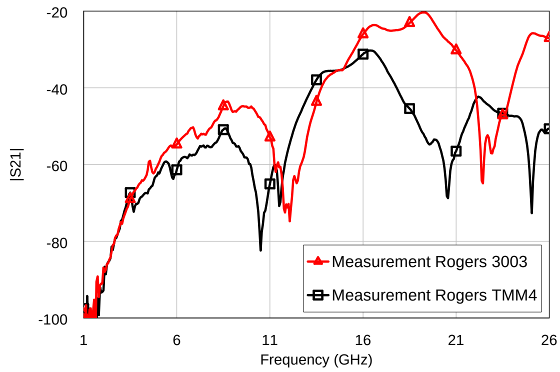


Fig. 11. Comparison of $|S_{21}|$ between measured data of two DUTs with 10 mm long open-ended microstrip lines fabricated on substrates Rogers TMM4 and Rogers RO3003.

It is possible to state that the specific combination of the Omni-Spectra launchers with the substrates used within this article proves that usable frequency range for the real-world applications is very limited in terms of frequency range. This experimental approach can be useful when choosing the most appropriate substrate for measurements in the specific launcher which is currently available.

5. Conclusion

A radiation and electro-magnetic (EM) field analysis of commercially available coaxial-to-microstrip transitions was presented. Radiation was quantified by measuring a crosstalk component between two Omni-Spectra's transitions using 'open' planar calibration standards on different substrates. Simulation results were compared to the measured data and good agreement was reported on both substrates. Results show interesting behaviour especially while comparing radiation properties for different substrates inside the test-fixture. It was observed that radiated power is smaller on thinner substrates and with higher relative permittivity. Deeper theoretical physical explanation has to be done in the future to validate the assumption.

Acknowledgements

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