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Full-Wave Analysis of Discontinuities in Uniplanar and Multiplanar Transmission Lines Using the Frequency-Domain TLM Method

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Abstract --- This paper introduces a rigorous analysis of a variety of transmission line transition discontinuities and module interconnect assemblies in MMIC and miniature MIC circuits using the frequency-domain TLM method. Numerical results of frequency-dependent s-parameters are presented which include the effect of finite thickness and conductivity of the metallization as well as mode interaction between cascaded discontinuities. The effects of the bonding wire for module assemblies is investigated. It is found that the properties of the interconnect are largely depended on the total length of the wire and are quite insensitive to the shape of the wire. This is in a good agreement with experimental observations.

1.Introduction

Typical MMIC and miniature MIC circuits are composed of a variety of different types of planar transmission lines including microstrip, coplanar waveguide(CPW), coplanar stripline and slotlines on single or multiple layers of substrate. Low loss transmission of energy between these different transmission media is a frequently encountered problem in the design of MMIC and miniature MIC circuits. This problem is not easily solved since the structures present a truly 3-D discontinuity with small circuit dimensions. Furthermore, discontinuities are closely spaced which leads to interactions between circuit parts and makes the use of quasi-static analysis methods unreliable. More sophisticated full-wave techniques which are capable of simulating all aspects of 3-D field interaction at and between discontinuities are usually very CPU-time consuming or memory space intensive. In this paper we will investigate a variety of transitions and module interconnections in more detail with a very powerful and flexible new numerical technique, the frequency-domain TLM method, which was published only recently by Jin and Vahldieck [1][2]. The objectives of this investigation is to simulate such structures and to extract useful information for optimum transition design.

2. Method

A detailed description of the frequency-domain TLM (FDTLM) method has been presented in [2]. Therefore, only some highlights are given here. In the FDTLM method, the space to be analysed is discretized by a transmission line network, as in the conventional time domain TLM (TDTLM) method. However, instead of exciting the network with a single impulse, an impulse train of sinusoidally modulated magnitude is assumed so that the magnitudes of the impulses vary with time described as $e^{j\omega N\Delta t}$, where ω is the modulation frequency, N the time step, and Δt the time interval between impulses. Therefore, if the magnitude of the impulse at a particular reference time is known, the magnitude of the impulse at any other time step can be readily found by multiplying with $e^{j\omega\Delta N\Delta t}$, where $\Delta N\Delta t$ is the total time interval between the reference time step and the time when the observation takes place. Hence, the time iteration procedure necessary in the conventional TDTLM method is no longer needed in the FDTLM method and the entire algorithm operates within one time step at an arbitrary time instance[2]. It should be noted that the practical implementation of the algortihm does not require an actual impulse excitation of the network. The assumption of an excitation as described above is just to simplify the algorithm and to provide the transition between the TDTLM and the FDTLM [2]. The FDTLM method has both the flexibility of the conventional TDTLM and the computational efficiency of frequency-domain techniques, and is particularly suitable for structures with complicated geometries such as the transmission line transitions and interconnect assemblies investigated in this paper.

The general calculation procedure of the frequency-dependent s-parameters with the FDTLM method consist of the following steps. For a given structure we first perform a two dimensional propagation analysis for the attached waveguides, which, in the present cases, are microstrip, CPW or slotline. From this analysis we find the fundamental modal field distribution. Then the discontinuity

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region is excited from one of the waveguides by an incident wave, which is its fundamental propagation mode obtained from the previous two dimensional analysis. The reflected and transmitted waves are calculated and the s-parameters are obtained. Since the field components for different modes can be separated in the FDTLM method[2], the reference plane for the s-parameter calculation may be located right in the discontinuity plane. This property allows to minimize the volume to be discretized by the transmission line network.

3. Numerical Results

Fig.1 shows the frequency-dependent s-parameters for a fineline to CPW transition considering a quarter wave transformer section for better matching of the finline mode, which is the receiver mode in a front-end application. A significantly improved match at about 25GHz is obtained.At this frequency the propagation constant of the intermediate slotline is β=1.1(mm^{-1}) and the quarter wavelength is about $\lambda g=2\pi/4\beta$ = 1.43 mm, which is very close to the physical length of the added section (d_1= 1.5 mm). Varying the gap g between the finline and the CPW, it was found that the results for two different gaps (g=0.1mm and g=3.0mm) are almost identical, indicating that this kind of transition is not sensitive with respect to the distance between the two transmission lines. The reflection from this kind of transition is mainly due to the finline step junction.

Fig.2 shows a uniplanar CPW-microstrip transition with an odd mode excitation. The walls of the metallic enclosure are 2xS2+W1, which means the CPW groundplane and the backmetallization (the latter one is the groundplane for the microstrip line) are on the same potential. Due to the additional substrate layer (air ε_r=1.0) between the back metallization and the alumina substrate (ε_r=9.6), there are no higher order modes up to 40 GHz. As expected, the reflection coefficient increases with frequency because the transition behaves like a transmission line junction with a parasitic shunt capacitor. The characteristic impedances are 54 Ohms for the CPW and 97 Ohms for the microstrip. Also here the reflection coefficient can be reduced by introducing a quarter wavelength (CPW) transformer between both transmission lines. The lowest value for S_{11} in Fig.2 occurs at about 28 GHz which corresponds to β=1.5(mm^{-1}) or a quarter wavelength λg~1.05mm, which is approximately the physical length of the intermediate section (d=1.0 mm).

An overlay transition between a microstrip and a CPW is shown in Fig.3. This transition provides electromagnetic coupling between two transmission lines of 54 Ohms characteristic impedance. Fig.4 illustrates the electrical performance of this junction versus the overlay length. As expected, the coupling between both transmission media increases with overlap area and at the same time the s-parameters become more frequency dependent.

Fig.5 shows the results for an interconnect assembly between microstrip and CPW on different substrates. Alumina substrate (ε_r=9.6) is used for the microstrip line while the CPW is deposited on GaAs(ε_r=12.9). It is found that the reflection of the interconnect increases significantly with frequency, suggesting that this kind of discontinuity can be represented as π-network of shunt capacitors (open-ended microstrip and CPW) and series inductor (wire bond). The shunt capacitances are due to the fringing fields at the open end of the microstrip and the CPW, and the series inductance comes from the bonding wire connecting both transmission lines. The inductance is approximately proportional to the length of the bonding wire[3]. Fig.6 shows the s-parameters as a function of the gap g for two different frequencies f=5 GHz and f=25 GHz. It is obvious that the reflection increases with increasing gap g. This is quite plausible because a wider gap means longer bonding wire and hence larger inductance value with large reflection coefficient. Fig.7 illustrates s-parameters as a function of the bonding wire height $a_m(=a_c)$. Also here, the reflection coefficient increases with increasing height because this results in longer bonding wire lengths.

As reported in [3], the properties of the interconnect assembly of Fig.5 are largely determined by the length of the bonding wire and are quite insensitive to the shape of the bonding wire. In the following we have investigated different shapes of the bonding wire to determine to what extent this assumption is true. For a given wire length, we change both the wire height a_m, a_c and the gap g simultaneously in such a way that the total length of the wire $L=a_m+a_c+g$ is kept unchanged. The results of the s-parameters with different a_m , a_c and g are shown in Fig.8. The curves are quite flat over a large range of dimensions, which confirms the results reported in [3]. In other words, when the bond wire changes from one extreme shape (g=0.3, $a_m=a_c$=0.1, Fig.8) to another (g=0, $a_m=a_c$=0.25, Fig.8) the s-parameters change only by a few percent.

4. Conclusions

This paper has introduced a rigorous analysis of a variety of transmission line transitions and module interconnect assemblies in MMIC's using the frequency-domain TLM method. Numerical results of frequency-dependent s-parameters have been presented which include the effect of finite thickness and conductivity of metallization as well as mode interaction between cascaded discontinuities. The effect of inserting an intermediate section of transmission line between two different transmission media has been analysed and it was found that CPW transitions can be made more broadband. The effects of the bonding wire for module assemblies was investigated. It was found that the properties of the interconnect are largely depended on the total length of the wire and are quite insensitive to the shape of the wire. This is in a good agreement with the experimental observations.

References:

[1]Hang Jin and Ruediger Vahldieck," A frequency domain TLM method," 1992 IEEE Int. Microwave symp. Dig. pp.775-778

[2]Hang Jin and Ruediger Vahldieck, " The frequency-domain transmission line matrix method - A new concept," IEEE Trans. Microwave Theory Tech., vol.40, pp.2207-2218, Dec. 1992

[3]Steve Nelson, Marilyn Youngblood, Jeanne Pavio, Brad Larson, and Rick Kottman," Optimum microstrip interconnects," 1991 IEEE MTT-S Digest, pp.1071-1074,1991,Boston.

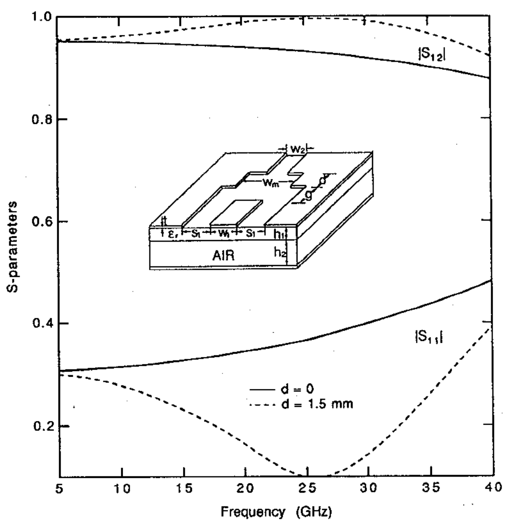

Fig.1 Frequency-dependent S-parameters of a cascaded CPW-finline transition (w_1=0.2 mm, s_1=0.6 mm, w_2=0.4 mm, w_m=1.0 mm, h_1=0.254 mm, h_2=0.4 mm, t=0.003 mm, ε_r=9.6 and the metallization σ=10000 s/mm)

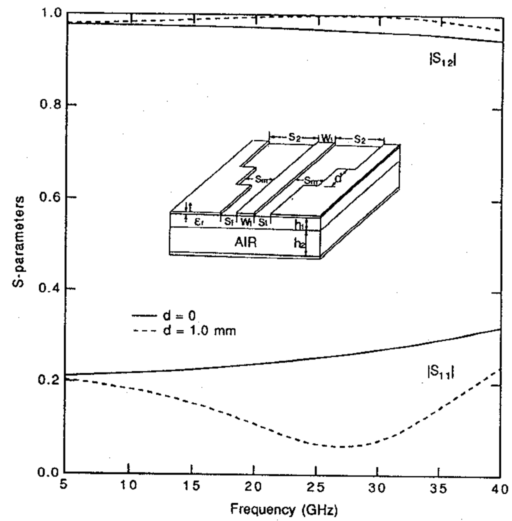

Fig.2 Frequency-dependent S-parameters of a uniplanar CPW-microstrip transition (w_1=0.2 mm, s_1=0.1 mm, s_m= 0.2 mm, s_2=5.0 mm, h_1=0.254 mm, h_2=0.2 mm, t=0.003 mm, ε_r=9.6 and the metallization σ=10000 s/mm)

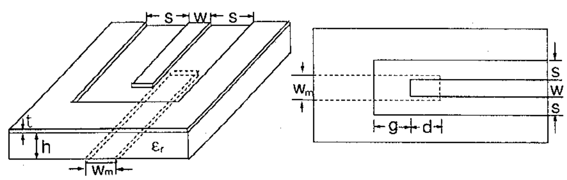

Fig.3 A CPW-microstrip overlap transition.

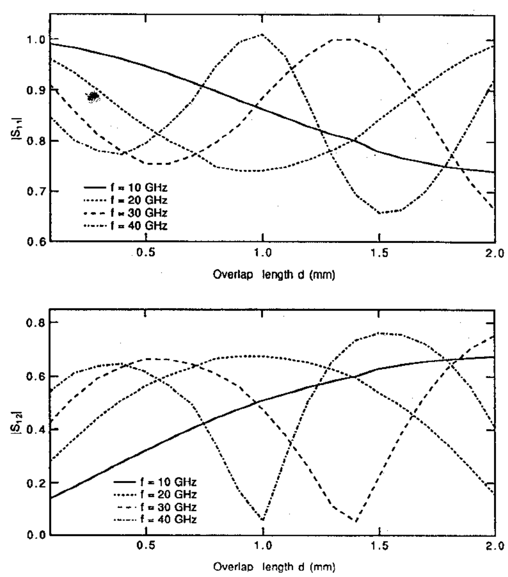

Fig.4 Variation of the s-parameter versus the overlap length d of a CPW-microstrip overlap transition as shown in Fig.3 for four different frequencies. (w=0.2 mm, s=0.1 mm, w_m=0.2 mm, h=0.254 mm, g=0.5 mm, ε_r=9.6 and the metallization σ=10000 s/mm)

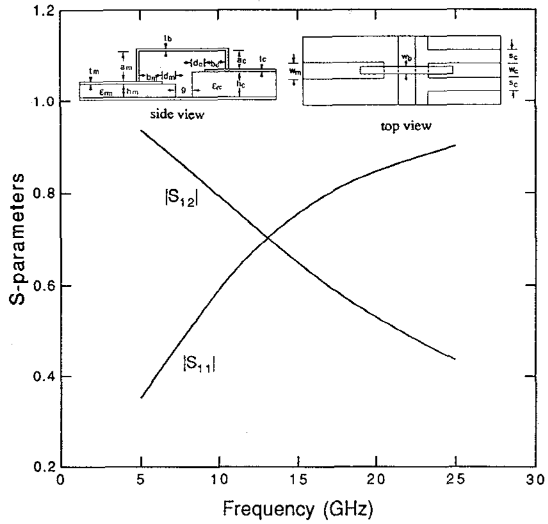

Fig.5 Frequency-dependent s-parameters of the interconnect assembly (b_m=b_c=d_m=d_c=0.1mm, h_m=h_c=0.2mm, g=0.3mm, a_m=a_c=0.25mm, w_m=w_c=0.2mm, w_b=0.1mm, s_c=0.2mm, t_c=t_m=t_b=0, ε_{rc}=12.9, ε_{rm}=9.8)

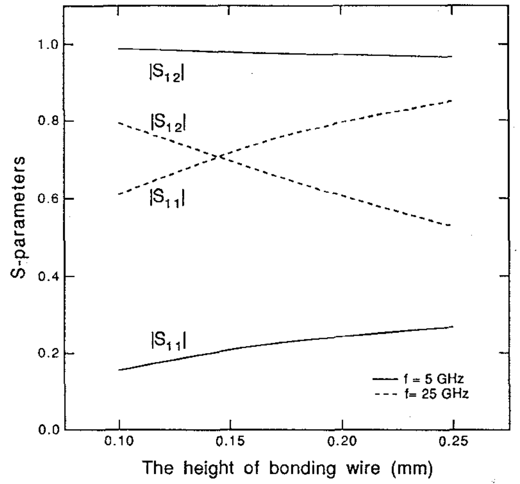

Fig.7 S-parameters as the functions of the height of bonding wire a_m(=a_c) for two different frequencies f=5 GHz and f=25 GHz (b_m=b_c=d_m=d_c=0.1mm, h_m=h_c=0.2mm, g=0.001mm, w_m=w_c=0.2mm, w_b=0.1mm, s_c=0.2mm, t_c=t_m=t_b=0, ε_{rc}=12.9, ε_{rm}=9.8)

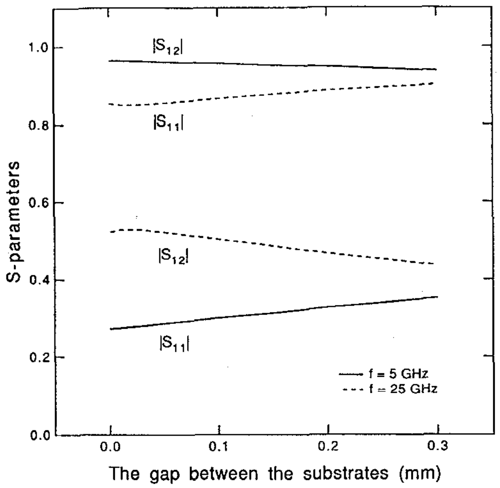

Fig.6 S-parameters as the functions of the gap g between the substrates for two different frequencies f=5 GHz and f=25 GHz (b_m=b_c=d_m=d_c=0.1mm, h_m=h_c=0.2mm, a_m=a_c=0.25mm, w_m=w_c=0.2mm, w_b=0.1mm, s_c=0.2mm, t_c=t_m=t_b=0,ε_{rc}=12.9, ε_{rm}=9.8)

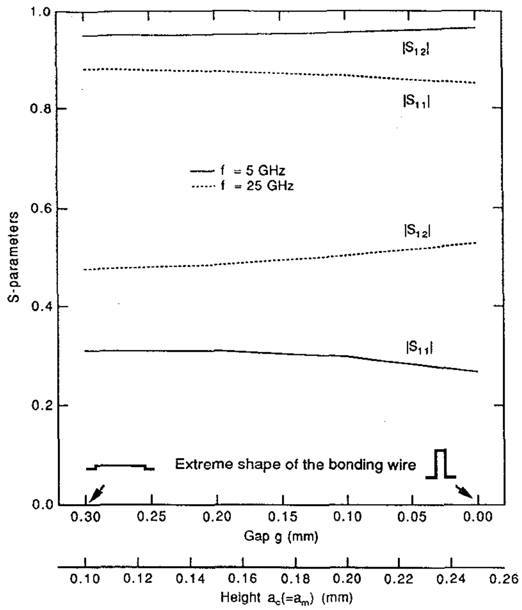

Fig.8.S-parameters as the functions of the height of bonding wire. The gap between the substrates is reduced simultaneously with increasing height so as to keep the total length of the bonding wire constant. (b_m=b_c=d_m=d_c=0.1mm, h_m=h_c=0.2mm, w_m=w_c=0.2mm, w_b=0.1mm, s_c=0.2mm, t_c=t_m=t_b=0, ε_{rc}=12.9, ε_{rm}=9.8)