Full-Wave Modeling of Via Hole Grounds in Microstrip by Three-Dimensional Mode Matching Technique

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***Abstract*—A rigorous full-wave analysis of microstrip via hole grounds is presented using a three-dimensional mode-matching technique in connection with a suitable segmentation of the structure into homogeneous parallelepipedal cells. The adoption of the novel impressed source technique reduces substantially the numerical effort compared to the transverse resonance technique and, in addition to the finite metallization thickness, accounts for possible package interaction. Theoretical results are compared with several experimental data from various sources, including our experiments, showing excellent agreement. Package effects have been observed experimentally and shown to be fully predicted by the theory.**

I. Introduction

THE USE of shunt posts in microstrip transmission lines is a common practice in microwave and millimeter-wave hybrid and monolithic circuits. Via hole throughs are used in multilayer printed circuit boards to connect by vertical pathways microstrips at different layers [1]. They are also used in single layer circuits to obtain wide-band short circuits (via hole grounds). Bandpass filters are another promising application of this technology [2].

The accurate characterization of the microstrip via hole discontinuity is an important issue in the successful design of the circuit. Approximate models of this type of discontinuity has been proposed based on quasi-static formulae [3], or using an equivalent planar waveguide model, i.e. a rectangular waveguide structure with magnetic sidewalls [2]. Though simple and numerically efficient, however, such models have intrinsically limited validity. For instance, they cannot take into proper account possible crosstalk and package effects associated with the presence of several discontinuities or due to the interaction with the package. Rigorous analyses of via hole throughs have recently been published using a purely numerical method such as the FDTD [4]. This method is versatile but highly computer intensive.

In Section II of this paper we present a method for the full-wave analysis of the microstrip-via hole ground discontinuity. The method is based on the 3-D version of the mode-matching technique introduced in [5], [6]. Using suitable segmentation procedures of the 3-D structure along with the impressed source technique, a rigorous yet numerically efficient approach has been developed. In Section III, theoretical results are compared with a number of measured data from various sources as well as with results from a commercial microwave simulator. An excellent agreement between the present theory and experiments is observed in all cases, also when the commercial simulator fails because of the occurrence of package effects.

II. Method of Analysis

Let us briefly recall the basic mode-matching (MM) method for the analysis of a complicated microwave structure, such as a filter, a multiplexer, etc. Actually, the MM method applies directly to simple discontinuity problems, such as steps, irises or bifurcations. In the case of a complicated structure, this is first subdivided into a number of discontinuities connected by transmission line sections. The electromagnetic (EM) field at the plane of each junction is then expanded in terms of the normal modes of one of the waveguides connected to that junction. Applying the microwave network formalism, i.e. associating voltages and currents to the electric and magnetic fields of each mode respectively, one obtains a multiport network model of the junction, where each port corresponds to one mode. In this manner, an equivalent network results for each discontinuity. The analysis of the microwave structure is so reduced to that of the overall equivalent network.

This approach is numerically very efficient for simple discontinuities and in cases when the waveguide modes are known, i.e. when the waveguide cross-sectional geometry allows an analytical solution for the Helmholtz equation. If the waveguide cross-section is a non-separable one, a numerical technique must be applied to compute the modes. For inhomogeneous guiding structures

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(e.g. finlines) the numerical effort is also quite high because of the presence of complex modes.

To alleviate numerical problems in the MM analysis of complicated waveguide structures such as multiport branch guide dividers, an efficient 3-D MM technique has been developed in [5], [6]. Instead of looking at the structure as a number of *discontinuities* connected by transmission line sections, the structure is subdivided into adjacent *volume elements* (or *cells*) connected one to the other by apertures on the common walls. By expanding the EM fields at the apertures in terms of waveguide modes (or any complete set of basis functions) one obtains an equivalent network representation for each volume element. By solving the resultant equivalent network of the structure, the EM field distributions on the apertures and, then, in the whole structure are finally computed. It has been shown that the association of an equivalent multiport network to each volume element instead of to each discontinuity leads to a substantial reduction in the numerical complexity of the algorithm.

Such a technique is now extended to the modeling of 3-D inhomogeneously filled structures, such as the via hole ground on a microstrip line. In contrast with the conventional MM technique, with the adoption of the cellular segmentation technique the presence of different dielectric materials is easily taken into account, no complex modes being involved.

The geometry of a via hole ground is sketched in Fig. 1. The microstrip is enclosed in a metallic box of proper size, i.e. such as not to perturb the reactive fields in the proximity of the discontinuity. This condition has to be satisfied if one has to model the isolated discontinuity. Observe however that the enclosure can be used to represent the actual circuit package. In this case, one obtains a description of the circuit behavior in the actual conditions, i.e. including parasitics effects due to the package.

It could be noted that, because of the fabrication procedure [7], the via hole is generally cone-shaped. In the EM modeling of such a discontinuity, however, a considerable reduction of the complexity of the problem is achieved considering the via as a cylinder of rectangular cross-section. This is assumed to introduce only minor alterations. This hypothesis has been confirmed *a posteriori* by comparison against measured data.

Taking advantage of the symmetry along x and z, we can analyze just one quarter of the structure, bounded by electric and/or magnetic walls as depicted in Fig. 2(a). This structure can further be segmented into four parallelepipedal regions $R_n (n = 1, \cdots 4)$. (Note that account is taken of the finite thickness of the metal strip). Region 1 corresponds to the region above the strip, region 2 is the volume aside the strip, and regions 3 and 4 are in the dielectric layer. The various regions interface one each other, and are shown separately in Fig. 2(b). Note that each region is homogeneously filled. Moreover, the three rectangular apertures between adjacent regions (1–2, 2–3, 3–4) have contours laying on perfectly conducting walls. As a consequence, rectangular waveguide modes are an

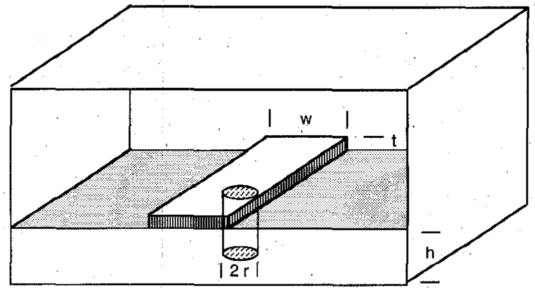

Fig. 1. Geometry of the boxed via hole ground.

appropriate basis for the field representation on such apertures.

Consider now a generic region R_n. The EM field inside the region is excited through N_p apertures $S_i (i = 1, 2, \cdots N_p)$ and can be computed from the knowledge of the electric field distribution on such apertures. Using Huygens' principle [8], the magnetic field in the region is expressed as

$$H(r) = j\omega\epsilon \oint_S n \times E(r') \cdot \underline{\underline{G}}(r, r')\, dS'$$
$$= j\omega\epsilon \sum_{i=1}^{N_p} \int_{S_i} n \times E(r') \cdot \underline{\underline{G}}(r, r')\, dS' \qquad (1)$$

the integral over S being reduced to the only portions S_i where $n \times E$ is not zero. In (1) the quantity $n \times E$ can be interpreted as a magnetic surface current flowing on S_i. $\underline{\underline{G}}$ is the dyadic Green's function for the magnetic field in R_n. It is a solution of

$$\nabla \times \nabla \times \underline{\underline{G}} - k^2 \underline{\underline{G}} = \underline{\underline{I}}\delta(r - r') \qquad \text{in } R_n \qquad (2)$$

with boundary conditions

$$n \times \nabla \times \underline{\underline{G}} = 0 \qquad \text{on } S \qquad (2')$$

where $k^2 = \omega^2\mu\epsilon_n$, $\underline{\underline{I}}$ is the unit dyadic, δ is the Dirac delta function.

The relationships among field quantities at the openings S_i can be formulated in terms of a generalized admittance matrix by expanding the tangential E- and H-fields into a suitable set of vector basis functions $\varphi_k^{(i)}$

$$n \times E^{(i)} = \sum_{k=1}^{N_i} V_k^{(i)} \varphi_k^{(i)} \qquad (3a)$$

$$H^{(i)} = \sum_{k=1}^{N_i} I_k^{(i)} \varphi_k^{(i)}. \qquad (3b)$$

The vector basis functions in (3a) and (3b) need not to be the same, but this choice leads to a Galerkin-type procedure having variational properties. The $\varphi_k^{(i)}$'s can be any complete orthonormalized set of vector basis functions. They have been chosen, as usual, as the modal eigenvectors of a waveguide having S_i as the cross section. The series in (3a,b) are truncated to N_i terms for numerical reasons.

The expansion coefficient $V_k^{(i)}$ and $I_k^{(i)}$ represent the equivalent voltage and current on the kth electrical port

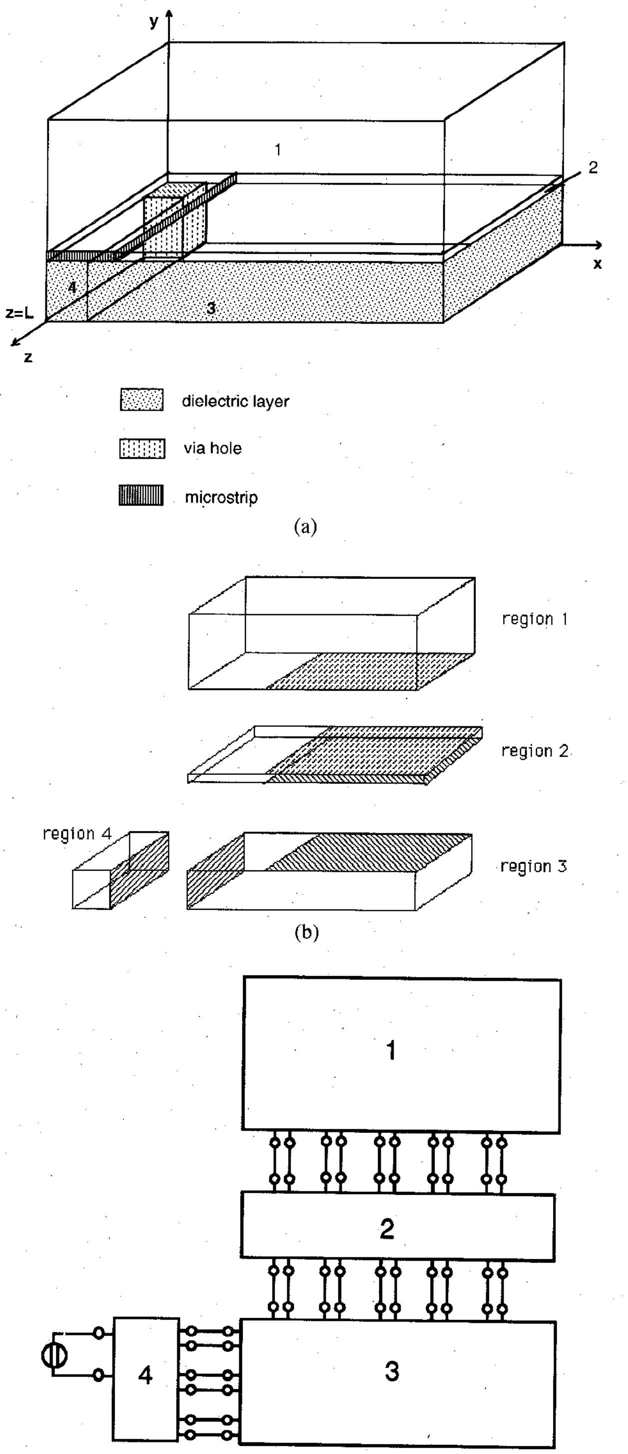

Fig. 2(a). Structure used for analysis. The microstrip is enclosed in a box with perfectly conducting walls. The analysis is carried out by using even/odd excitation, thus placing a magnetic/electric wall at $z = 0$. For symmetry reasons, the wall placed at $x = 0$ is a magnetic one. (b) Segmentation of Fig. 2(a). (c) Network description of the boxed microstrip with a via hole ground. The generator connected to region 4 corresponds to the impressed electric field.

on the output S_i. They are related to the respective field quantities on S_i by

$$V_k^{(i)} = \int_{S_i} \boldsymbol{n} \times \boldsymbol{E}^{(i)}(\boldsymbol{r}) \cdot \boldsymbol{\varphi}_k^{(i)}(\boldsymbol{r})\, d\boldsymbol{r} \tag{4a}$$

$$I_k^{(i)} = \int_{S_i} \boldsymbol{H}^{(i)}(\boldsymbol{r}) \cdot \boldsymbol{\varphi}_k^{(i)}(\boldsymbol{r})\, d\boldsymbol{r}. \tag{4b}$$

Combining (4b) with (1) and (3a) we obtain

$$I_m^{(j)} = \sum_{i=1}^{N} \sum_{k=1}^{N_i} Y_{mk}^{(ji)} V_k^{(i)} \tag{5}$$

where

$$Y_{mk}^{(ji)} = \int_{S_j} \int_{S_i} \boldsymbol{\varphi}_m^{(j)}(\boldsymbol{r}) \cdot \underline{\underline{G}}(\boldsymbol{r}, \boldsymbol{r}') \cdot \boldsymbol{\varphi}_k^{(i)}(\boldsymbol{r}')\, d\boldsymbol{r} d\boldsymbol{r}' \tag{6}$$

defines the generalized admittance matrix of R_n. This quantity represents the amplitude of the mth current component entering the jth opening produced by a unit kth component of the voltage at the ith opening, all the other voltage components being zero. Observe that each physical aperture is represented by a set of electrical ports.

Each region is therefore completely described, from a network point of view, by its generalized admittance matrix. The four admittance matrices corresponding to regions 1–4 are then interconnected (Fig. 2(c) to enforce the continuity of the currents, i.e. of the tangential components of the magnetic fields at the interfaces. This step is equivalent to solve the set of coupled integral equations resulting from the continuity of the tangential components of the magnetic fields at the interfaces. The presence of the voltage source will be discussed later on.

Computation of the Admittance Matrices

Let us first observe that to compute the generalized admittance matrix (6) no matrix inversion is necessary. As pointed out in [9], this is due to the fact that the admittance matrix is the natural representation of a cavity with apertures on conducting walls.

The numerical evaluation of (6) requires the computation at many frequency points of the coupling integrals between the basis functions (that are frequency independent) and the Green's function. To reduce the numerical effort, one should take advantage of suitable fast convergent expressions for the involved Green's functions, bearing also in mind their frequency dependence. An exhaustive discussion on this aspect is outside the scope of this paper. Let us just mention a few basic points.

Different expressions of the Green's functions are available for closed rectangular cavities [8], [10]. Besides the eigenfunction expansion (involving summation over a triplet of indexes) also expansion in terms of modes along the various directions (involving double instead of triple summation) can be used. The numerical advantage of summing over just two indexes instead of three is apparent. The choice of expanding the Green's functions in terms of modes along x, y, or z depends on the convergence of the relative series, which in turn depends on the relative position of the ports [6].

Consider, in fact, that the coupling between two ports placed along, say, the x-direction, occurs through a finite number of x-propagating modes and a number of x-decaying modes below cutoff. The latter number is smaller the larger the distance between the ports. On the basis of such considerations, fast convergent series can be obtained for the mutual admittance between any pair of ports. Details of this procedure can be found in [6].

Another important aspect from the numerical point of view is the frequency dependence of the series for the Green's function. It is desirable that the frequency dependence could be easily extracted so that all the coupling integrals can be computed just once. This is possible with the eigenfunction expansion and, when using the waveguide mode expansion, only when the involved apertures $S_i S_j$ are on parallel sides. For apertures on adjacent orthogonal sides of a region, the choice of the most efficient representation for Green's function is somewhat questionable, depending also on the number of frequency points.

Impressed Source Technique

The electromagnetic field analysis has led to the expression (6) for the generalized admittance matrix of the entire structure. To extract the parameters of the discontinuity one could apply the tangent method used in connection with the transverse resonance technique [11]. At each frequency point, repeated field analyses must be performed in the numerical search for the resonant dimensions of the enclosing cavity, which has therefore no given size.

The impressed source technique introduced here reduces substantially the computer effort by solving a deterministic rather than an eigenvalue problem. A number of advantages are obtained: i) only two field analyses, one for the even mode one for the odd mode, are sufficient to determine the 2 unknown parameters of the lossless symmetrical discontinuity; ii) since the enclosing cavity has fixed dimensions, it can be used to simulate the actual package, in such a way as to closely model the actual operating conditions and account for possible package interaction; iii) the coupling integrals need to be calculated only once for the given geometry of the cavity.

The excitation of the field in the structure can be obtained by various field distributions. In the present case, an electric field distribution has been impressed at the front side ($z = L$) of region 4 (Fig. 2(a)). Correspondingly, in the multiport representation of region 4 (Fig. 2(c)) a voltage generator appears representing the impressed field. A single generator has been indicated in Fig. 2(c) for simplicity. Actually, the number of generators can be higher as it equals the number of modal basis functions used to represent the impressed field at the front side of region 4. In the proximity of the source, as well as near the via hole discontinuity, higher order microstrip modes are excited. However, provided that the structure is long enough (along z), only the fundamental mode exists in the central region where higher order modes are died out. As shown below, the scattering parameters of the microstrip via hole discontinuity are then readily computed by evaluating the E-field distribution along z in the central region.

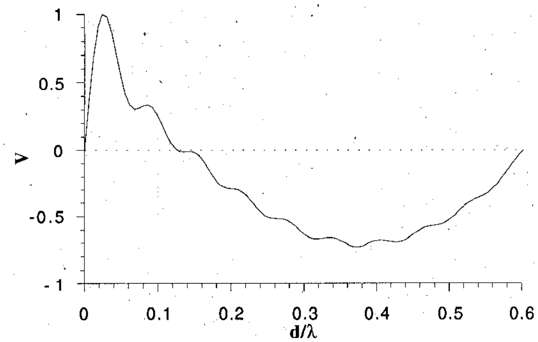

Fig. 3. Normalized voltage distribution along the strip computed as a function of the distance $d = L - z$ from the source.

The electric field distribution of the dominant mode of the microstrip can also be selected as the excitation to be impressed at the front sides of regions 1, 3, 4 (Fig. 2(c)). This eliminates the excitation of higher order modes from the source, but requires additional effort to precompute the modal field distribution and additional coupling integrals.

Extraction of the Scattering Properties

Once the equivalent circuit of Fig. 2(c) has been solved, we know the field at the boundaries of the $n = 1, 2, 3, 4$ regions constituting the segmented structure. To compute the scattering parameters of the via hole, it is now necessary to determine the incident and reflected wave amplitudes of the fundamental mode. In a region far enough from both the impressed source and the discontinuity, higher order modes have negligible amplitudes so that the electric field behavior along z is of the form

$$V(z) = V^+ e^{-j\beta z} + V^- e^{j\beta z} \qquad (7)$$

The propagation constant β of the microstrip being computed in advance, the values of V^+, V^- can easily be computed by a least square approach using sampled values of $V(z)$ taken apart from both the source and the discontinuity.

Fig. 3 shows an example of the voltage along the strip conductor, computed from the source ($d = 0$) to the surface of the via. The computation has been made at the interface (aperture) between region 3 and 4. This avoids the computation of the fields inside the cells, thus simplifying the numerical effort. The operating frequency is $f = 10$ GHz. The number of spectral terms used in the z direction is 20. The presence of higher order modes in the proximity of the via is not evident. It is noted, on the other hand, the presence of the Gibbs phenomenon in the proximity of the source, as well as a small ripple (especially near the source) due to the series truncation. Possible errors in the scattering parameter extraction due to such a ripple are minimized by the least square procedure.

It may be useful to mention that to avoid the computation of β, an alternative procedure exists based on the Prony algorithm [12]. This algorithm computes the values of a_i and α_i of a function

$$y(z) = \sum_{i=1}^{N} a_i e^{\alpha_i z} \tag{8}$$

when samples of $y(z)$ are given at 2N equidistant points. This algorithm can thus be applied also in the presence of higher order modes.

III. Results and Discussion

The 3-D MM technique described has been checked against a number of experimental data on various structures from different sources.

Fig. 4 shows computed and measured transmission parameter $|s_{21}|$ of a via hole grounded microstrip line realized on a low dielectric constant ($\epsilon_r = 2.32$) substrate. Experimental data [13] in frequency range 2–15 GHz are compared with the static formula (2) of [3], as well as with theoretical data computed by a commercial software package and by the present technique. The static inductance is seen to provide a very good approximation in the lower frequency range, while the commercial software simulation becomes increasingly more accurate at higher frequencies. The present theory, on the contrary, is in very close agreement in the whole frequency range. Observe that in our rigorous analysis the thickness of the microstrip has been taken into account; in particular, a strip thickness of 10 μm has been assumed.

A further comparison has been made between theory and experiments on a via hole on a GaAs substrate manufactured by GEC-Marconi. The geometry of the via is sketched in Fig. 5(a)–(c) shows the scattering parameters $|s_{21}|$ and $|s_{11}|$ in the range 2–20 GHz. Experimental data, provided by GEC-Marconi [14], are compared with the present theory and with the commercial software simulation. In Fig. 5(c), the curve marked RL has been obtained considering the via as a shunt resistance R in series with an inductance L, the foundry data being $R = 0.187\ \Omega$ $L = 0.022$ nH [14]. The excellent accuracy of such lumped model is to be ascribed, on the one hand, to the optimization of the element values to fit the experimental data of the specific via and, on the other hand, to the use of higher dielectric constants. This in fact reduces the dimensions of the microstrip via, thus extends the validity of the low frequency range approximation. The better performance of vias on high dielectric constant substrates is confirmed comparing Figs. 4 and 5.

Results computed by the present theory appear in excellent agreement with the experiments. A small disagreement on $|s_{21}|$ is observed at the lowest frequency (-46 instead of -41 dB). This is to be ascribed to the loss not being included in the model. It can be concluded that the effect of the losses has some effects in the low frequency range only, where, on the other hand, the via behaves as a good short. As the frequency increases, reactive effects prevail.

The effect of varying the via's diameter has been investigated on the basis of the theoretical model described. The results for the same structure of Fig. 4 except for different via's diameters are shown in Fig. 6. As expected, the short circuit effect of the via is improved in the whole frequency range as the diameter increases.

Additional measurements have also been performed,

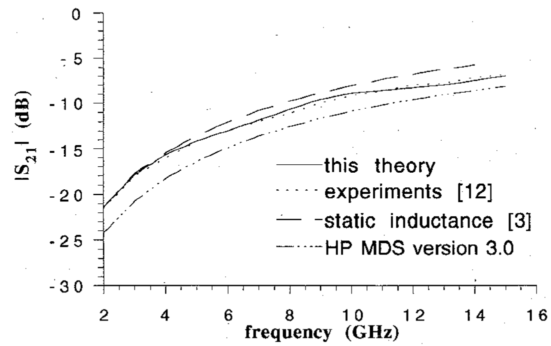

Fig. 4. Computed and measured insertion loss of a via hole ground in microstrip with $w = 2.3$ mm, $h = 0.794$ mm, $2r = 0.6$ mm and $\epsilon_r = 2.32$.

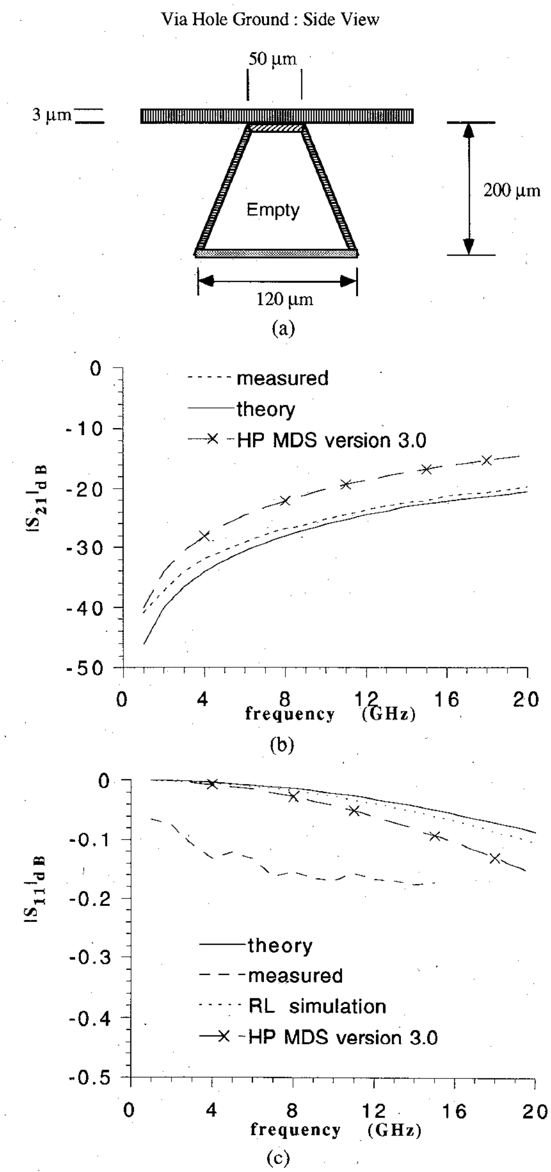

Fig. 5(a). Schematic of a microstrip via hole ground on GaAs ($e_r = 12.9$). Strip width is $w = 570$ mm. (b), (c) Theoretical and experimental results for the via hole of Fig. 5(a).

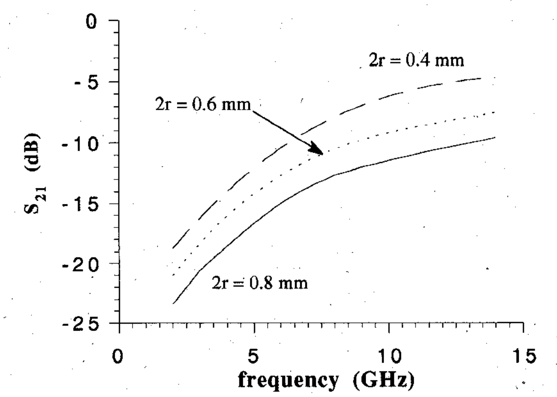

Fig. 6. Computed insertion loss of the same via of Fig. 4, but for different via's diameters.

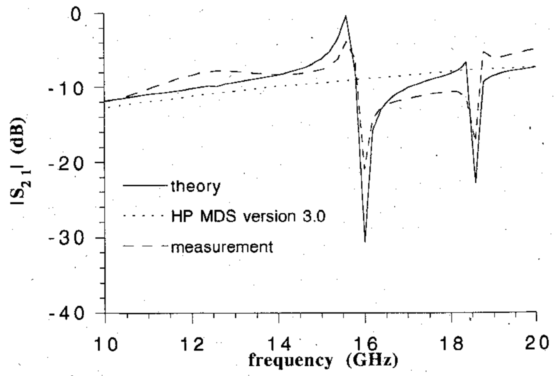

Fig. 7. Computed and measured insertion loss of a via on a 50 Ω microstripline in a package. Via hole diameter = 0.7 mm, substrate permittivity ϵ_r = 2.54; substrate thickness h = 0.79 mm; package size = 10 × 10 × 25 mm.

implementing several samples on various substrates (OAK601, CuClad, etc.), with ϵ_r in the range 2.45–4.8, substrate thickness between 0.79 and 1.524 mm, and metal thickness from 17 to 35 μm. The lines were assembled in a brass enclosure of size 10 × 10 × 25.5 mm, using SMA connectors. Experiments were carried out with via hole diameters of 0.7 and 0.8 mm. Both completely filled and with 0.1 metal thickness vias were tested, showing no appreciable difference in electrical behavior. A HP 8510 network analyzer system was used to test the samples in the 0.045–18. GHz band. To reduce as much as possible the effects of the transitions, a transmission-reflection-line calibration standard was implemented for each substrate. Measurements have been performed using both TRL and TRM calibration techniques.

Fig. 7 shows one of the results obtained. Again, the measurements are compared with both our theory and the commercial software package. The interesting result here is the occurence of package resonances which strongly interfere with the via's behavior. The interaction with the package is fully predicted by the present theory, while, of course, it is ignored by the commercial simulator.

IV. Conclusions

A rigorous full-wave analysis of microstrip via holes has been presented. The method is based on a 3-D mode-matching associated with a suitable segmentation of the structure into parallelepipedal elements. The method applies a novel impressed source technique which overcomes some inefficiencies of tangent method adopted in the transverse resonance technique (TRT) [11]. Besides the possibility of accounting for the finite metal thickness, as with the TRT, the impressed source approach substantially reduces the numerical expenditure and at the same time makes it possible to account for possible package effects. Theoretical results have been compared with several experimental data from various sources, including original experiments. An excellent agreement has been observed in all cases. In particular, the possibility of predicting package effects has been demonstrated. Losses in practical circuits are seen to affect the low frequency behavior only. At higher frequency reactive effects associated with the via hole discontintuity prevail.

Acknowledgment

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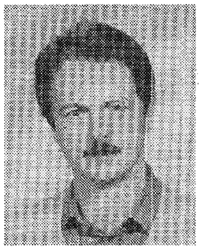

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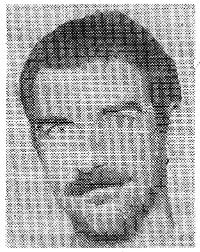

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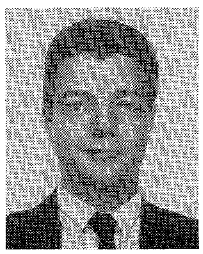
Soon after graduating he joined the Electronic Engineering Department of the Florence University, where he currently works as a Researcher Fellow. Currently he teaches "Electronic devices" at the University of Perugia. His main interest is in the field of microwave devices and circuits, with a particular attention to radar and telecommunication applications.



Luca Roselli was born in Florence in 1962. He received the "laurea" degree in "electronic engineering" in 1988 from the University of Florence.

In 1991 he joined at the Institute of Electronics of the University of Perugia as Researcher Fellow. His field of interest is mainly in microwave device characterisation and application