A GENERAL APPROACH TO THE RESONANCE MEASUREMENT
OF ASYMMETRIC MICROSTRIP DISCONTINUITIES

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Summary

The paper describes an experimental procedure suitable for broad-band characterization of two-port microstrip discontinuities. The resonance frequencies of a cavity embedding the discontinuity under test and of a suitable set of reference resonators are measured and computer processed to obtain the scattering parameters of the discontinuity itself. The method features extreme ease of application thanks to the limited number and simple topology of the required test circuits, as well as highly accurate and repeatable results.

Description of the method

Fig. 1 is a picture of the test and reference circuits to be used for measuring a) a microstrip impedance step and b) a microstrip right-angle bend over the 0 to 18 GHz range. The standard topology of the test circuit is schematically represented in Fig. 2.

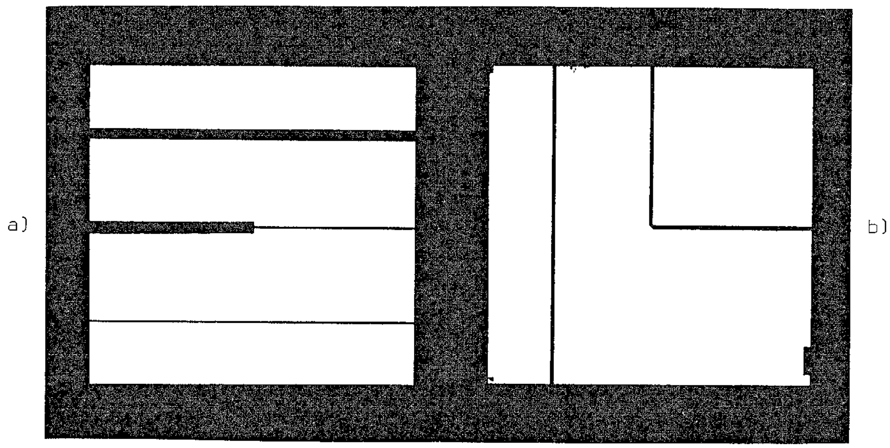


Fig. 1
Pictures of test and reference circuits used for broadband measurement of: a) an impedance step; b) a chamfered right-angle bend, etched on 2"x2" alumina slabs

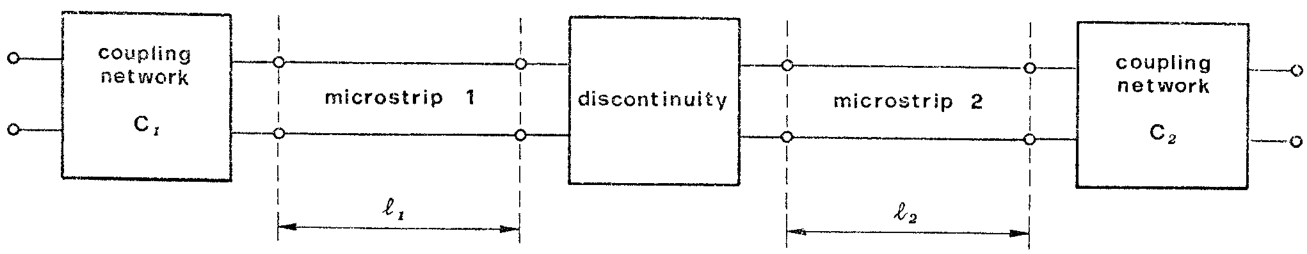


Fig. 2
Schematic representation of microstrip test resonator

The network consists of two sections of uniform microstrip line separated by the discontinuity under test and loosely coupled to the outside by means of the coupling networks C_1, C_2. The quantity to be measured is always represented by the scattering matrix of the discontinuity normalized with respect to the wave impedances of the fundamental (i.e., quasi-TEM) modes of microstrips 1 and 2, respectively. Furthermore, the discontinuity is modelled as a reciprocal loss-free network. Thus the following representation can be used for its scattering parameters:

$$\begin{aligned} S_{11} &= u \exp\{j\phi\} \\ S_{21} = S_{12} &= \sqrt{(1-u^2)} \exp\{j\psi\} \\ S_{22} &= -u \exp\{j(2\psi-\phi)\} \end{aligned} \qquad (1)$$

where u,ϕ,ψ, are unknown functions of frequency to be determined. In turn, these can be replaced by a set of discrete scalar unknowns making use of approximation methods. In other words, each of the above quantities is expressed as a function of frequency depending on a number of undetermined parameters now representing the true unknowns of the problem. The kind of functional dependence of u,ϕ,ψ on frequency can usually be chosen in two different ways, that is

1) by resorting to classic curve-fitting schemes such as polynomial or cubic-spline approximation;
2) by use of some sort of equivalent circuit for the discontinuity parasitics.

While the former is certainly better-suited for obtaining extremely accurate results, the latter will generally provide a deeper insight into the physical behavior of the discontinuity.

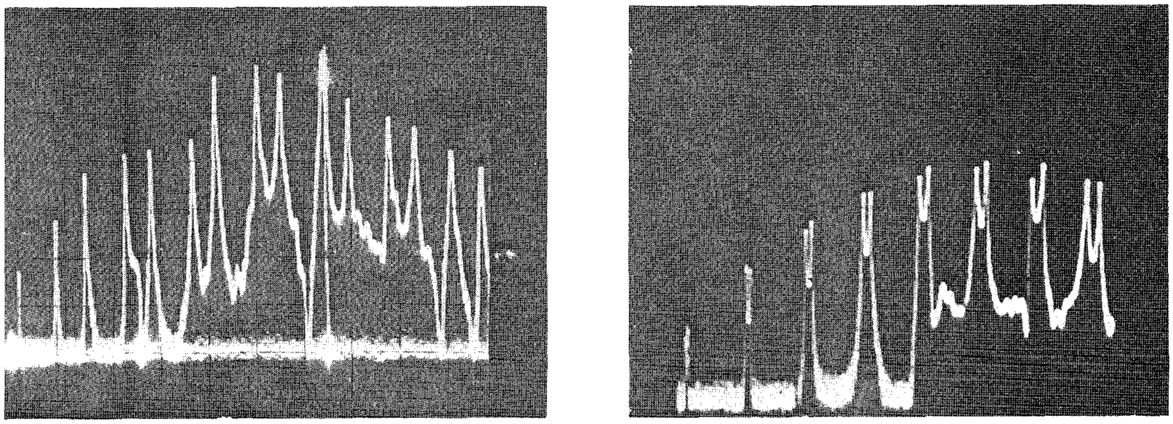
In order to find the unknowns, a set of experimental equations must be written down. The best way of doing this for the present purposes is to make use of the resonances of the test circuit, since at resonance the effects of otherwise minor circuit parameters (such as the discontinuity parasitics we are interested in) may be greatly emphasized [1]. In practice, the coupling networks in Fig. 2 are realized by means of very small capacitive couplings at both ends of the test circuit, thus obtaining a transmission-type resonant cavity. Typical resonance patterns (magnitude of transmission coefficient versus frequency) are given in Fig. 3 for the cases of a) a microstrip impedance step and b) a gap in microstrip.

ches two reference circuits consisting of uniform microstrips having the same widths as lines 1 and 2 of the test circuit. Obviously one reference circuit will be enough (Fig. 1 b) in the case of a symmetric discontinuity. As it clearly appears from Fig. 1, making $\ell_1=\ell_2$ both references will have the same length $2\ell_1$, so that at the n th resonance frequency of reference number i, namely $f_n^{(i)}$, we will have

$$\frac{\ell_i + \ell_{oi}}{v_{pi}} = \frac{n}{4f_n^{(i)}}, \qquad (4)$$

$$i=1,2; \quad n=1,2,3, \ldots .$$

According to the above, the measurement procedure may now be outlined as follows:



a) b)

Fig. 3
Picture of resonance patterns of 2" microstrip cavities embedding: a) an impedance step; b) a 50 μm gap

A definite advantage of the transmission-type cavity is that no problems of critical coupling are encountered, so that the cavity may be coupled as loosely as desired. In particular, the coupling coefficients can always be made so small that the resonance frequencies be unaffected by the presence of the coupling networks. When this is the case, the condition for resonance of the test circuit may be analytically stated as follows:

$$u \sin(\psi-\phi+\theta_1-\theta_2) - \sin(\psi-\theta_1-\theta_2) = 0 \qquad (2)$$

where

$$\theta_i = \frac{\ell_i + \ell_{oi}}{v_{pi}}, \quad i=1,2 \qquad (3)$$

and

ℓ_i = physical length of i th microstrip

ℓ_{oi} = equivalent open-end length of i th microstrip

v_{pi} = phase velocity of i th microstrip.

In (2), (3), ℓ_{oi} and v_{pi} should be considered, in general, as frequency-dependent quantities.
To obtain an accurate characterization of the discontinuity, it is essential that (3) be experimentally determined for the same substrate used to fabricate the test circuit. For this reason, on the dielectric slab supporting the microstrip cavity (see Fig. 1) one also et-

1) The reference lines are coupled to the measuring instrument (typically a network analyzer) and their resonance frequencies are accurately detected by a digital frequency meter.
2) Explicit expressions for the frequency-dependent phase delays (3) are obtained by means of (4) and (say) a cubic-spline approximation algorithm.
3) The test circuit is coupled to the measuring instrument and its resonance frequencies are detected.
4) The unknowns are found by numerical optimization so as to provide the best fit between the solutions of (2) and the measured resonance frequencies in a minimax or a least-square sense.

A few remarks are in order. First note that the measurement does not rely upon an independent characterization of the microstrip open-end effect. In fact, the latter is accounted for automatically together with microstrip dispersion after performing the steps 1) and 2), thus completely avoiding the lengthy procedure described in Ref. 2. Furthermore, the use of least-square approximations of the discrete experimental data is useful for averaging out random measurement errors and thus improving accuracy. Finally, it has been found by computer simulation that the presence of losses both in the microstrips and in the discontinuity does not appreciably affect the resonance frequencies of the transmission-type cavity. Thus any dissipative effects may be completely neglected with no loss of accuracy as far as the reactive components of the discontinuity parasitics are being investigated.

Sample results

In this section we report the results obtained by application of the method described above to a typical microstrip discontinuity. The case considered is the impedance step shown in Fig. 1 a). The microstrip circuit was etched on 0.635 mm alumina (ε_r = 10 nominal) and the strip widths were 1.302 mm and 0.142 mm, corresponding to zero-frequency characteristic impedances of 32.5 Ω and 85 Ω, respectively.
The way chosen to describe the impedance step was to model the discontinuity parasitics by means of suitable equivalent circuits. Two different topologies, illustrated in Figs. 4 and 5, were investigated.

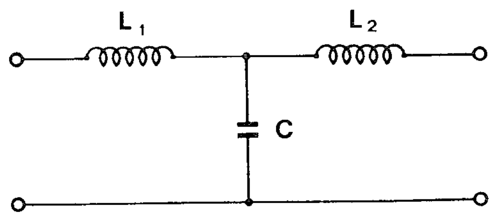


Fig. 5
Lumped equivalent circuit of microstrip impedance step

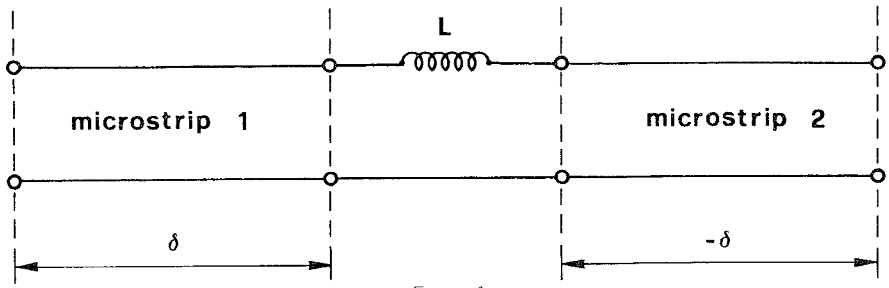


Fig. 4
Lumped-distributed equivalent circuit of microstrip impedance step

With the circuit of Fig. 4, the step parasitics are essentially represented by a lumped inductance, and δ has the meaning of a reference-plane shift [3]. The optimum values for this case were found to be

$$L = 79.2 \text{ pH}$$
$$\delta = 0.156 \text{ mm} \quad , \qquad (5)$$

yielding a maximum shift between computed and measured resonance frequencies of 15.8 MHz, and a maximum relative error of 0.15%. This truly represents an excellent accuracy.
A comparison of (5) with the results obtained by well-known theoretical calculations is interesting. Making use of the mode-matching method [4], the discontinuity parasitics are also modelled by a lumped inductance,which is frequency-dependent and may be very closely approximated by

$$L(f) = 115.8 + 0.0454\ (f - 22.4)^2 \qquad (6)$$

where L(f) is given in pH and f in GHz. This corresponds to an average inductance of 125.2 pH over the 0 to 18 GHz range. Thus for the present case the mode-matching method is found to overestimate the discontinuity inductance by about 60%. On the other hand, the static result of Ref. 5, giving L ≃ 59 pH, shows better agreement with the measured data, the relative error being about -25%.
As for the reference-plane shift δ, no data is available in the literature for comparison. However, it is interesting to note that the experimental value of δ for the case considered is exactly coincident with the equivalent open-end length of the wider microstrip as computed by Itoh [6] (Fig. 6 of Ref. 6 actually yields $\Delta\ell$ ≃ 0.16 mm).

The circuit of Fig. 5 was found to be substantially equivalent to the previous one. The related optimum values were

$$L_1 = -15.6 \text{ pH}$$
$$L_2 = 29.2 \text{ pH} \qquad (7)$$
$$C = 0.0263 \text{ pF} \ ,$$

with maximum errors of 15.3 MHz and 0.14%. The measured discontinuity capacitance is very close to that computed by static methods [7], which is C ≃ 0.03 pF. Note, however, that the simultaneous inclusion of the static inductance *and* capacitance in the equivalent circuit would lead to overestimating the discontinuity parasitics by about 70%.

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