R-1

MICROSTRIP MEASUREMENTS

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Abstract

Extensive measurements are essential for the characterisation of microstrip and design using this medium. Resonator methods are significant and these are given prominence in this tutorial/overview paper. Brief considerations are included of parallel-coupled microstrips, Q-factor measurements, and TDR techniques.

Introduction

The accurate design of microstrip circuits requires comparably accurate and reliable information on the behaviour of the microstrip lines to be used.

Microstrip parameters are subject to variations due to tolerances in substrate permittivity and thickness, and in the metallisation pattern.[1] Accurate microstrip design formulas are therefore only of limited use and theoretical descriptions of behaviour must be supported by careful measurements. Resonator methods are useful for many of the measurements and several are described here for both single and parallel-coupled microstrips.

Time-domain reflectometry (TDR) techniques are also briefly considered.

Instrumentation

The great majority of microwave measurements are best performed in the frequency domain because, with present systems technology, a high accuracy is achievable and a large amount of information is then potentially available; particularly using a swept-frequency system in a network analyser. However, there exist some requirements where time-domain measurements are valuable and high resolution time-domain reflectometer (TDR) systems should then be used. Pulse rise times of less than 28 ps are specially available with TDR equipment.

At the most sophisticated level. operating in the frequency domain, computer-augmented fully automatic measurement systems are available in network analysis; the cost is naturally relatively high and can only really be justified when high accuracy is required. For many microstrip measurements such comprehensive systems are not necessary and, indeed, often only scalar measurements are sufficient. The microstrip circuit under test is typically connected to the test port of a dual directional coupler via a coaxial-to-microstrip or waveguide-to-microstrip transition. The need for leveling is essentially removed by automatically determining the voltage ratio:

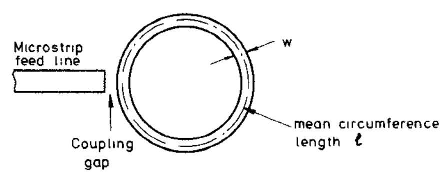
$$\frac{V_r}{V_i} = \frac{\text{reflected signal (from microstrip circuit)}}{\text{incident signal}}$$

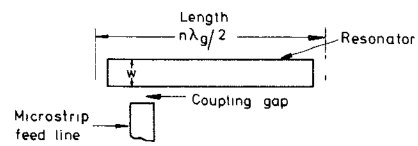
This ratio is automatically measured over the full frequency range of interest. A frequency resolution of ±2 MHz on 10 GHz is typically achieved with such techniques and relative amplitudes to 0.25 dB/cm can be resolved.

Resonator Methods

If a microstrip line is formed as some closed loop on a substrate or left open or short circuited at both ends, it represents a resonator at certain microwave frequencies. Excessive loading effects, which would otherwise seriously effect the measurements are minimised by loosely coupling such a test resonator through to the measurement port of the network analyser.

It appears that the *ring resonator* was first introduced by Troughton [2] and the most conventional circular form is shown in figure 1(a). These resonators can be used for dispersion measurements and, occasionally, Q-factor evaluations.





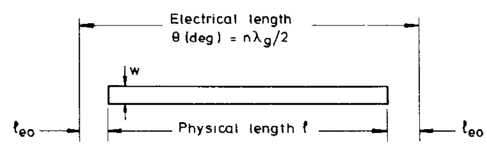


Figure 1. Ring resonator (a), side-coupled open-circuited resonator (b) and end-effect equivalent lengths (c)

For any order n of resonant frequency f and length ℓ the effective microstrip permittivity may then be expressed as

$$\varepsilon_{eff}(f) = \left(\frac{nc}{2\ f\ \ell}\right)^2 \qquad \ldots(1)$$

where c is the free space velocity of light: $2.9979.10^8\ ms^{-1}$.

Due to curvature effects the length ℓ is *not* the mean physical circumference, but it *can* be determined satisfactorily.[3] Difficulties can ensue from the fields in the coupling gap region, field interactions across the ring, or the use of 'large diameter' rings (variations in substrate permittivity and thickness).

The side-coupled open-circuited resonator of Fig 1(b) is sometimes used.[2] Open-circuit end-effect equivalent lengths (Fig 1(c)) then significantly affect the measurement. Coupling problems are evident and the proper location of the feedline along the resonator is difficult.

Provided the end-effect lengths may be considered substantially independant of physical lengths then *pairs* of resonators (Fig 2) may be used to evaluate dispersion.[4] The main result is then given by [1]:

$$\varepsilon_{eff}(f) = \left\{ \frac{nc\ (2f_1 - f_2)}{2\ f_1\ f_2\ (\ell_2 - \ell_1)} \right\}^2 \qquad \ldots(2)$$

where f_1 and f_2 are resonant frequencies. This technique has been very successfully used for accurate dispersion measurements to 18 GHz.[5]

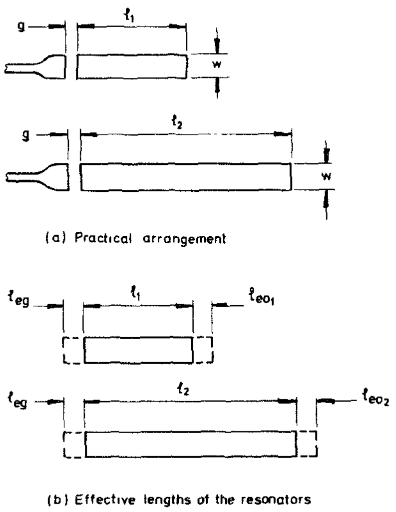


Figure 2. Open-circuited pair of straight resonators

In many filters, matching arrangements, and other circuits data is required on open-circuit end-effect equivalent lengths. One method for the determination of these lengths is shown in Fig 3. The quarter-wavelength line, of physical length ℓ_2, transforms the open circuit at the fore-shortened end of ℓ_2 to an effective short circuit in the vicinity of the gap g_2. This is the crucial feature of this method, because it means that with two gaps the initial circuit can be solved (with lengths ℓ_1) involving gap-end-effect extensions. For 'stage two' of the measurements the quarter-wavelength line ℓ_2 is removed by chemical etching. If we assume that the fore-shortened open-circuit end-effect lengths are all identical and equal to ℓ_{eo} and the gap-end-effect lengths are equal to ℓ_{eg} then both quantities can be separately obtained.

For adequate accuracy in measurement the gaps must be made identical to within a few micrometers for resonators manufactured on high permittivity substrates.[1]

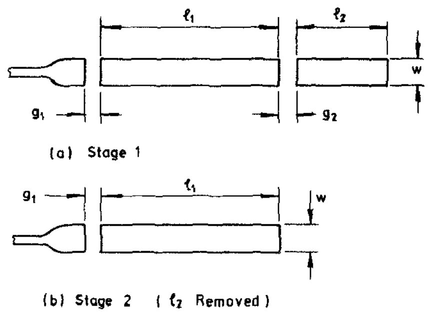


Figure 3. Two-stage resonator technique using quarter-wave gap-coupled line

Both previously described straight-resonator measurement methods have suffered from the drawback of requiring some prior knowledge of the end effect equivalent lengths. The 'symmetrical resonator' method (Fig 4) does not require this rather uncertain piece of design information.

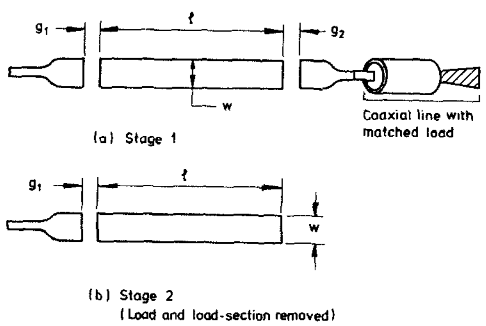


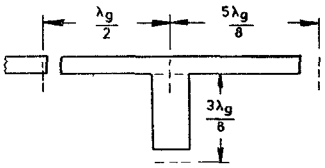
Figure 4. Symmetrical two-stage resonator technique

This arrangement also avoids the critically placed quarter-wave transforming line associated with the previous method. The gaps g_1 and g_2 are kept very nearly equal and set for loose coupling. If it is assumed that the effective microstrip permittivity is known from previous methods then the gap equivalent length is given by the expression:

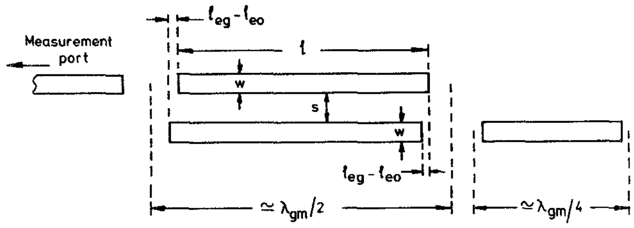
$$\ell_{eg} = \frac{nc}{4\ f_1\ \sqrt{\varepsilon_{eff}(f)}} - \ell \qquad \ldots(3)$$

The equivalent length ℓ_{eo} then follows as a function of f_2, but with ℓ_{eg} also subtracted.

Resonator methods may also be used to evaluate other discontinuity parameters and Fig 5(a) illustrates an arrangement for impedance transformation ratio (n) of a microstrip T-junction. The transformed impedance (in the vicinity of the junction) must be made finite and well-defined for measurement. To obtain this the intermediate 8th-wavelength sections of line are used since these yield input reactances equal to $-jZ_o$ ohms. In order to complete the measurements the secondary arm of the T-junction has to be removed by chemical etching, and the circuit has to be remeasured. The demands on the accuracy of initial circuit fabrication and subsequent etching are generally high.



(a)



(b)

Figure 5. T-junction measurement (a), and determination of phase velocities in parallel-coupled microstrips (b)

Reliable techniques for the determination of the Q-factor of a microwave resonator are well known and network-analyser-compatible swept-frequency techniques for measuring the Q-factor have been described in detail elsewhere.[1] These techniques are well suited to microstrip resonator measurements. The coupling coefficient must be kept within the range 0.15 to 0.75 for acceptable accuracy (that is to keep errors below 10%).[1]

Microstrip line losses may be determined by the comparison technique based upon two resonators with nominally identical microstrip parameters but different lengths, such as Fig 2(a). The Q-factors of two resonators may be combined to obtain the unloaded-Q and hence the line losses since the end loadings are similar for each resonator. There are situations where the equivalent admittance is required, at the open ends of the straight resonator (eg Fig 1(c)). Examples are microstrip antennas and the design of microstrip circuits for operation at high microwave frequencies. Having measured the external-Q (Q_e) we can easily find the end reflection coefficient ρ using the following expression:[1]

$$Q_e = \frac{\pi m \rho}{1 - \rho^2} \qquad \ldots(4)$$

Formulas are then available for the equivalent end susceptance and conductance.[1]

A summary of the resonator techniques is presented as Table 1.

Table 1

Figure Nos and Resonator type: / Uses and Features	Resonator Methods: 1(a) Ring	1(b) Side-Coupled	2 Paired
Dispersion	0		0
End-effects			
Q-factor	0	0	0
Need for circuit removal			0
A *priori* data required		●	●
Use of separate substrate			0
No of measurements obtainable	0	0	0

	3 With $\frac{1}{4}\lambda_g$ section	4 Symmetrical
Dispersion	0	0
End-effects	0	0
Q-factor	0	0
Need for circuit removal	●	●
A *priori* data required	●	0
Use of separate substrate		
No of measurements obtainable	0	0

0 signifies special suitability or advantages

● signifies lack of suitability or drawbacks

In many MIC filters and couplers *parallel-coupled microstrips* are required. The most important parameters are the even- and odd-mode characteristic impedances and phase velocities. Several techniques have been used for the measurement of these parameters but that due to Richings [6] appears to be the most extensive. The arrangement is shown in Fig 5(b) where two distinct resonances are observed and their frequencies are measured. End-effects are taken into account and phase velocities are readily evaluated.

The accuracy of parallel-coupled microstrip measurement techniques is subject to limitations of a nature similar to those for single microstrips.

Time Domain Methods

Time domain reflectometry (TDR) techniques may sometimes be used in a complementary role for microstrip measurements. In TDR the input reflection coefficient for the network under study is displayed as a function of time. Characteristic impedance may quickly be determined, at least to first-order accuracy, and variations along the microstrip line can be evaluated and approximately located. Some interesting work has been carried out using TDR and TDT (Time Domain Transmission) to evaluate discontinuities. [7] A 50-ohm microstrip line on an alumina substrate yields a delay of approximately 8.6 ps/mm. Using this value it is found that a TDR with a system pulse rise time of about 30 ps can only resolve distance separations of approximately 2 mm at best, along 50-ohm microstrip lines. In practice the finite rise time of the pulse ramp, and the effects of multiple reflections and noise complicate the displayed waveforms considerably and limit the resolution.

Conclusions

In this paper a variety of microstrip measurement techniques have been reviewed. Some aspects are probably not at all well known, in particular the interpretation of the external Q-factor and the use of the 'symmetrical resonator'.

One of the principal limitations affecting the choice of a microstrip measurement technique is the fundamental inability to realize effective broadband short circuits at high frequency in this medium. This fact may not be very widely appreciated since some workers still report measured results using 'short circuit' conductor plates to define a resonator.[8] As a consequence the results will not be very accurate, especially since test frequencies as high as 50 GHz are often employed.

Further work is required to characterise microstrip at higher frequencies, almost certainly to at least 90 GHz, and on some important substrates such as quartz and gallium arsenide. Anticipated problems include the complex behaviour of discontinuities which require full network analyser and synthesised source facilities at frequencies well above 18 GHz.

Another significant aspect of microstrip, the measurement of the frequency dependence of the characteristic impedance, receives detailed attention in a separate paper being presented at this symposium. [9]

References

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Acknowledgement

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