A Broad-Band Model for a Coaxial-to-Stripline Transition

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***Abstract*—This paper investigates a significant source of error encountered when characterizing a circuit element embedded in a stripline circuit at microwave frequencies. The errors introduced by the coaxial-to-stripline transition are examined and a frequency-independent model is developed for the transition over the band 2–12 GHz. The usefulness of the model is demonstrated experimentally by measuring a known load consisting of a short length of high-impedance line terminated in a short circuit.**

I. Introduction

THE CHARACTERIZATION of a circuit element is more difficult at microwave frequencies than at lower frequencies because it is often not possible to take measurements at the plane of the element itself. Instead measurements are made at, and referred to, some reference plane physically removed from the device. The device is then said to be embedded in the intervening structure [1]. If the embedded element is a two-terminal device, such as a detector diode, then the embedding structure can be regarded as a two-port network terminated in the device, with the measurement plane at the input as shown in Fig. 1.

To characterize the device, it is first necessary to determine the characteristics of the embedding network.

In a triplate stripline circuit the embedding network usually consists of a coaxial launcher and length of stripline which is terminated in the element to be characterized. Since the launcher and stripline can be regarded as lossless, the most significant source of measurement error comes from the transition between the launcher and the stripline [2].

The problem of the transition in stripline has received little attention, but a considerable amount of work has been done on the equivalent problem in microstrip [3]–[6].

The purpose of this paper is to investigate the nature of the errors introduced by the transition to stripline and to derive a broad-band model for it which will enable an embedded device to be accurately characterized.

II. The Stripline Launcher

The material used for the stripline circuits was aluminium backed Polyguide manufactured by Electronized Chemicals Corporation. The Polyguide was 0.159 cm thick and clad on one side by the same thickness of aluminium. The other side was clad with 35.6-μm thick copper. The coaxial launcher was an AMERICON SMA type, number 2070-1403.

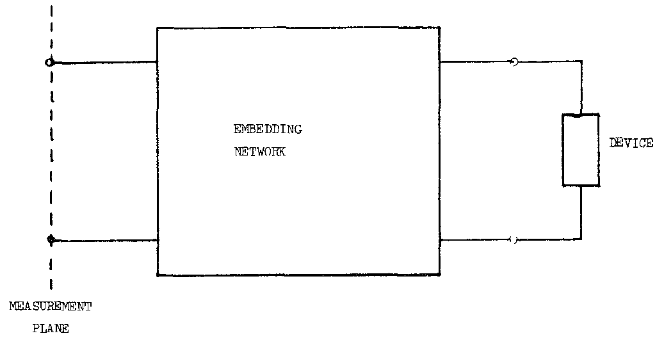

Fig. 1. Schematic representation of a two-terminal device embedded in stripline.

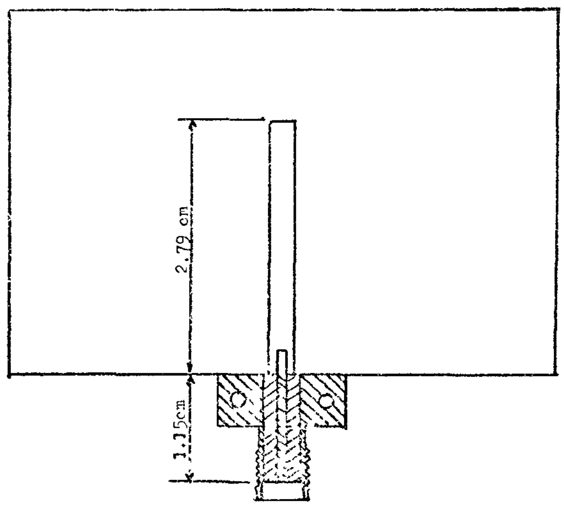

Fig. 2. Cross section of the SMA launcher and stripline assembly.

Contact between the launcher and stripline is effected by a tab extending from the launcher center conductor, and held in contact with the stripline by the pressure between the two pieces of Polyguide. The tab measures approximately 2.5 mm long by 1.3 mm wide and 0.1 mm thick.

III. Measurement Technique

To characterize the transition, a stripline circuit was made consisting of a straight 50-Ω line 2.79 cm long. The dielectric constant of Polyguide is nominally 2.32 and can reasonably be regarded as being frequency independent so

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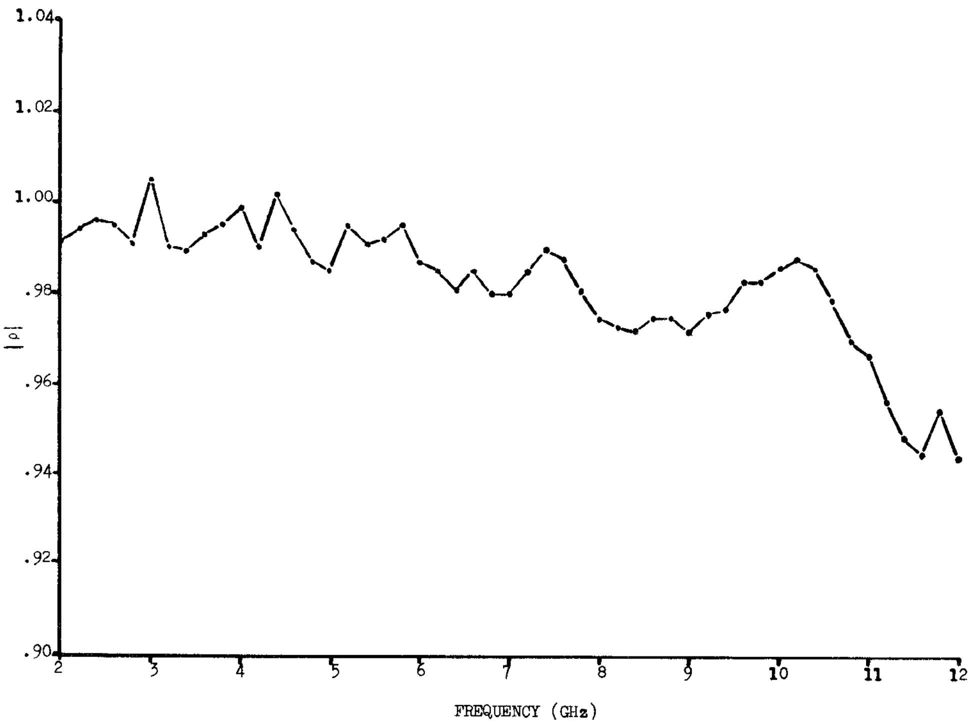

Fig. 3. Variation of amplitude of reflection coefficient with frequency of open-circuit stripline circuit.

the electrical length of the line was taken to be 4.25 cm. To this length must be added an extra electrical length of 0.05 cm due to the open-circuit end effect [7] of the stripline. The physical length of the launcher between the junction with the end of the stripline and the coaxial interface at the other end was measured to be 1.15 cm. The dielectric material used in the launcher is PTFE, the dielectric constant of which was taken to be 2.1, making the electrical length of the launcher 1.67 cm.

The overall theoretical electrical length of the circuit, then, is 5.97 cm. Fig. 2 shows a cross section of the launcher and stripline circuit. If the characteristic impedances of the launcher and stripline are both 50 Ω and can be regarded as perfect uniform transmission lines then all that is necessary to characterize a load terminating the stripline is to correct the reflection coefficient measured at the coaxial interface for the phaselength of the lines. In practice the transition from coaxial to stripline introduces an additional line, the phaselength of which is a function of frequency and the terminating load.

IV. Characterization of Stripline Circuit

The assembled stripline and launcher were characterized by measuring the magnitude and phase of the reflection coefficient at the coaxial interface of the launcher on a Hewlett-Packard automatic network analyzer over the band 2–12 GHz in 200-MHz steps, applying correction for network analyzer errors up to and including the APC-7 SMA adaptor. The correction procedure yields a reflection coefficient measurement accuracy of ± 0.05 dB in magnitude and $\pm 2°$ in phase. From these measurements a quantity termed the excess phase was calculated by subtracting the theoretical reflection coefficient phase due to an open-circuit perfect uniform transmission line of electrical length 5.97 cm from the measured phase. The measured reflection coefficient magnitude and excess phase characteristics are shown in Figs. 3 and 4, respectively.

Fig. 3 shows that the circuit, as expected, is substantially lossless, since $|\rho|$ is greater than 0.98 up to 7.8 GHz and only falls to 0.94 at 12 GHz. Fig. 4, however, shows that there is an excess phase characteristic with a periodic amplitude variation and a negative mean slope. The error introduced into phase measurements varies between 1° at 2 GHz and over 40° at 12 GHz. Errors of this magnitude are considered to be excessive and should be corrected when characterizing an embedded circuit element.

The negative mean slope corresponds to a line of electrical length of about 0.08 cm and indicates that the overall electrical length of the structure is slightly greater than the previously calculated value, while the periodic amplitude variations indicate the existence of a perturbing discontinuity within the structure. Since the launcher and stripline conductors are both uniform lines, this discontinuity must be the transition between them. It is also probable that the discontinuity is the cause of the apparent increase in the electrical length of the assembly (see Section V).

V. A Model for the Transition

The main discontinuity at the transition is caused by the launcher tab which, since it lies on top of the stripline conductor, increases its thickness and consequently reduces its characteristic impedance. The tab can, therefore, be represented as a short length of transmission line.

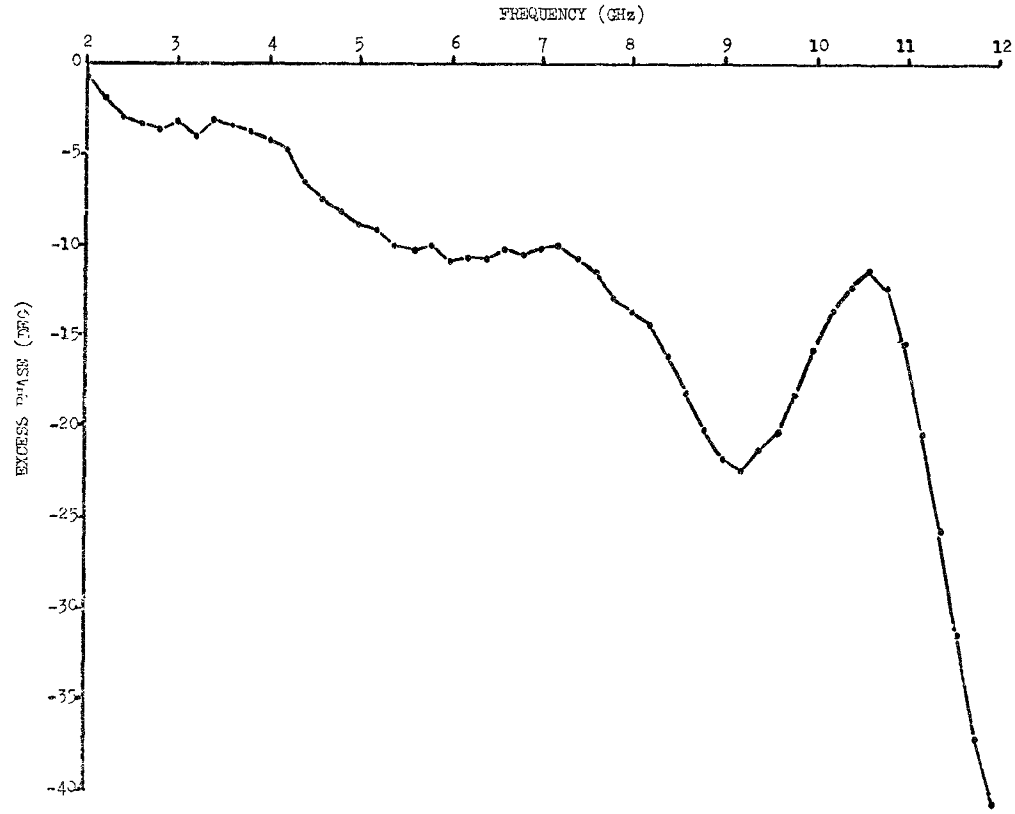

Fig. 4. Variation of phase of reflection coefficient with frequency of open-circuit stripline circuit.

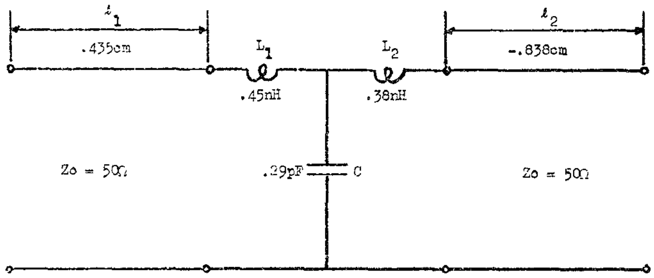

Fig. 5. A model for the coaxial-to-stripline transition.

The transition was initially considered as a lumped-element T-network approximation to a short transmission line. Ajose *et al.* [3] have developed a similar model for a transition to microstrip, but to obtain satisfactory agreement between the computed and measured excess phase characteristic for stripline the model has to be modified.

It has been assumed that the T-network was situated at the physical transition between the launcher and the end of the stripline. As the tab has a finite length it was decided to allow the position of the T-network to be varied by permitting the optimization program to vary the launcher and stripline lengths. The resulting transition model and the computed excess phase characteristics are shown in Figs. 5 and 6, respectively. The agreement between the measured and computed phase characteristics is satisfactory over the whole band with a maximum difference of only 3.5° at 2 GHz and considerably less above 3 GHz.

The model comprises a 50-Ω line of electrical length 0.44 cm, a 0.45-nH and a 0.38-nH inductor in the series arms of the T-network, and a 0.29-pF capacitor as the shunt arm. The T-network is terminated in a negative length of 50-Ω line −0.84-cm long. It is interesting to note that the T-network has not been simply moved away from the physical transition but has absorbed some of the line length. This is shown by the fact that the magnitudes of l_1 and l_2 are different and the sum of them is negative which means the overall electrical length of the launcher and stripline circuit is less than previously calculated. This conflicts with the deduction from the measured excess phase that the overall electrical length was greater than calculated. The apparent paradox has two possible explanations. The first is that the T-network itself must contribute an extra shift equivalent to an extra line length which is greater than the sum of l_1 and l_2. The second is the effect of spreading of current from the launcher tab where it makes contact with the much wider stripline.

The reproducibility of the phase measurements described above was examined by completely dismantling and reassembling the circuit, shown in Fig. 2, three times. The excess phase was measured every 1 GHz between 8 and 11 GHz. Table I shows the standard deviation of the excess phase at each measurement frequency, together with the mean. The excess phase deviation is sufficiently small to regard the measurement on one launcher as acceptably reproducible. The variation in launcher characteristics was examined by repeating the X-band measurements with five launchers.

The excess phase standard derivation is shown in the

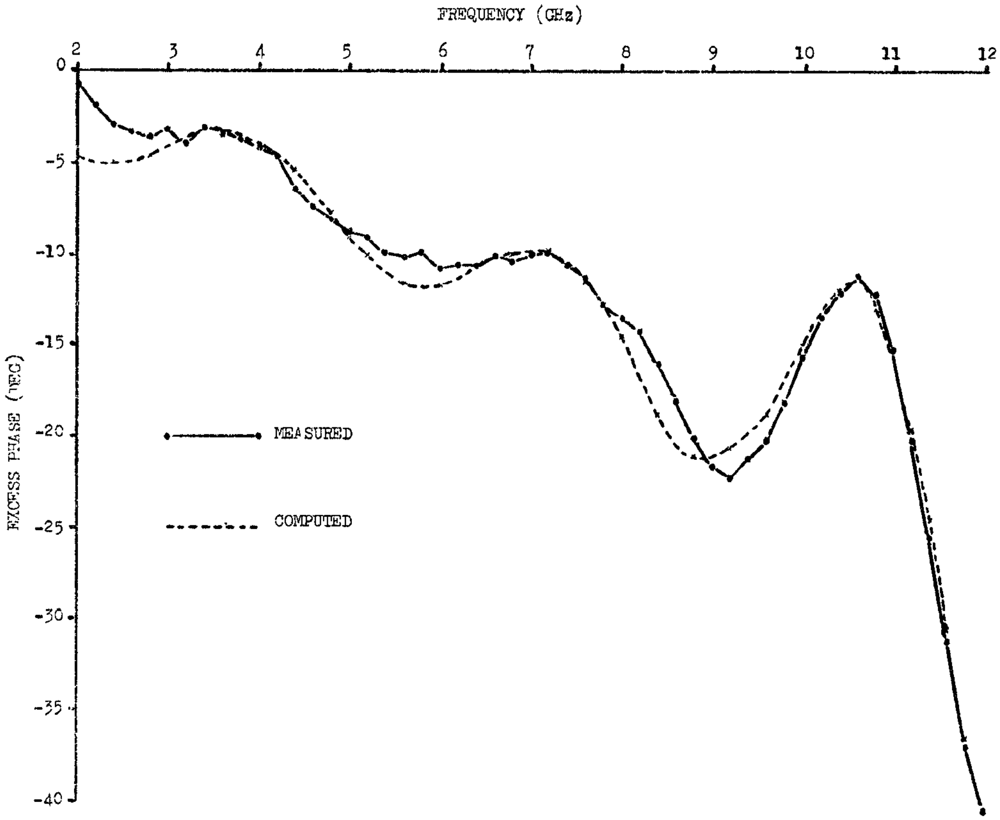

Fig. 6. Variation of reflection coefficient phase with frequency of transition model and measured phase.

TABLE I

Frequency(GHz)	8	9	10	11	12
Excess Phase Standard Deviation (°)	1.0	0.41	0.82	0.82	1.8
Mean Excess Phase (°)	9.7	13	26	20	21

TABLE II

Frequency(GHz)	8	9	10	11	12
Excess Phase Standard Deviation (°)	1.6	2.2	2.1	3.1	4.1
Mean Excess Phase (°)	9.6	15	26	19	23

Table II. This table shows that the standard deviation of the phase error is adequately small compared with the mean value at each frequency. A more accurate representation of the launcher equivalent circuit values requires individual characterization of each launcher. The remainder of the measurements described below were undertaken with the first launcher.

VI. Alternative Models for the Transition

A modification was made to the basic *T*-network model to include two additional shunt capacitances which represent the small discontinuities at each end of the tab. The computer optimization of this model gave a marginally better agreement between the measured and computed phase characteristics by reducing the deviation over the band from 1.24° to 0.94° which was not regarded as a sufficient improvement to justify use of the more complex model in experimental work. The modified model is shown in Fig. 7, from which it can be seen that the extra components have resulted in changes to all the other element values compared with the simple model. From these results, then, the circuit shown in Fig. 5 would appear to adequately represent the coaxial-to-stripline transition.

VII. Demonstration of Usefulness of Transition Model

The usefulness of the transition model can be demonstrated by measuring a load through the transition and correcting the measurement for the transition and comparing this with the uncorrected value and the theoretically expected value.

The load chosen was a series 0.2-mm diameter wire 2.9 mm long terminated in a demountable short circuit. A satisfactory demountable short circuit was constructed from two O.B.A. screws, one through each ground plane of the stripline mount. The two screws made contact with the lead of the element to be characterized in the plane of the stripline center conductor. This arrangement had the advantage that the series element could readily be changed. It is first necessary to characterize the short

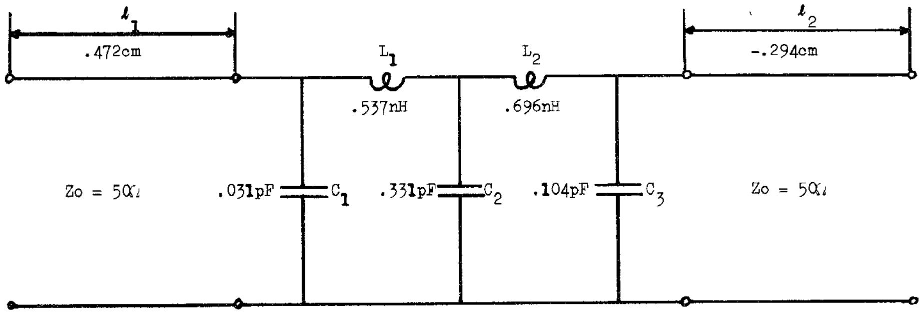

Fig. 7. An alternative model for the coaxial-to-stripline transition.

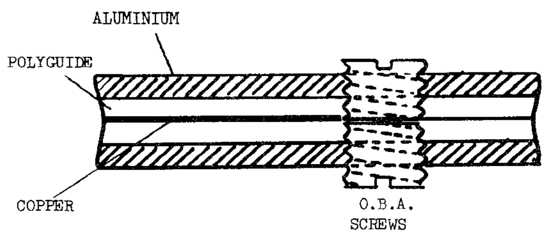

Fig. 8. Position and construction of short-circuit termination.

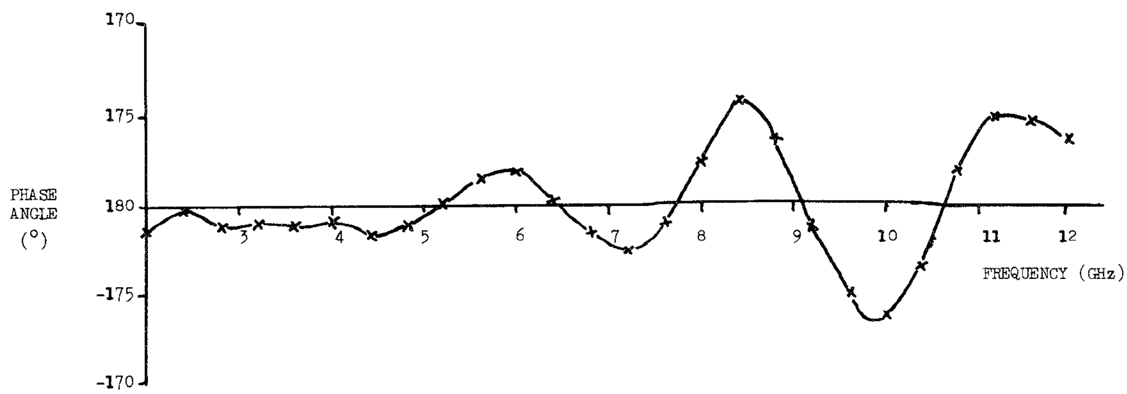

Fig. 9. Phase characteristic of demountable short circuit.

circuit and it was initially assumed that the transition model was correct, and reflection coefficient measurements were made on the mount (using the automatic network analyzer) after modifying it with the short-circuit screws positioned such that the end of the 50-Ω stripline abutted the outside diameter of the screws as shown in Fig. 8. Positive contact between the line and the screws was effected by a short length of wire. It was found that after deembedding the termination, a short circuit existed at a point 0.4 mm beyond the end of the 50-Ω stripline with a peak-to-peak phase spread of only $\pm 6°$ over the band 2–12 GHz, as shown in Fig. 9. Demounting and reassembly of the short circuit showed a reproducibility of $\pm 1°$ which was considered satisfactory.

For the measurement of the known load, the 50-Ω stripline was shortened by 2.9 mm and the 0.2-mm diameter wire connected between the end of the line and the short circuit as shown in Fig. 10(a).

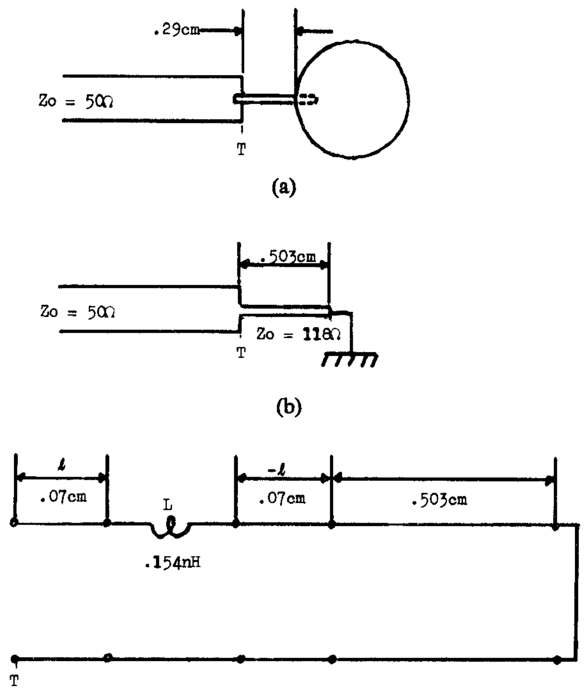

Fig. 10. The load used to confirm the transition model experimentally. (a) Physical construction. (b) Stripline representation. (c) Electrical model.

Now the wire has a characteristic impedance of 118 Ω given by the formula

$$Z_0 = \frac{138}{\sqrt{\epsilon_r}} \log_{10}\left[\frac{4b}{\pi d}\right] \Omega$$

where ϵ_r is the relative dielectric constant of Polyguide, b is the ground-plane spacing, and d is the diameter of the wire.

The circular cross section wire can be converted to an equivalent stripwidth of the same characteristic impedance from the graph given in [8]. Converting the line lengths to electrical lengths and including the extra length due to the position of the short-circuit phase, Fig. 10(a) can be redrawn as shown in Fig. 10(b).

According to Altschuler and Oliner [7], a step change in width of the center conductor of a strip transmission line can be accurately represented as a series inductance L, displaced from the physical discontinuity by a length l. The inductance is given by

$$\omega L = \frac{60\pi b}{\lambda} \log_e\left[\csc\left(\frac{\pi}{2} \cdot \frac{Z_{01}}{Z_{02}}\right)\right] \Omega$$

where Z_{01} and Z_{02} are the line impedances where Z_{01} is the lower impedance, and λ is the free-space wavelength. In this instance $Z_{01} = 50$ Ω, $Z_{02} = 118$ Ω, and the inductance is calculated to be

$$L = 0.154 \text{nH}.$$

The length is given by

$$l = \frac{b \log_e 2}{\pi}$$

where b is the ground-plane spacing, in this case 0.319 cm. The length is calculated to be

$$l = 0.07 \text{ cm}.$$

The known load can, therefore, be represented by the circuit shown in Fig. 10(c).

Fig. 11 shows the measured reflection coefficient phase of the deembedded load with, and without, correction for the transition, and the computed phase characteristics. The agreement between the corrected and computed phase characteristics is satisfactory over the entire band the effect of the correction has been to reduce the phase error at 10.5 GHz from 34° to 5°. At 11.4 GHz the improvement is only from 26° to 22°. Below 3 GHz there is very little difference between the corrected and uncorrected curves, but above this frequency, with the exception of around 11.4 GHz, a significant improvement in the agreement between the measured and theoretical curves is observed when the measurements are corrected for the effects of the transition.

VIII. Conclusions

The coaxial-to-stripline transition can produce significant errors into measurements on circuit elements embedded in stripline. A broad-band model has been developed which usefully describes the transition over the band 2–12 GHz. The application of the model has been demonstrated by measuring a known load and correcting the measurements for the effect of the transition to show that the phase error can be significantly reduced to give closer agreement with theory.

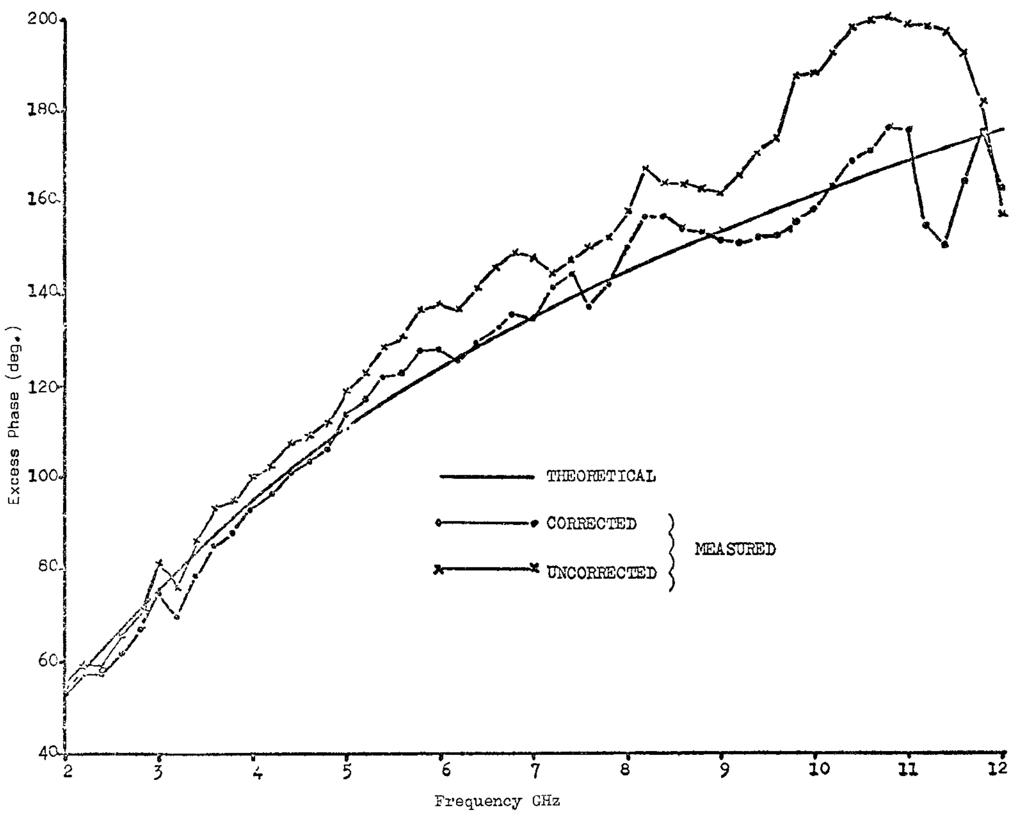

Fig. 11. Theoretical, corrected, and uncorrected measurements of variation of phase of load with frequency.

ACKNOWLEDGMENT

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