of the probe consists of nonconducting material. The field-perturbation effects of the initial model were not determined; however, thermographic [2, p. 705] tests for these effects will be performed for the next model to quantify any errors associated with the leads or the small amounts of highly conducting material used to make contact with the thermistor.

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# A Computer Program for the Direct Calibration of Two-Port Reflectometers for Automated Microwave Measurements

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***Abstract*—A short computer program is developed to solve explicitly a very useful reflectometer error model currently receiving high utilization in microwave measurement systems. The program architecture is designed to enhance its utilization as a stand-alone subroutine operating in conjunction with a measurement program, or to replace iterative solution software in existing automated measurement systems.**

## INTRODUCTION

The history of automated network analyzers is well documented and their contribution to accurate and high-speed microwave measurements is nothing short of phenomenal. However, because of heavy capital equipment investments required to purchase a fully automated system, many manual systems are still being utilized. Over the last few years, the introduction of inexpensive minicomputers such as the HP-9830 and the availability of programmable sources and measurement instruments have made it possible to build "homemade" automated network analyzers [1] at a substantially lower cost than the completed system purchase price. Utilizing the minicomputer as a controller (talker) and some standard data transfer/interface bus schemes, listening instruments such as sources may be set to the desired driving frequencies, and listening/talking devices such as scanning digital voltmeters may perform measurements. Software routines may be utilized to process the data thus acquired.

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## PROGRAM DESCRIPTION

A very important attribute of these systems is that if the measurement system is properly modeled, measurements can be made upon sets of precision standards and the error model can be solved. Unknowns can then be measured and corrected for errors [2]. High degrees of accuracy are thus possible for measurements of unknown two-ports.

Since most minicomputers utilized for instrumentation purposes are limited in core size, program size and efficiency are of paramount importance. The currently available software for automatic network analyzer application [3] solves the error vectors of particular models by iterative methods because of quadratic coupling between a number of system equations derived from the normal calibration standards. Although the iterative algorithm converges quickly (2 or 3 tries), the process must be repeated at each frequency point. Therefore, it would be very desirable to utilize a direct solution for the system equations.

Recent work has shown [4], [5] that explicit solutions are possible for the commonly utilized error models. Rehnmark [5] solves explicitly the 10-error vector model called GPS3 in the Hewlett-Packard Software Manual [3], while [6] solves explicitly the 12-error model GPS2. The 10-error model, shown in signal flowgraph form in Fig. 1, is significant because it may be used to model a useful, commercially available, and remotely programmable reflectometer (HP8746). The explicit solution for the unknown scattering parameters imbedded in the signal flowgraph of Fig. 1 is as follows:

$$S_{11} = \frac{\left(\dfrac{M_O - e_{00}}{e_{01}e_{10}}\right)\left[1 + e_{22}\left(\dfrac{M_3 - e_{33}}{e_{32}e_{23}}\right)\right] - e_{22}\left(\dfrac{M_3 - e_{30}}{e_{32}e_{23}}\right)\left(\dfrac{M_O - e_{03}}{e_{01}e_{10}}\right)}{D} \tag{1}$$

$$S_{22} = \frac{\left(\dfrac{M_3' - e_{33}}{e_{23}e_{32}}\right)\left[1 + e_{11}\left(\dfrac{M_O - e_{00}}{e_{01}e_{10}}\right)\right] - e_{11}\left(\dfrac{M_3 - e_{30}}{e_{23}e_{32}}\right)\left(\dfrac{M_O' - e_{03}}{e_{01}e_{10}}\right)}{D} \tag{2}$$

$$S_{21} = \frac{\left(\dfrac{M_3 - e_{30}}{e_{32}e_{10}}\right)}{D} \tag{3}$$

$$S_{12} = \frac{\left(\dfrac{M'_O - e_{03}}{e_{01}e_{23}}\right)}{D} \tag{4}$$

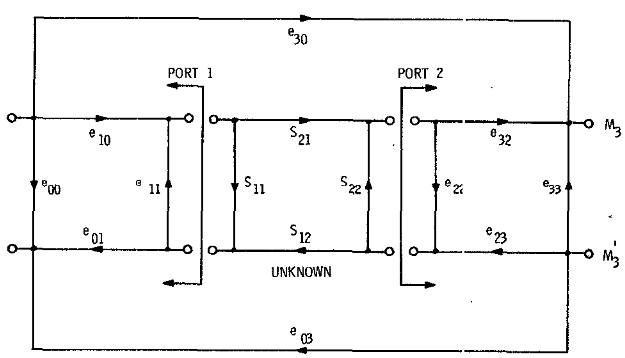


Fig. 1. 10-error model.

where

$$D = \left[1 + e_{11}\left(\frac{M_O - e_{00}}{e_{01}e_{10}}\right)\right]\left[1 + e_{22}\left(\frac{M'_3 - e_{33}}{e_{23}e_{32}}\right)\right] - e_{11}e_{22}K$$

where

$$K = \left(\frac{M'_O - e_{03}}{e_{01}e_{10}}\right)\left(\frac{M_3 - e_{30}}{e_{23}e_{32}}\right).$$

Note that solutions of $e_{01}$, $e_{10}$, $e_{23}$, or $e_{32}$ are not required.

The second subscript letter, if present with the indicated measurement, denotes the standard measured (i.e., $S$ = short, $O$ = open and $T$ = through line). Deletion of the second subscript denotes the indicated measurement is performed on the unknown.

The error vectors are defined by the following measurements:

$$e_{00} = M_{OM} \tag{5}$$

$$e_{11} = \frac{M_{OS} + M_{OO} - 2M_{OM}}{M_{OO} - M_{OS}} \tag{6}$$

$$e_{01}e_{10} = \frac{2(M_{OS} - M_{OM})(M_{OO} - M_{OM})}{(M_{OS} - M_{OO})} \tag{7}$$

$$e_{33} = M'_{3M} \tag{8}$$

$$e_{22} = \frac{M'_{3S} + M'_{3O} - 2M'_{3M}}{M_{3O}' - M_{3S}'} \tag{9}$$

$$e_{23}e_{32} = \frac{2(M'_{3S} - M'_{3M})(M'_{3O} - M'_{3M})}{M_{3S}' - M_{3O}'} \tag{10}$$

$$e_{30} = M_{3M} \tag{11}$$

$$e_{03} = M'_{OM} \tag{12}$$

$$e_{33}e_{01} = (M'_{OT} - e_{03})(1 - e_{11}e_{22}) \tag{13}$$

$$e_{10}e_{32} = (M_{3T} - e_{3O})(1 - e_{11}e_{22}). \tag{14}$$

## PROGRAM ARCHITECTURE

Carefully examining the relationships in Table I leads to the conclusion that by measuring the first ten parameters in Table I, the flowgraph of Fig. 1 can be solved. However, other operational parameters are required, i.e., offset voltage and the value of attenuation in the test channel. The former is simply the value of dc voltage at the final amplifier with no drive signal present. The latter is the amount of attenuation in the test channel such that maximum resolution at the test signal D/A converter is always obtained. A 20–30-dB return loss differential may be obtained easily between a good short and load. Hence, an attenuator must be utilized to maintain a previously determined level in the output. Therefore, every measurement outlined in Table I will require an associated value of test channel gain as data. Further, this data will be required at every frequency point. The flowchart in Fig. 2 illustrates a program to solve the flowgraph of Fig. 1, which may be utilized as a modular subroutine. Data is acquired through a common data block.

## CONCLUSION

A basic language program has been written to implement the algorithm offered in Fig. 2 using the solution of [1]–[4] for the HP9830 minicomputer. The program size is approximately equal in size to the iterative solution. As a result, core requirements do not decrease. Given the measurements required (measurements of standards and unknowns) the program will yield the scattering parameters of the unknown two-port. The value of the solution agrees excellently with the iterative solution offered commercially. However, this solution is approximately two to three times faster, and since this process is repeated for every frequency point, the net advantage is obvious. Also, since it is modular in design, minimum

TABLE I

DESCRIPTION OF MEASUREMENTS REQUIRED TO SOLVE FLOWGRAPH OF FIG. 1 FOR UNKNOWN SCATTERING PARAMETERS

| No. | Measurement | Standard | Measure |
|---|---|---|---|
| 1 | $M_{OM}$ | Load on Port One | $M_O$ |
| 2 | $M_{3M}$ | | $M_3$ |
| 3 | $M'_{3M}$ | Load on Port Two | $M'_3$ |
| 4 | $M'_{OM}$ | | $M'_O$ |
| 5 | $M_{OS}$ | Short on Port One | $M_O$ |
| 6 | $M_{OO}$ | Open on Port One | $M_O$ |
| 7 | $M'_{3S}$ | Short on Port Two | $M'_3$ |
| 8 | $M'_{3O}$ | Open on Port Two | $M'_3$ |
| 9 | $M'_{OT}$ | Connect Port One to Port Two with through line | $M'_O$ |
| 10 | $M_{3T}$ | | $M_3$ |
| 11 | $M_O$ | UNKNOWN | $M_O$ |
| 12 | $M_3$ | DEVICE | $M_3$ |
| 13 | $M'_O$ | PLACED | $M'_O$ |
| 14 | $M'_3$ | BETWEEN | $M'_3$ |
| | | Port one and two | |

*Note:* For unprimed measurements the driving generator is connected to the left-hand port. For primed measurements the generator is connected to the right-hand port.

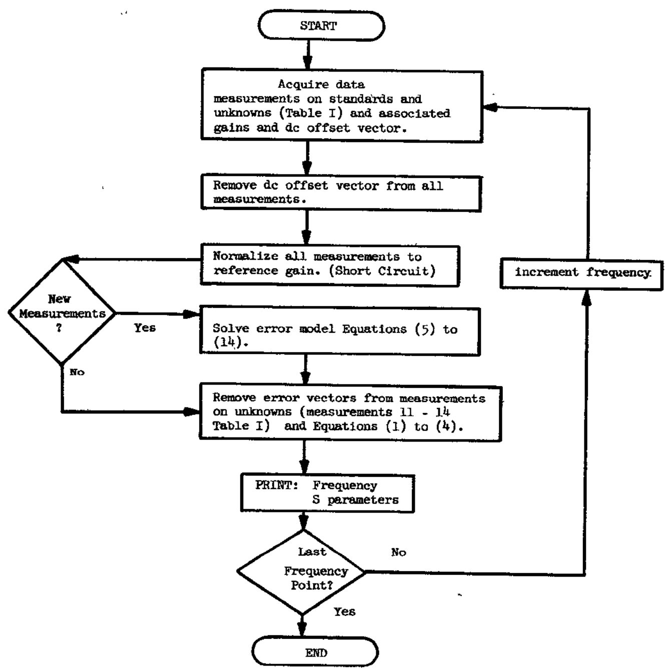


Fig. 2. Flow chart for computer program to solve error model of Fig. 1.

time is spent in interfacing solutions with existing software. A listing of this subroutine in BASIC may be obtained from the author of this paper.

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# A Coaxial to Microstrip Transition

E. H. ENGLAND

***Abstract*—A method of obtaining an improved transition, from 0.141-in (3.55-mm) semirigid coaxial to microstrip is described. Further improvements by means of compensation include two fixed types having a reflection coefficient less than 0.005 and an adjustable form capable of producing a "transparent" transition.**

A variety of commercial transitions (launchers) are available, normally in combination with a connector. When used in conjunction with 0.5-mm-thick substrates, these transitions typically have a reflection coefficient ($\Gamma$) of 0.03 or more, including the associated connector. Often this is not good enough for precision work.

The transition tab discontinuity was first described by Caulton *et al.* [1] as being capacitive; this is confirmed by time domain reflectometry (TDR) measurements. At 28-pS overall system rise time, TDR also reveals the capacitance to be small in value and discrete. Its effective position appears to be near the tip of the tab and is thought to be partly due to field concentration at any sharp corners on the tab. With this in mind, such corners were rounded off and the length and shape of the tab optimized, experimentally, to produce a minimum $\Gamma$.

The transition thus developed is shown in Fig. 1; it is formed directly on the semirigid coaxial thereby avoiding the discontinuity presented by a connector. The tab is shaped, using a needle file, in a simple jig. When used in conjunction with the jig shown in Fig. 2 to launch into 50-$\Omega$ microstrip on 0.5-mm sapphire ($C$ axis perpendicular to plane of substrate), this tab gives $\Gamma < 0.01$.

Thought was given to reducing the remaining capacitive effect of the tab described still further. Inductive compensation would give a band-limited solution. Instead, the line capacitance was reduced at the appropriate position by means of a hole in the ground plane (GP) below the tab, as in Fig. 3, placed above a larger hole in the supporting baseplate.

Alternatively, a hole shaped as in Fig. 4 can be used with a solid baseplate if the GP thickness is increased to 40 $\mu$m. Either of these methods reduces $\Gamma$ to less than 0.005.

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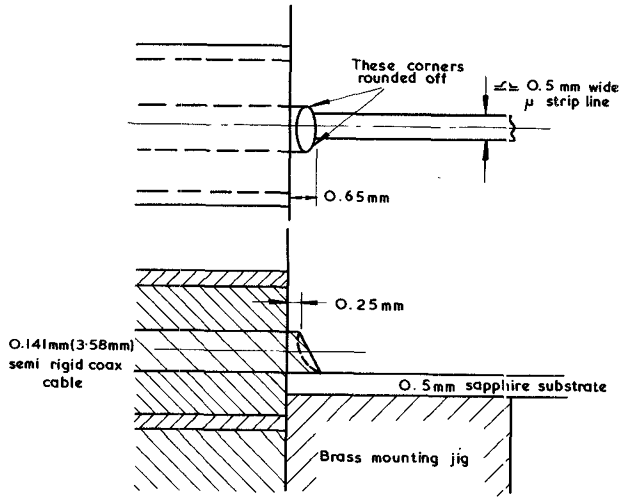


Fig. 1. Details of transition tab.

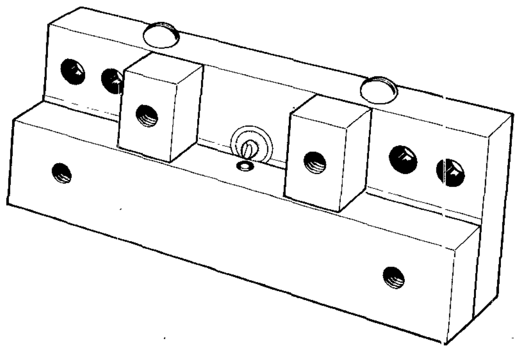

Fig. 2. Transition and substrate holding jig.

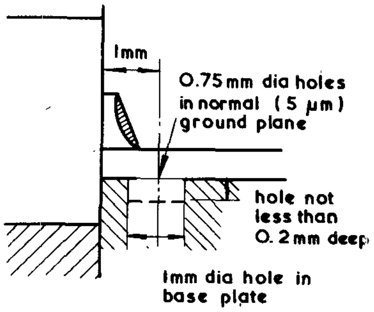


Fig. 3. Fixed compensation hole in normal ground plane.

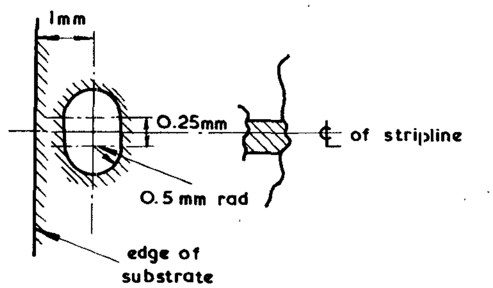


Fig. 4. Fixed compensation hole in thickened ground plane.