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COMPUTER-CORRECTED REFLECTOMETER USING
THE HP-8410 AND AN APPLE II

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ABSTRACT

This report describes an implementation of the short-open-load method of calibrating and correcting errors in a complex reflectometer. The computations are simple enough to be performed with reasonable speed by an 8-bit microcomputer, allowing swept-frequency measurements. This implementation uses a Hewlett-Packard 8410A/8411A/8414A/8746B network analyzer as the reflectometer; a specially built analog-to-digital interface; an Apple II Plus computer; and a voltage-controlled (YIG tuned) sweep oscillator. After calibration, the measurement speed is about 310 msec per frequency, including graphical display of the corrected reflection coefficient. At S-band, the setup achieves an effective directivity > 40 dB immediately after calibration (limited by the standard load), and this degrades to ~ 26 dB after several hours. The software provides an option to substitute a shorted delay line for the open circuit calibration standard. Programming is mostly in Applesoft Basic, with assembly language subroutines used for complex arithmetic and control of the interface.

I. Introduction

A complex reflectometer measures impedance at a point on a uniform transmission line by determining the complex reflection coefficient at another point along the line. Such an instrument can be built by inserting two directional

couplers on the line between a matched source and the unknown load, with one coupler arranged to sample the incident wave and the other the reflected wave, and connecting the coupler outputs to a complex ratio meter. Ideally, the source impedance will be exactly matched to the line; the couplers will have perfect directivity, identical coupling coefficient, and will not cause reflections on the line; and the ratio meter will be linear and well matched. In practice it is difficult to achieve these conditions to high accuracy, especially if the instrument is to be operated over a wide range of frequencies.

Fortunately, it is now widely recognized that it is not worthwhile to try to improve the accuracy of the microwave components to any great extent, provided that a computer is available. If the components have moderate accuracy (undesired signals at least 10 dB below desired ones), the errors can be determined and stored for each frequency of interest, and measurements of any unknown at the same frequencies can then be corrected.

There are three basic sources of error, each of which is complex; these are imperfect source match, imperfect directivity, and unknown complex gain in the ratio meter. Most other sources of error are equivalent to one of these (e.g., d.c. offset in the ratio meter is like imperfect directivity). To calibrate these errors, at least three different, known reflections must be connected in place of the unknown. As we shall see in the next section, a short circuit, open circuit, and matched load are good choices.

At the NRAO central development lab, we frequently need to measure impedances in the 1 to 23 GHz range. We have an old HP8410A network analyzer with 8746B S-parameter test set, which nominally operate up to 12.4 GHz. However, the accuracy degrades rapidly as the upper frequency limit is approached. The instrument contains adjustments for complex gain and the length of a reference

line, allowing compensation for some internal errors; but we often want to make swept measurements over a wide band (around one octave), and the optimum adjustment settings vary significantly with frequency. We were, therefore, motivated to investigate this method of computer corrections. In addition, we wanted to see whether the network analyzer might be pushed beyond its nominal 12.4 GHz limit.

In what follows, I first give the simple theory of the calibration and correction procedures; next, I discuss some practical considerations of general applicability; then, I describe the hardware and software of the NRAO implementation and give some test results; and finally, I discuss errors which are neglected in the model and mention improvements which could be made.

II. Theory

Figure 1a is a model of a complex reflectometer of the type described in the introduction. At the measurement plane, the incident and reflected wave amplitudes are a and b , respectively, and we are interested in determining the reflection coefficient $\Gamma = b/a$. The instrument attempts to measure the ratio b'/a' of the wave amplitudes from the coupler, and this ratio might be equal to b/a if it were not for the various errors mentioned earlier. The ratio meter's output is not exactly b'/a' , but rather $x = gb'/a' + g_0$ due to complex offset g_0 and gain g .

Figure 1b is another model of the same reflectometer [1,2]. Here an "ideal" reflectometer is cascaded with a linear two-port network called the "error box" whose parameters describe the non-ideal characteristics of the actual reflectometer. With a little thought the reader can convince himself that there exist parameters for the error box which will make Fig. 1a and Fig. 1b equivalent. The calibration

problem can then be visualized as determining these parameters; that is, specifying the relationships among the wave amplitudes a , b , a'' , b'' .

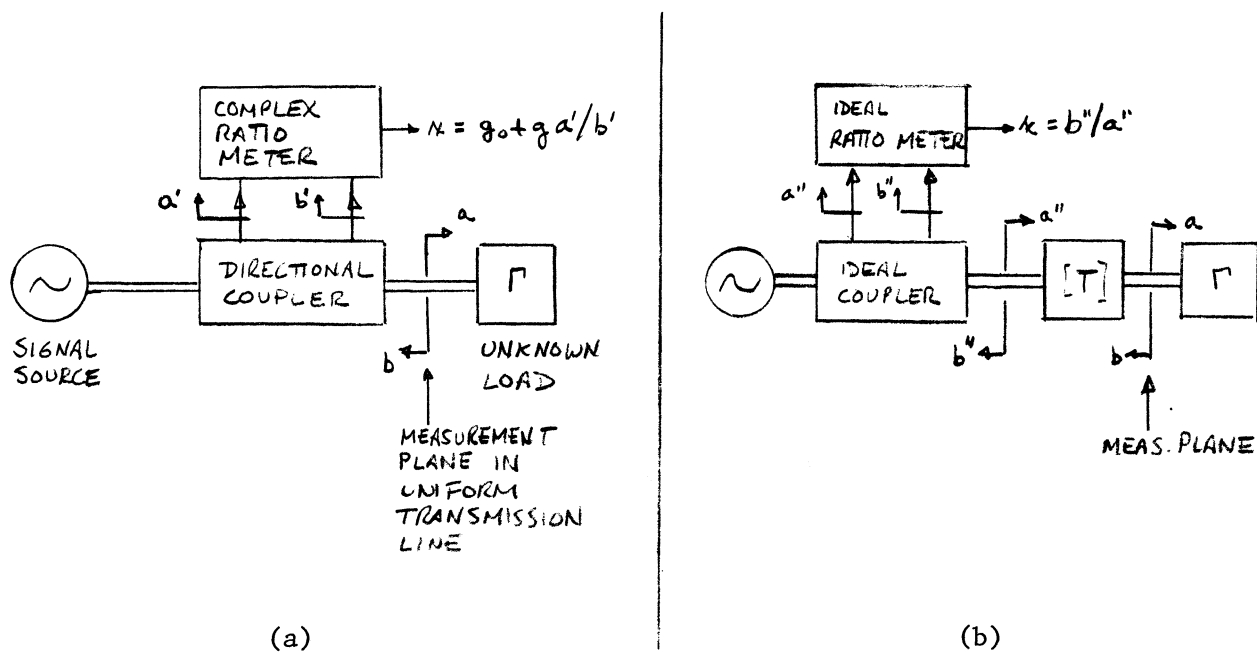


Fig. 1. Two equivalent models of a complex reflectometer.

For a general linear two-port,

$$a'' = T_{11}a + T_{12}b \quad (1)$$

$$b'' = T_{21}a + T_{22}b$$

where $[T_{ij}]$ is known as the wave cascading matrix. The ratio meter reading is then

$$x = \frac{a''}{b''} = \frac{T_{11}a + T_{12}b}{T_{21}a + T_{22}b} = \frac{T_{11} + T_{12}\Gamma}{T_{21} + T_{22}\Gamma} \quad (2)$$

Solving (2) for Γ , we also have

$$\Gamma = \frac{xT_{11} - T_{21}}{T_{22} - xT_{12}} \quad (3)$$

Notice that we do not need to find all four parameters $\{T_{ij}\}$, because (2) and (3) could be divided through by any one of them. For computational accuracy, we should avoid dividing by quantities near zero; if we suppose that our instrument approximates an ideal reflectometer, so that $T_{12}, T_{21} \ll T_{11}, T_{22}$, then we could divide by T_{11} or T_{22} . Choosing T_{11} gives

$$x = \frac{C + D\Gamma}{1 + B\Gamma} \quad (4)$$

$$\Gamma = \frac{x - C}{D - xB} \quad (5)$$

where $B = T_{12}/T_{11}$, etc.

Now, if three different, known loads $\Gamma_1, \Gamma_2, \Gamma_3$ are connected at the measurement plane yielding measurements x_1, x_2, x_3 , then (4) gives a system of linear equations in B, C, D:

$$-B\Gamma_k x_k + C + D\Gamma_k = x_k, \quad k = 1, 2, 3 \quad (6)$$

the solution to which can be written

$$\begin{bmatrix} B \\ C \\ D \end{bmatrix} = \begin{bmatrix} -\Gamma_1 x_1 & 1 & \Gamma_1 \\ -\Gamma_2 x_2 & 1 & \Gamma_2 \\ -\Gamma_3 x_3 & 1 & \Gamma_3 \end{bmatrix}^{-1} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} \quad (7)$$

Choice of Calibration Standards

To avoid singularity in the matrix in (7), the Γ_k 's should be as different as possible. Moreover, the calculations are simplified (for swept measurements)

if one or more of the standards is as frequency-independent as possible. An obvious choice is $\Gamma_1 = -1$, a short circuit; since this can almost always be implemented with great accuracy, we will assume it from now on. In principle the best choice for the other standards is then $\Gamma_2 = \Gamma_3^* = \cos 60^\circ + j \sin 60^\circ$, but this is hard to implement over a wide bandwidth. Moreover, there is an advantage in making one of the standards a good match, say $\Gamma_3 \approx 0$; then parameter C, which represents directivity error and d.c. offset, is determined directly from x_3 [see (4) or (6)]. If the system were perfectly linear, C could be determined just as accurately from (7) with $|\Gamma_k| = 1$ for all k, but small non-linearities could then lead to errors in C which would be significant when measuring low-reflection unknowns. This is avoidable if a good match is available as a standard; we now make this assumption, setting $\Gamma_3 = 0$. The best choice for the remaining standard is then $\Gamma_2 = +1$, an open circuit. In coax, this can be closely approximated by a shielded connector for frequencies well below the cutoff of the first non-TEM mode. But sometimes it is more convenient to use a delayed short, where the delay is a length of line around one-quarter wavelength long at the center of the measurement band. In waveguide, a broadband "open" cannot be made, and the latter method is usually adopted. For the present, we will leave Γ_2 unspecified.

Setting $\Gamma_1 = -1$ and $\Gamma_3 = 0$ in (7) and carrying out the matrix operations gives

$$\begin{bmatrix} B \\ C \\ D \end{bmatrix} = \begin{bmatrix} x_1 & 1 & -1 \\ -\Gamma_2 x_2 & 1 & \Gamma_2 \\ 0 & 1 & 0 \end{bmatrix}^{-1} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix}$$

$$= \begin{bmatrix} \frac{(x_3 - x_1)\Gamma_2 + x_3 - x_2}{(x_2 - x_1)\Gamma_2} \\ x_3 \\ x_3 + x_1 \frac{(x_3 - x_2)(\Gamma_2 + 1)}{(x_2 - x_1)\Gamma_2} \end{bmatrix} \quad (8)$$

Once the instrument is calibrated at any frequency using (8), measurements of unknowns at the same frequency can be corrected using (5).

III. Some Practical Considerations

Frequency stability. The signal source must have repeatable tuning, so that the calibration measurements and the test measurements, which are separated in time, are made at the same frequencies. Therefore, a digitally synthesized signal locked to a stable reference oscillator is desirable. But a fairly stable analog-tuned source may be adequate if the reflectometer has reasonably good directivity and if its internal line lengths are adjusted to provide equal-delay paths to the two inputs of the ratio meter. In this case, the calibration parameters will not vary very rapidly with frequency, so that small drifts can be tolerated. Nevertheless, if the entire setup is many wavelengths in size (as it usually is), then internal reflections will cause fine-scale structure in the calibration parameters such that the source stability may limit the measurement accuracy.

Phase lock of ratio meter. Microwave complex ratio meters normally operate by converting the input signals to a low I.F. using a local oscillator which is phase locked to one of them. The spectral purity of the source must therefore be adequate to allow accurate phase locking. This depends on the loop bandwidth, which in turn determines the speed of lock acquisition, and which might limit the speed of swept measurements.

Accuracy of standards. The short circuit can almost always be assumed to be perfect ($\Gamma \equiv -1$). If we let the short define the location of the reference plane, then no phase error occurs; if the contacts are well designed and reasonable materials are used (e.g., Cu, Ag, Au), then magnitude errors are negligible in a

50 ohm system. If a delayed short is used, its accuracy depends primarily on knowledge of the propagation constant and length of the delay line. If an open circuit is used, the equivalent terminating capacitance must usually be known; this capacitance varies weakly with frequency. Table I gives experimental data supplied by Hewlett-Packard on some shielded open circuits manufactured by them; some theory is also available [3,4]. The third standard, $\Gamma = 0$, is probably the most troublesome; a load with $|\Gamma| < 10^{-3}$ is much harder to make than a short with $1 - |\Gamma| < 10^{-3}$. This can be overcome with a sliding load, as discussed in Section VI, but this complicates the calibration and is not included in the present implementation.

Non-linearity. The models of Fig. 1 assume that the ratio meter is linear in complex amplitude. This assumption is violated not only if the signal suffers compression in the mixers and amplifiers, but also if the real and imaginary channels have unequal gain or are not exactly in quadrature. The latter effects can be corrected, as discussed in Section VI, but again this is not included in the present implementation.

TABLE I: Capacitance of Hewlett-Packard* Shielded Open Circuits

$$C = C_0 + C_1 f + C_2 f^2$$

H-P Model (cal. kit)	Connector Type	$\frac{C_0}{\text{pf}}$	$\frac{C_1}{\text{pf/GHz}}$	$\frac{C_2}{\text{pf/GHz}^2}$
85033A, opt. 001	SMA male	.049	2×10^{-5}	3.5×10^{-5}
	SMA male	.009	3.5×10^{-5}	2.8×10^{-5}
85032A	N male**	.11	2.5×10^{-5}	1.7×10^{-10}
	N female	.057	2×10^{-5}	3.6×10^{-5}
85033B	APC 3.5 male**	-.0097	2.15×10^{-3}	-3.19×10^{-5}
	APC 3.5 female**	.021	-7.62×10^{-4}	0
11866A	APC 7	.079	0	4×10^{-5}

* data supplied by Dick Jones of Richmond sales office.

** requires "extender pin."

IV. Implementation

Hardware. Figure 2 gives a block diagram of the setup used for 1-12 GHz measurements. The Avantek YIG-tuned oscillators appear to provide adequate spectral purity and stability, although so far most of our testing has been below 4 GHz. The frequency counter is needed only to allow absolute calibration of the sweep oscillator. This is important if a calibration standard has a strong frequency dependence which must be calculated, as is the case for a delayed short. If a shielded open is used, precise frequency calibration may not be necessary.

We used an H-P S-Parameter Test Set (8746B) as a broad band, dual directional coupler; it also provides a convenient variable-length reference line. A pair of ordinary directional couplers could easily be substituted. An H-P Network Analyzer (8411A/8410A/8414A) is used as the complex ratio meter, with analog outputs (± 1.5 volt nominal) proportional to the real and imaginary parts of the voltage ratio.

An ADIOS module [5] provides the interface to an Apple II+ computer via three analog connections: a control voltage to the sweep oscillator, and the "horizontal" (real part) and "vertical" (imaginary part) outputs from the network analyzer. The ADIOS inputs and outputs have a range of $\pm 10V$. Thus, any voltage-tuned sweep generator with a control voltage in this range, and any ratio meter with outputs in this range, could be substituted for those used here.

Software. A program has been written to allow acquisition and storage of calibration data over a range of frequencies (up to 128 points at present); and to allow subsequent measurement, correction, and display of unknown reflection

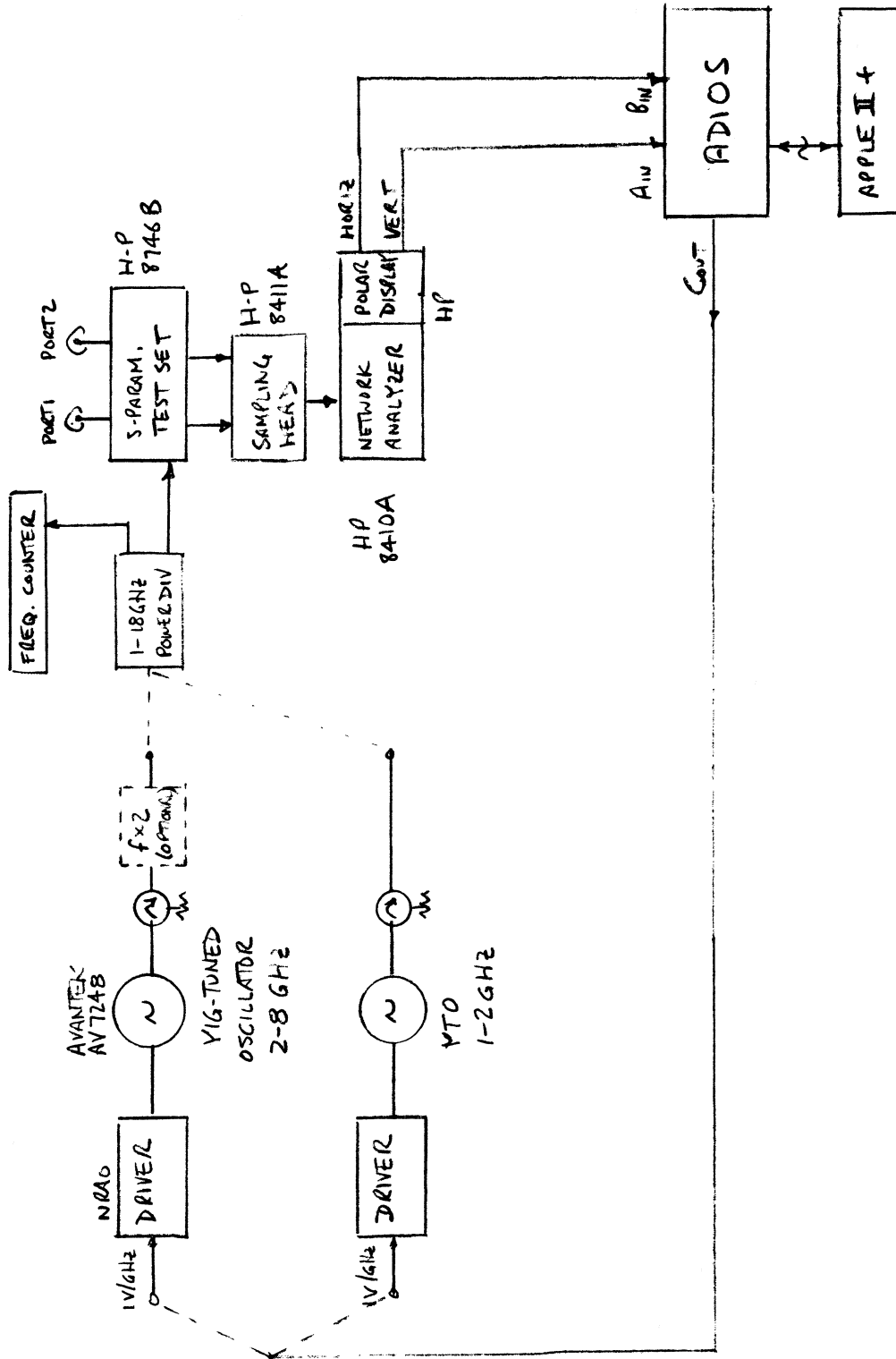


Fig. 2. Block diagram of NRAO implementation.

coefficients at the same frequencies. Features of the program include:

1. Both short-open-load and short-delayed short-load calibrations are supported. In the latter case, the delay line is assumed uniform, with characteristic impedance equal to the reference impedance (50 ohms in our system); but it may be lossy and of any (known) electrical length. For the open circuit, the user specifies the terminating capacitance as a second-degree polynomial in frequency (cf. Table I).
2. The voltage-frequency characteristic of the sweep oscillator can have any scaling and offset, subject to the limits of the ADIOS D to A converter (-10V to +10V range, 0.3 mV resolution). The program uses straight-line interpolation between user-specified limits.
3. Corrected measurements are listed in a table and plotted on the Apple's high-resolution graphics screen. The screen maps the complex Γ -plane, centered at the origin, out to an adjustable radius (default radius = 1). Overlays are available with circles of constant impedance (standard Smith chart), constant admittance, and constant reflection coefficient (polar coordinates).
4. When measuring an unknown, the program steps through the frequency range either once or repeatedly. With repeated sweeps, the result of the previous measurement at a given frequency is erased from the plot before the current result is plotted; this allows the user to see only the latest sweep, and thus to follow the effect of adjustments of the device being measured. In such an interactive application, the speed of measurement is important, as discussed below.
5. Options include printout of the calibration data in a formatted table, and a hard-copy dump of the graphics screen.

The program is written mainly in Applesoft BASIC, but it relies on some subroutines from the NRAO Binary Library [6], which was written in an assembly language. These subroutines provide for control of the ADIOS module [7] and for complex arithmetic operations [8]. Nearly all of these could have been implemented in BASIC, but with a severe penalty in measurement speed (estimated factor of 4 to 6).

Detailed user instructions for this program are given in Appendix A.

V. Test Results

Measurement speed. After calibration, measurement and display of an unknown reflection coefficient takes about 310 msec per frequency. Separate tests of program segments allow a partial breakdown as follows:

Service ADIOS (data transfer, no processing)	9 msec
Service ADIOS and compute corrected Γ	130 msec
Service ADIOS, compute Γ , print and plot result	310 msec

The ADIOS module uses integrating A-to-D converters, and integration can be overlapped with processing. The program sets the integrating time to use nearly all of the 310 msec available.

In calibrating, the time is usually dominated by the need for the operator to change connections; the system takes 310 msec per frequency to measure each of the three standards, and then about 500 msec per frequency (estimated) to evaluate eqn. (8). The measurement time could be reduced by decreasing the integrating time, but this seems unnecessary.

Accuracy. Absolute accuracy is hard to assess because it depends on several interacting factors, including the quality of the signal source, the accuracy of

the calibration standards, and the repeatability of connectors. Nevertheless, a few tests have been performed which give some indication of the accuracy. It should first be mentioned that A-to-D quantization and computational roundoff errors can be safely ignored unless two of the calibration standards have nearly equal reflection coefficients; normally, quantization error is about -40 dB (300 msec integration) and roundoff error is about -90 dB (Apple floating point, 4-byte mantissa).

The following tests were done in the 2 to 4 GHz range, except as noted, after calibration at an SMA female connector (H-P SMA to APC7 adapter) using a Wiltron model 22S50 short/open standard and a Weinshel model M1408 SMA termination. The 22S50's short and open reference planes are offset so as to compensate for the open's capacitance, so zero capacitance was assumed in the program. The termination was measured on a scalar network analyzer and found to have $|\Gamma| < -38$ dB to 4 GHz, and < -30 dB to 12 GHz.

Table II lists the results of measuring the calibration standards a few minutes after calibration, but after each had been removed and reconnected. The maximum errors were 4.4×10^{-3} (-47 dB) for the low- Γ load and 1.1×10^{-2} (-39 dB) for the high- Γ loads. These errors are believed due to frequency drift of the source (~ 1 MHz) plus connector repeatability. Also shown in Table II are similar measurements made about 15 hours after calibration (overnight drift); by then, the frequency had drifted 2 to 4 MHz, but nearly all errors remained below 1% (-40 dB).

Figure 3 shows the results of measuring a sliding load and a sliding short at a fixed frequency of 3.0 GHz.

Figure 4(a) shows the variation of calibration data over a wide frequency range, 2 to 12 GHz, plotted every 0.5 GHz. There was no difficulty in locking

TABLE II: Re-Measurement of Calibration Standards

	<u>A few minutes after calibration</u>			<u>15 hrs. after calibration</u>			
	<u>f</u>	<u> Γ </u>	<u>phase, rad.</u>	<u>f</u>	<u> Γ </u>	<u>phase, rad.</u>	
<u>Load</u>	2	3.89E-3	1.4052	2	5.28E-3	.3281	
	2.5	4.17E-3	1.3851	2.2	6.20E-3	- .6926	
	3	3.77E-3	1.3028	2.4	3.13E3	.0666	
	3.5	3.73E-3	1.3192	2.6	4.67E-3	.7446	
	4	4.44E-3	1.2997	2.8	5.77E-3	.2187	
				3	4.55E-3	- .0699	
				3.2	3.03E-3	.2227	
				3.4	5.36E-3	.5824	
				3.6	6.17E-3	.1869	
				3.8	3.93E-3	.0601	
				4	3.67E-3	.5568	
	<u>Short</u>	2	.9929	3.1340	2	.9831	3.1239
		2.5	.9929	3.1364	2.2	.9844	3.1343
		3	.9892	3.1378	2.4	.9935	3.1347
3.5		.9924	3.1344	2.6	.9883	3.1342	
4		.9995	3.1344	2.8	.9938	3.1380	
				3	.9946	3.1421	
				3.2	.9967	3.1403	
				3.4	.9938	3.1394	
				3.6	.9938	3.1436	
				3.8	.9961	3.1469	
				4	.9981	3.1447	
<u>Open</u>		2	.9924	1.27E-3	2	.9939	5.30E-4
		2.5	.9953	3.45E-3	2.2	.9936	-1.79E-3
		3	.9969	2.96E-4	2.4	.9912	-1.06E-3
	3.5	.9923	1.46E-3	2.6	.9873	1.20E-3	
	4	1.003	1.10E-3	2.8	.9993	-4.70E-4	
				3	.9953	-2.09E-4	
				3.2	.9937	2.06E-3	
				3.4	.9953	5.16E-3	
				3.6	.9973	4.09E-3	
				3.8	.9950	4.53E-3	

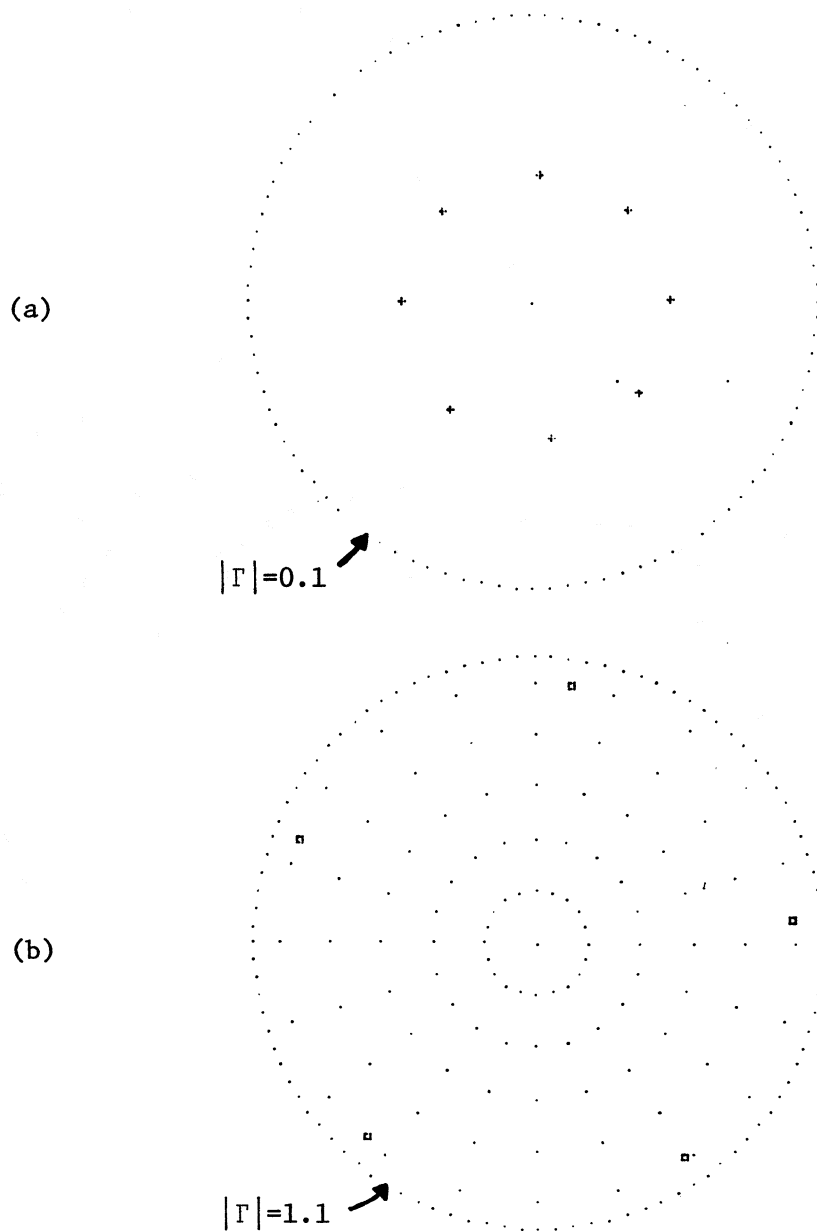


Fig. 3. Measurements of a sliding load and sliding short at 3.0 GHz:
(a) H-P model 911C sliding load, 1 cm increments;
(b) Maury Microwave model 1929-2 sliding short,
1 cm increments.

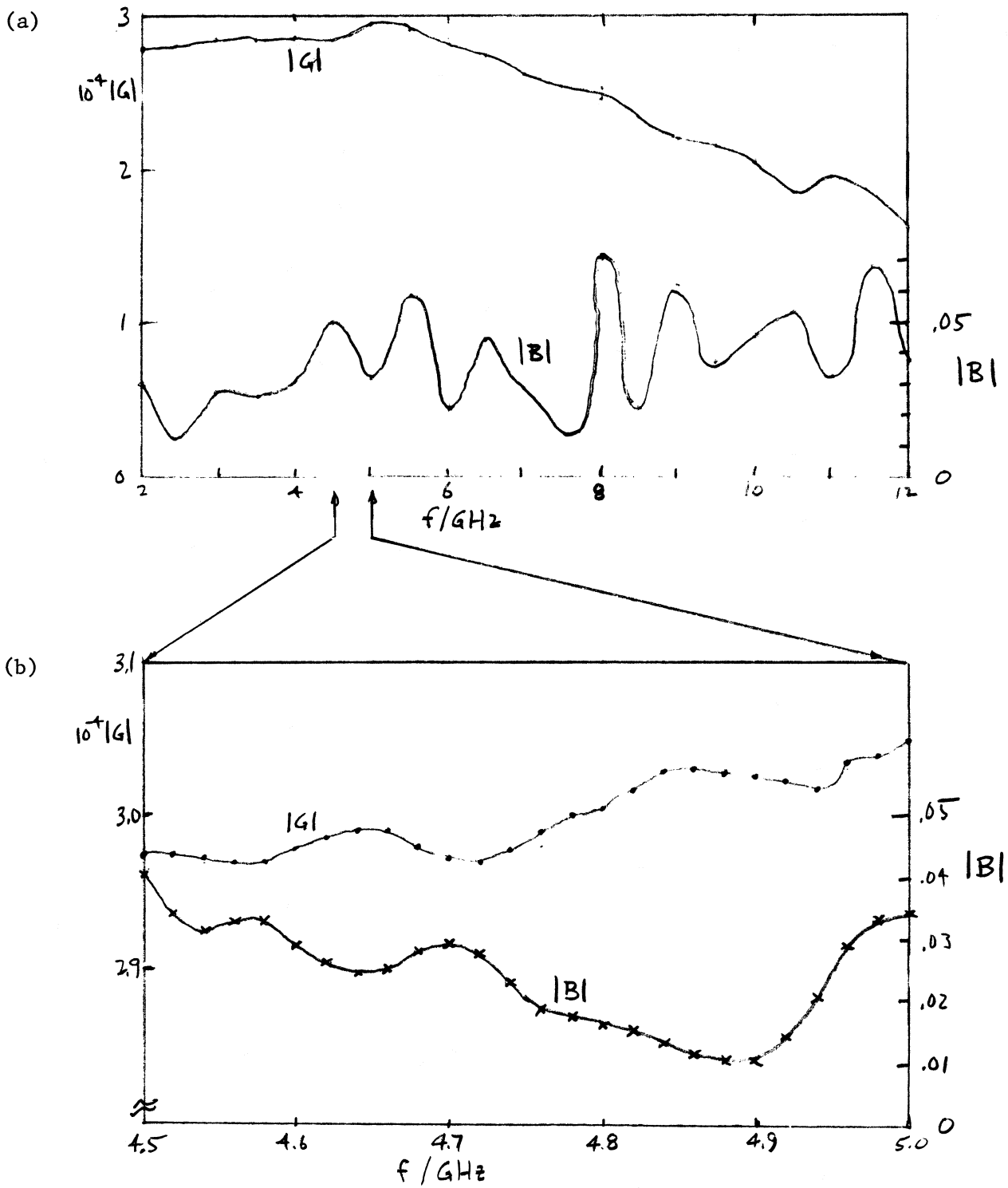


Fig. 4. Variation of calibration parameters $|B|$ and $|G|$ with frequency: (a) wide band; (b) narrow band.

the network analyzer at 12 GHz, and apparently there is only a gradual deterioration in performance at the upper end. Figure 4(b) displays a sample of the fine structure: 4.5 to 5.0 GHz is plotted every .02 GHz. The parameters plotted are derived by rewriting eqn. (5) as

$$\Gamma = \frac{x - C}{D - xB} = \frac{x - C}{G - (x - C)B}$$

so that $G = D - CB$ measures the complex gain, after correcting for the zero offset C .

Finally, some brief tests were done in the 12 to 16 GHz range, beyond the nominal limit of the instrument. The network analyzer remained in lock, although insufficient power was available to keep the front panel meter in the "operate" range. The gain, $|G|$, continued to drop smoothly. Re-measurement of the calibration standards gave errors of around 1% in Γ .

VI. Uncorrected Errors and Possible Improvements

The model of Figure 1 assumes that the complex ratio meter's response is linear, except for a possible d.c. offset. But it is possible that the real and imaginary channels may have unequal gains, and that they may not be perfectly in quadrature. In that case, a more accurate model would be

$$\begin{bmatrix} x_r \\ x_i \end{bmatrix} = \begin{bmatrix} g_{rr} & g_{ri} \\ g_{ir} & g_{ii} \end{bmatrix} \begin{bmatrix} \text{Re}(a'/b') \\ \text{Im}(a'/b') \end{bmatrix} + \begin{bmatrix} g_{or} \\ g_{oi} \end{bmatrix} \quad (9)$$

where subscripts r and i refer to real and imaginary parts, respectively. The model of Fig. 1 corresponds to $g_{rr} = g_{ii} = \text{Re}(g)$ and $g_{ir} = -g_{ri} = \text{Im}(g)$. Eqn. (9)

contains six real parameters, compared with four (two complex) for the simpler model. Note that if $|a'/b'|$ is constant and $\angle a'/b'$ is varied, then x describes an ellipse in the complex plane; the additional parameters may be thought of as the ellipticity and position angle of this ellipse.

Most complex ratio meters operate by heterodyning the input signals to a constant I.F., where the measurement is actually made. Therefore, the gain difference and quadrature error (or ellipse shape) just considered can usually be assumed to be independent of measurement frequency. In that case, the necessary additional corrections can be implemented without increasing the number of R.F. termination standards. One approach is to provide a means, at one frequency only, of varying the phase of the apparent reflection coefficient while keeping the magnitude constant. If the magnitude and phases are absolutely known, then three measurements suffice to determine the six parameters of (9). But if N measurements are made with unknown (but constant) magnitude and unknown (but different) phases, then there are $6 + 1 + N$ unknowns altogether and $2N$ (real) measurement results; a solution is possible for $N \geq 7$. (The H-P 8746B includes a variable-length extension to the reference path which closely approximates these measurement requirements.) From these results, one must still separate out those parameters which will be assumed constant with frequency and those which must be re-calibrated at each frequency; I will not give those details here.

It is suspected that most of the remaining errors in this system are due to inaccuracy in the calibration standards, particularly the low- Γ load. The system should be capable of errors in the -40 to -60 dB range, but it is hard to find a load this good, especially above 2 GHz. The usual solution is to use a sliding load, where $|\Gamma|$ is unknown but constant, and $\angle \Gamma$ is continuously

variable. In principle, measurements at three different (and unknown) slide positions are sufficient to determine the measurement that would have been made if Γ were zero; $|\Gamma|$ is determined as a byproduct (cf. Fig. 3a). In practice, it would be better to use 4 or 5 positions, especially if a wide frequency range is being calibrated, to ensure a good distribution of phases at all frequencies.

APPENDIX A

User Instructions for ANA Program

The program described in this report may be run in an NRAO Apple II+ system (see [9]) if the following files are available on a disk:

<u>File Name</u>	<u>Description</u>
ANA	EXEC file with initialization commands
ANA.A	Applesoft main program
LIB.HGR	NRAO binary library [6], version 3.1 or higher
SMITH	Image of Smith chart (normalized impedance)
SMITHY	Image of normalized admittance chart
SHAPES	Binary shape table with plotting symbols

The command EXEC ANA will then load and start the program. The ADIOS module should be connected as follows: COUT to sweep generator frequency control; AIN from real part of ratio meter; BIN from imaginary part; switches both to \pm and 10V.

Execution options are then selected by typing ESC followed by a digit. A summary of the options available is maintained at the bottom of the screen; here is a more complete explanation of each:

- 1: PARS - show list of current parameter settings. The list gives the internal variable name of each parameter; the value may be changed by an assignment statement in immediate mode.
- 2: CAL - enter calibration sequence (see below).
- 3: MEAS - enter measurement sequence (see below).
- 4: SHOW - switch display to HI-RES graphics screen. To return to text display, hit any key.

- 5: NEW - erase graphics screen and draw new background chart; Z, Y, and Γ charts are available.
- 6: COPY - make copy of graphics screen on printer
- 7: (this option not shown on bottom line of screen) - show list of frequencies at which calibration data has been stored, by index number.

Except when one of these options is being executed, the computer remains in the immediate mode of Applesoft BASIC. Thus, variables may be examined and their values changed, the program may be listed, and miscellaneous computations may be done. As long as no program lines are changed or added, stored data is unaffected.

Calibration sequence. The program first calibrates the voltage-controlled sweep oscillator by asking for its voltage limits and the corresponding frequencies. The lower and upper voltages are actually applied, so measured frequencies can be entered if available. These values are defaulted to the last values previously entered (zeros initially), and the defaults are displayed in parentheses; to use the default, type a carriage return alone (two CR's if two values requested, etc.). Next, the frequency range and increment are requested. No check is made that these fall within the limits previously given; if not, linear extrapolation results.

The program then asks whether one of the standards will be an open circuit or a delayed short, and it solicits the appropriate parameters for the selected case. Next, the user is prompted to connect each of the three standards in sequence. A double-beep tone (in addition to the screen prompt) indicates that the program is ready for the next standard. A printer listing of the calibration data can be obtained by responding "Y" to the "PRINTOUT?" prompt; any other

character bypasses printing. Finally, the program sets the measurement frequencies to include all calibration frequencies, reserves the top line of the screen for display of the current measurement frequency range, and exits to BASIC.

Measurement sequence. The program measures an unknown load at a subset of the calibration frequencies, computes its reflection coefficient using the corresponding calibration data, and writes the result both as a line on the text screen (magnitude and phase) and a plotted point on the graphics screen. There are two options: the frequency set can be scanned once or repeatedly; and the display can show either the listing or the plot during the measurements. With repeated scans, typing any key causes the program to stop after the present scan. When the plot is displayed, typing any key switches to the text display; conversely, ESC-4 switches to the plot display.

The measurement frequencies used are controlled by the parameters I1, I2, and I3 (ESC-1 shows current values); these are, respectively, the first and last index numbers and the increment in index number of the calibration frequencies. To see a list of calibration frequencies and their index numbers, type ESC-7. The calibration sequence sets $I1 = 0$, $I2 = II$, $I3 = 1$, where $II + 1$ is the number of calibration frequencies. For example, if you have calibrated from 2.0 to 4.0 GHz in 0.1 GHz steps and now wish to measure 2.0 to 4.0 in 0.5 GHz steps, just type

```
I3=5<CR>
```

If you later want to measure 3.0 to 3.5 in 0.1 GHz steps, type

```
I1=10: I2=15: I3=1<CR>
```

Finally, the program allows use of a "marker" on one frequency. Just set variable MK to the index of the frequency to be marked; if that frequency is

measured, its point will be plotted with a double-sized symbol. The marked frequency will be displayed on the top line of the text screen.

The plotted "points" are images of one of three plotting symbols. The symbol is changed in rotation (1, 2, 3, 1, 2...) each time the measurement sequence exits. To select a particular symbol, set variable SY = 1, 2, or 3 before typing ESC-3.

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