OF-2-6

A Millimeter-Wave Vector Network Analyzer

J. V. Bellantoni, G. C. Dalman, and R. C. Compton
School of Electrical Engineering
Cornell University

Abstract—A network analyzer for making accurate measurements in the 75-110 GHz band will be presented. The analyzer design is a novel variation on the six-port principle, and can be implemented at relatively low-cost. A 27-40 GHz prototype operating as a reflectometer will be demonstrated. Calibrations and measurements are performed interactively with the aid a Macintosh II equipped with a high resolution color display.

Background

Developments in device and component technology have stimulated the demand for fast, accurate instruments that measure scattering parameters at millimeter wavelengths. Among existing vector network analyzers, the six-port[1,2] is the most widely used alternative to expensive down-converter systems. In a six-port analyzer, four power detector readings are used to extract magnitude and phase information without the use of a vector voltmeter. We have developed an analyzer[3] that simulates the six-port analyzer but requires only two detectors. The analyzer is easy to build, and provides a low-cost alternative to commercially available millimeter-wave measurement systems.

The analyzer operates by sampling forward and reverse signals using back-to-back directional couplers. The tapped signals are combined in a hybrid which feeds the sum and difference into two detector diodes. An additional two measurements are obtained by switching a phase delay into one of the paths. These four measurements can be manipulated using six-port theory[1,2] to calculate the complex reflection coefficient. Given the high cost of sensitive, low-noise millimeter-wave detectors, the simplification from four detectors down to two is a significant advantage. Full s-parameter measuring capabilities can be realized by duplicating the hardware to create a transmission channel.

Also new in this analyzer design is the use of an electronic, reflection phase shifter. The phase shift is achieved by a beam lead PIN diode (Alpha DSG 6470) that is bonded across a tapered finline and integrated into a waveguide hybrid. The diode is controlled with a driver that converts a TTL signal from the computer into the spiked waveform necessary for high switching speeds.

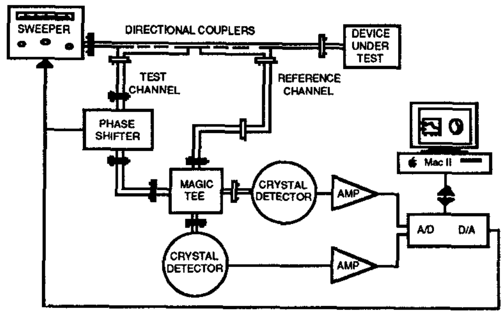

Figure 1 Schematic layout of the 27-40 GHz prototype reflectometer. The sum and difference of the forward and reverse travelling waves are sampled at two crystal detectors and passed via analog to digital convertor to a Macintosh computer. Two more independent measurements are made by electronically switching a phase shift into the reference channel.

System Configuration

We have developed two waveguide prototype systems, one for the 27-40 GHz band and a second for the 75-110 GHz band. The basic setup for the 27-40 GHz system is shown in Figure 1. An HP 8690A Sweep Oscillator is externally swept by a ±5 Volt signal at the EXT FM input. A magic tee feeds the sum and difference signals into two HP 11517A detectors. The video signals from the detectors are transferred via analog to digital converter to computer at a rate of 19,200 Baud.

A Macintosh II controls the sweeper, phase shifter and processes the data. With color display, math-coprocessor, and expansion slots it is ideal for this application. Control programs, written in Pascal, call the Macintosh Utilities and provide the basis for a simple, user-friendly interface (Figure 2). Pull-down menus allow the user to select from calibration, measurement, and display options. Alert boxes and an online help menu are used to guide the operator through each stage of measurement.

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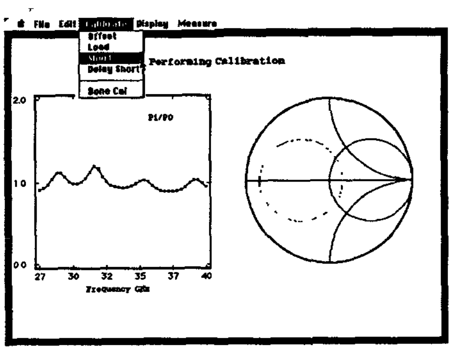

Figure 2 Display from the analyzer program showing the pull-down calibration menu. The calibration is performed in three steps by measuring the diode linearity, then a sliding short, and a sliding load.

Calibration and Measurement

Several different calibration schemes were investigated. The first approach assumed the couplers have infinite directivity. This scheme is simple and fast, but inaccurate. An equivalent circuit for the couplers, phase shifter and hybrid was formulated to determine the sources of error. This circuit was analyzed using a computer aided design package *Puff*, codeveloped by one of the authors[4]. The couplers and magic tee are specified as user definable elements whose s-parameters are read from data files stored on disk. *Puff* was used to study errors due to the non ideal characteristics of the components.

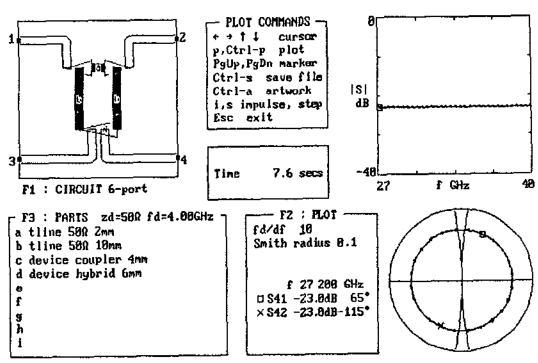

Figure 3 Puff output showing the six-port simulations. The sweeper signal enters at port 1 and the device under test is positioned at port 2. Port 3 and 4 monitor the signal at the detectors

Based on these simulations and measurements, a second more general calibration scheme was implemented. An approach which used a non-contacting sliding short and a sliding load was found to give the most accurate and reproducible results. From six-port junction theory[1,2] we know the power at each of the measurement ports is related to the reflection coefficient Γ by

$$P_i = \kappa_i \left|\Gamma - C_i\right|^2 \qquad i = 3, \ldots, 6$$

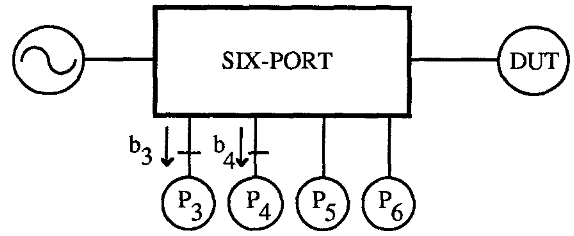

Figure 4 Schematic diagram of the 6-port analyzer. The power detected at port three (P_3) is proportional to $|b_3|^2$.

In order to eliminate errors due to fluctuations in sweeper output power, P_3, P_5, and P_6 are normalized by P_4. The resulting normalized powers form circles in the Γ plane. The intersection of these circles determines the value for Γ. A calibration procedure is required to determine the unknown constants C_i and κ_i. We follow the approach outlined by Engen[5] in which three of the four power measurements are used to map the data onto an intermediate complex plane $w = b_3/b_4$ (Fig. 4). This procedure is well suited to six-ports in which there is no power reading that is proportional to the power incident on the device under test. Once w is calculated, the analyzer is effectively reduced to a four port vector-voltmeter which may then be calibrated using a standard scheme like those used in the HP 8510 network analyzer. The complex quantity w is related to Γ by a bilinear transformation of the form

$$\Gamma = \frac{w - e_{df}}{e_{rf} + e_{sf}(w - e_{df})}$$

the constants e_{df}, e_{sf} and e_{rf} can be calculated by measuring three standards, for example a short, delay short and a load.

At lower power levels the voltage output of the diode detectors is proportional to rf power. However, to increase the dynamic range and decrease the effects of noise, high power levels are desirable. At high levels, the detector output voltage V_{diode} is proportional to square of the rf power P_{rf}. We can write

$$V_{\text{diode}} = AP_{\text{rf}} + BP_{\text{rf}}^2$$

the constants A and B are determined at each frequency, using a least squares fit, by varying the sweeper power and comparing the diode response with an HP 432A power meter. Calibrating for this non-linearity in the diode output voltage led to significant improvements in measurement accuracy. The diodes were also found to have large VSWR's. Two isolators were added to minimize these reflections.

Several hardware layouts were tried before reaching an optimal configuration. If there is a large difference in path lengths between the reference and reflected signals, then small frequency fluctuations in successive sweeps will manifest themselves in large phase errors. Care must also be taken in the relative strengths of the test and signal channels. If the reference channel has a magnitude less than the reflected channel for a specific DUT, then under certain conditions the two signals can cancel to make P_4 small or zero. This can lead to a numerically unstable calculation of Γ.

Accuracy

There are two main sources of inaccuracy that have been identified in the 27 to 40 GHz prototype. Noise from the detector diodes and instrumentation amplifiers generates non systematic errors in the measurements. Analytical formulae for the error that this noise introduces can be derived, providing certain assumptions hold[2]. To avoid making these assumptions, a series of monte carlo simulations were performed by computer. These simulations allow the effects of noise as well as alternative measurement schemes to be investigated.

In the simulations, measurement data is generated by computer with a prescribed noise level. The noise is calculated using a standard uniform random number generator followed by a transformation to produce a normal distribution with given standard deviation[6]. The simulations provide useful insights on the sources of error and how to minimize them. They were used to map out the scatter in measurements in the Γ plane due to noise.

Figure 4 shows the uncertainty for our 27 GHz to 40 GHz system. In plotting the error, we have assumed a 12 bit analog to digital converter and a noise component with a standard deviation of 4 bits. We have deliberately exaggerated the noise error to illustrate the distribution in uncertainties. The errors appear to be larger on the top side of the Smith chart than on the bottom. In practice the actual error is approximately eight times smaller than plotted in figure 4.

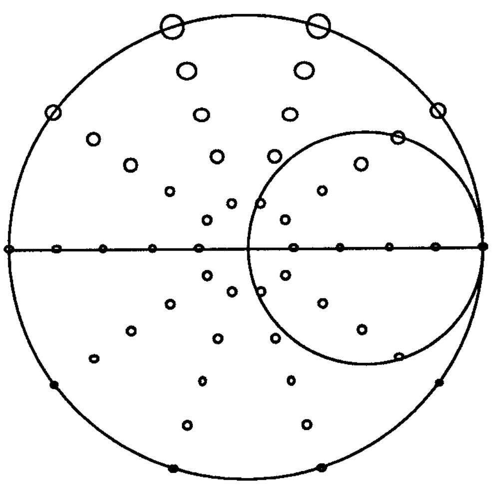

Figure 5 Noise from the detector diodes, intrumentation amplifiers and analog to digital conversion gives rise to the uncertainties in the reflection coefficient. The uncertainties have been magnified by a factor of eight.

In calculating Γ we have not used the ratio P_6/P_4. Because of the high degree of correlation between P_6/P_4, P_5/P_4 and P_3/P_4 we found that including the ratio P_6/P_4 did not significantly reduce the effect of noise. However, the ratio P_6/P_4 can be used to eliminate sign ambiguities when measuring active circuits.

The second source of error comes from non-ideal behaviour of the calibration standards. A sliding short and a sliding load are used in the calibration procedure. After calibration, the accuracy and reproducibility of these standards were determined by measuring their reflection coefficients at several different positions. Figure 6 shows measurements for the sliding short. Non-idealites in the sliding short manifest themselves in small errors in the phase and magnitude of the reflection coefficient. Based on these measurements we estimate the analyzer to be accurate to better than 1% in magnitude and phase.

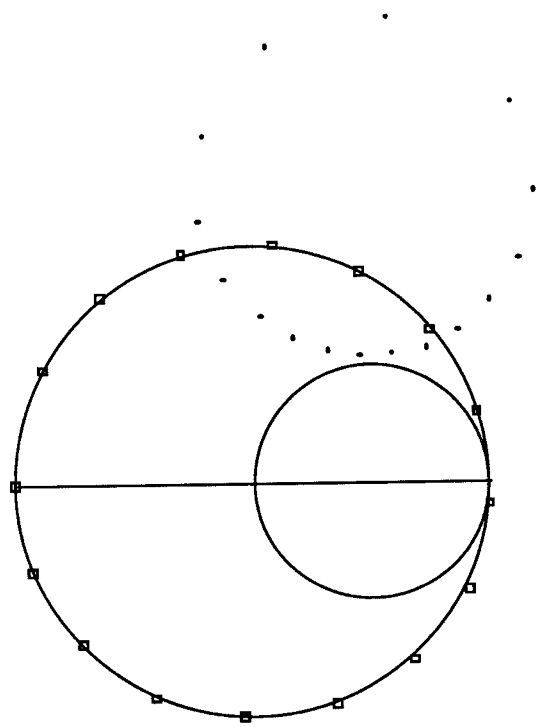

Figure 6 Reflection coefficient Γ for a sliding short at 30 GHz (squares). Also plotted are the corresponding w values (dots) in the w plane.

Similar tests were performed on the sliding load. The load was built by machining a slab of absorber into a pyramid and glueing it to the end of a micrometer drive. The VSWR was found to be less than 1.05 over the waveguide band. While it varied slightly with frequency, it remained at a level comparable to the noise (Figure 7).

The sliding short and load tests indicate that the system accuracy is presently limited by noise. The noise originates from several sources. The main source appears to be noise generated by the computer and the analog to digital converter. Increased accuracy is obtained by averaging over several measurements. The present converter is located in a box external to the Macintosh II. When operating, this generates considerable noise. Improved performance may be possible by going to a lower noise analog to digital system or chopping the sweeper.

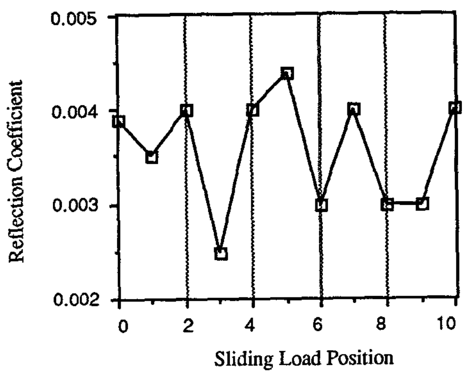

Figure 7 Magnitude of the reflection coefficient for the sliding waveguide load at 30 GHz.

The 27-40 GHz analyzer has been used to measure reflection coefficients of a number of components including semiconductor devices and was used to develop a waveguide to microstrip transition[7].

The 75-110 GHz system uses a Micro-Now Backward wave oscillator and a mechanical phase shifter in the test path. Refinements in the 75-110 GHz analyzer have concentrated on reducing the effects of frequency drift in the sweeper.

Acknowledgements

The authors would like to thank C. A. Lee, H. Kondoh, J. A. Faist, and D. M. Cleary for their work on preliminary versions of the analyzer. W. L. Williams provided many useful suggestions for improving the measurement accuracy. This work was supported by the Defense Advanced Research Projects Agency, Alpha Industries and the Raytheon Company.

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