High-Frequency Impedance Analyzer

A new one-port impedance analyzer measures high-frequency devices up to 1.8 GHz. Using a current-voltage method, it makes precise measurements over a wide impedance range. A special calibration method using a low-loss capacitor realizes an accurate high-Q device measurement. Many types of test fixtures are introduced because they are a key element in any test system.

by Takanori Yonekura

In research and development, component qualification, and RF and digital manufacturing, there is increasingly a need to make impedance measurements on chip components, such as chip inductors, capacitors, varactor diodes, and p-i-n diodes, and on other surface mount devices. Often the capacitances and inductances are very small and have impedances much greater or much less than 50 ohms at the operating frequencies. Traditionally, vector network analyzers are used to measure impedance in the RF range, but they are limited to measuring impedances near 50 ohms.

The new HP 4291A RF impedance analyzer (Fig. 1) is designed for passive surface mount device testing and material analysis at frequencies from 1 MHz to 1.8 GHz. Using an RF current-voltage measurement technique, it provides improved measurement accuracy over a wide impedance range, making it possible to test components at operating frequencies and to evaluate non-50-ohm components more accurately.

New surface mount component test fixtures for use with the new analyzer save time and eliminate the need for custom fixtures. These fixtures and the analyzer's built-in calibration and compensation routines ensure measurement accuracy.

This paper is a revised and updated version of reference 1 and is published here with permission of the Electronic Components Institute Internationale Ltd.

The analyzer's firmware provides direct impedance reading, frequency-swept measurements and many advanced functions such as equivalent circuit analysis, limit lines, and markers.

General Impedance Measurements

A general impedance measurement schematic using two vector voltmeters is shown in Fig. 2. In this case, the true impedance (Z_x) of a device under test (DUT) is determined by measuring the voltages between any two different pairs of points in a linear circuit.

$$Z_x = K_1 \frac{K_2 + V_r}{1 + K_3 V_r}, \tag{1}$$

where K_1, K_2, and K_3 are complex constants and V_r is the voltage ratio V_2/V_1.

There are three unknown parameters related to the circuit in equation 1. Once we know these parameters, we can calculate the impedance of the DUT from the measured voltage ratio $V_r = V_2/V_1$. The procedure that estimates these circuit

Fig. 1. The HP 4291A RF impedance analyzer measures impedances of components and materials at frequencies from 1 MHz to 1.8 GHz. New surface mount test fixtures and built-in calibration and compensation routines ensure accuracy.

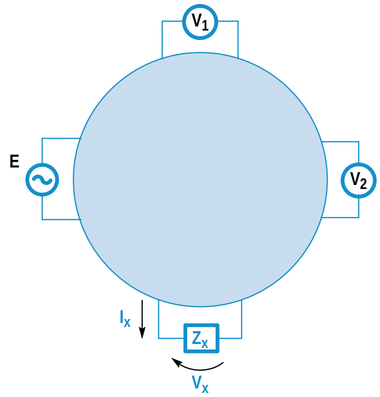

Fig. 2. General schematic for impedance measurement using two vector voltmeters.

parameters is called *calibration* and one method is *open-short-load* (OSL) calibration. Calculation of Z_x from the measured voltage ratio V_r according to equation 1 is called *correction*.

Transducers

We call a linear circuit such as the one in Fig. 2—one that relates a signal source, two vector voltmeters, and a DUT—a *transducer*. Transducers are the key element in impedance measurements. Two types of transducers of interest here are the directional bridge and the transducer in a current-voltage (I-V) method. Let's compare these in terms of their sensitivity to gain variance in the vector voltmeters in Fig. 2.

Directional Bridge. Directional bridges (see Fig. 3) are used in many network analyzers, mainly to measure impedances near 50 ohms. In this case, the bilinear transformation is:

$$Z_x = R_o\frac{1+\Gamma}{1-\Gamma}, \tag{2}$$

where $\Gamma = (Z_x - R_o)/(Z_x + R_o)$ is the reflection coefficient, $V_r = V_2/V_1 = (-1/8)\Gamma$, and $R_o = 50$ ohms is the characteristic impedance. The parameters in equation 1 are $K_1 = -8R_o$, $K_2 = -1/8$, and $K_3 = 8$.

Now assume that the vector voltmeters in Fig. 3 are not ideal but have some gain variance. The measured voltages V_1 and V_2 and the calculated impedance Z_x are:

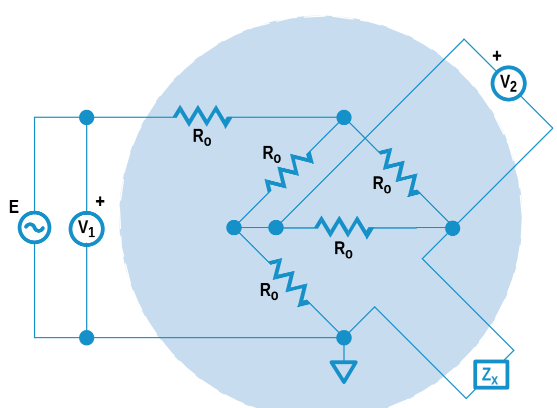

Fig. 3. Directional bridge circuit.

$$V_1 = E\alpha_1$$

$$V_2 = (-1/8)E\Gamma\alpha_2$$

$$Z_x = R_o(1+\Gamma)/(1-\Gamma),$$

where α_1 is the gain of vector voltmeter 1, α_2 is the gain of vector voltmeter 2, $\Gamma = -8V_r\alpha_r$ is the measured reflection coefficient, $V_r = V_2/V_1$ is the voltage ratio, and $\alpha_r = \alpha_1/\alpha_2$ is the ratio of the voltmeter gains.

We define the calculated impedance sensitivity S to the voltmeters' gain variance as follows:

$$S = \frac{\delta Z_x/Z_x}{\delta\alpha_r/\alpha_r}. \tag{3}$$

This sensitivity can be considered as the inverse of the magnification of gain variance. The smaller S is, the smaller the error in the calculated impedance.

For the directional bridge, equation 3 is:

$$S = \frac{\delta Z_x}{\delta\Gamma}\frac{\delta\Gamma}{\delta\alpha_r}\frac{\alpha_r}{Z_x} = \frac{1}{2}\frac{Z_x^2 - R_o^2}{Z_xR_o}$$

$$\begin{aligned} S &= (-1/2)R_o/Z_x && \text{for } |Z_x| \ll R_o \\ &= 0 && \text{for } |Z_x| = R_o \\ &= (1/2)Z_x/R_o && \text{for } |Z_x| \gg R_o. \end{aligned}$$

This implies that this type of transducer has little sensitivity to the voltmeter gain variance when the DUT impedance is near R_o (50 ohms). The gain variance of the voltmeters behaves as an offset impedance with a magnitude of $(1/2)R_o |\Delta\alpha_r/\alpha_r|$ when the DUT impedance is much smaller than R_o, where $\Delta\alpha_r$ is the change in the gain ratio α_r. The gain variance of the voltmeters behaves as an offset admittance with a magnitude of $(1/2)G_o|\Delta\alpha_r/\alpha_r|$ when the DUT impedance is much larger than R_o, where $G_o = 1/R_o$.

Fig. 4 shows this characteristic.

I-V Method. Fig. 5 shows the simplest transducer for the I-V method. The bilinear transformation in this case is:

$$Z_x = R_oV_r, \tag{4}$$

where $R_o = 50$ ohms is a resistor that converts the DUT current to a voltage and $V_r = V_2/V_1$. The parameters in equation 1 are $K_1 = R_o$, $K_2 = 0$, and $K_3 = 0$.

We also assume that there is some gain variance in the vector voltmeters in Fig. 5. The measured voltages V_1 and V_2 are related to the calculated impedance Z_x as follows:

$$V_1 = \alpha_1E\frac{R_o}{Z_x + R_o}$$

$$V_2 = \alpha_2E\frac{Z_x}{Z_x + R_o}$$

where α_1 is the gain of vector voltmeter 1, α_2 is the gain of vector voltmeter 2, and $V_r = V_2/V_1$ is the voltage ratio. Thus,

$$Z_x = R_oV_r\,\alpha_r,$$

where $\alpha_r = \alpha_1/\alpha_2$ is the ratio of the voltmeter gains.

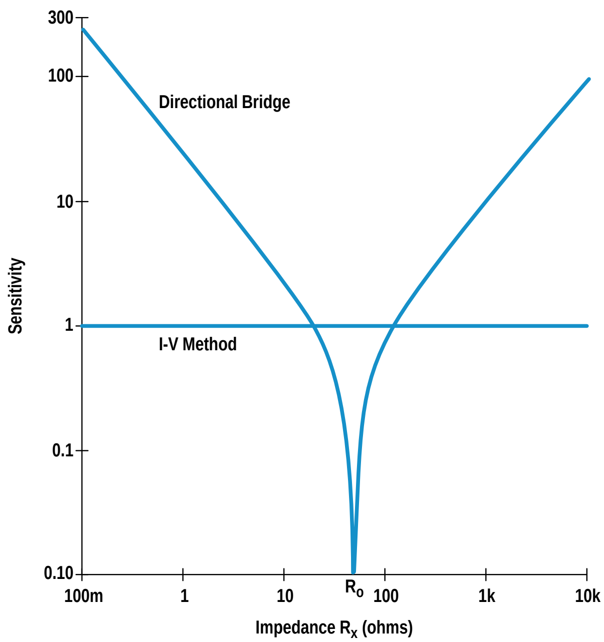

Fig. 4. Sensitivity of the directional bridge and I-V methods to voltmeter gain variance.

In this case, the sensitivity is given by:

$$S = \frac{\delta Z_x / Z_x}{\delta \alpha_r / \alpha_r} = 1.$$

The error ratio $(\Delta Z_x/Z_x)/(\Delta\alpha_r/\alpha_r)$ is always constant and equal to unity. For example, if the voltmeter gain ratio α_r changes by 1%, an impedance error of 1% is incurred for any DUT.

New RF Impedance Analyzer

The foregoing analysis shows that the voltmeter gain variance is neither suppressed nor magnified for all DUT impedances by an I-V method transducer. This characteristic is desirable for wide impedance measuring capability. Therefore, we adopted this type of transducer for the new HP 4291A one-port RF impedance analyzer.

Fig. 6 shows the basic circuit of the transducer in the HP 4291A, which is a modified version of Fig. 5. The high-impedance configuration (switch closed) realizes perfect open and imperfect short conditions, while the low-impedance configuration (switch open) realizes imperfect open and perfect short conditions.* In either switch position, the output impedance at the DUT port is always R_o.

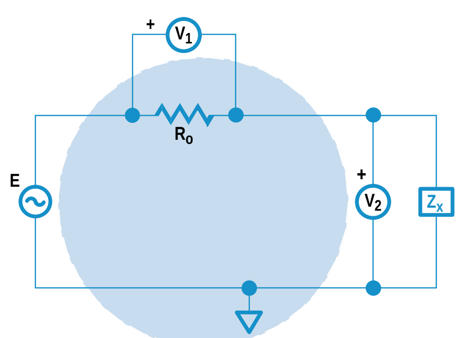

Fig. 5. I-V method.

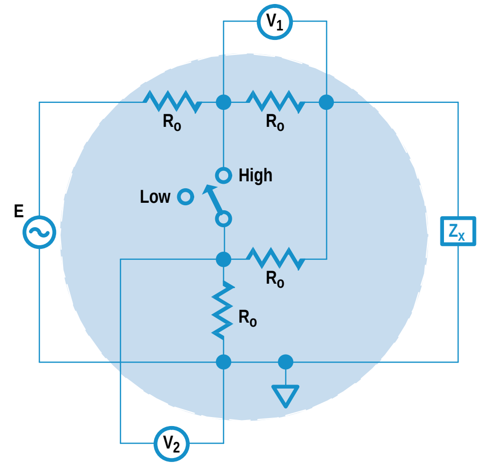

Fig. 6. Basic transducer circuit of the HP 4291A RF impedance analyzer with switch indicating circuit configurations for high-impedance and low-impedance measurements.

Fig. 7 shows the actual HP 4291A transducer circuit configurations. A number of considerations influenced the design of these circuits. First, because a wideband switch with small nonlinearity and small transients over a wide signal range is not easily realized, we divided the circuit of Fig. 6 into two separate circuits. Second, because the minimum frequency of the analyzer is 1 MHz, the floating voltmeter (V_1), which corresponds to the current meter, is easily realized by using a balun. Third, we adopted a circuit in which the voltmeter readings change by the same percentage if the floating impedance of the balun changes. Thus, the voltage ratio V_r does not change and stable impedance measurements are realized.

Block Diagram

A conceptual block diagram of the HP 4291A including the transducer discussed above is shown in Fig. 8. Only the parts relevant to this discussion are included. The mainframe is similar to the HP 4396A network and spectrum analyzer.[3]

Two key features of the HP 4291A are time division multiplex operation and impedance ranging. Two voltmeters are obtained by time division multiplexing one voltmeter. The multiplexing period is 2 ms. This ensures that slow drift of the voltmeter gain does not affect the impedance measurement. With this method, the signal path after the multiplexer can be extended. The HP 4291A uses a 1.8-m cable between the transducer and the instrument mainframe. This allows wide flexibility in constructing a test system using automatic device handlers. The single-path configuration results in good temperature characteristics even with an extended cable.

At frequencies below 200 MHz there is an expanded range. In the expanded range there is a gain difference between the voltage channel and the current channel ahead of the multiplexer. This impedance ranging offers stable measurements for DUTs with impedances that differ greatly from 50 ohms.

Fig. 9 shows HP 4291A impedance measurement specifications. General error factors are the uncertainties of standards

* A perfect open condition means that voltmeter V_1 reads zero with the DUT port open. A perfect short condition means that voltmeter V_2 reads zero with the DUT port shorted.

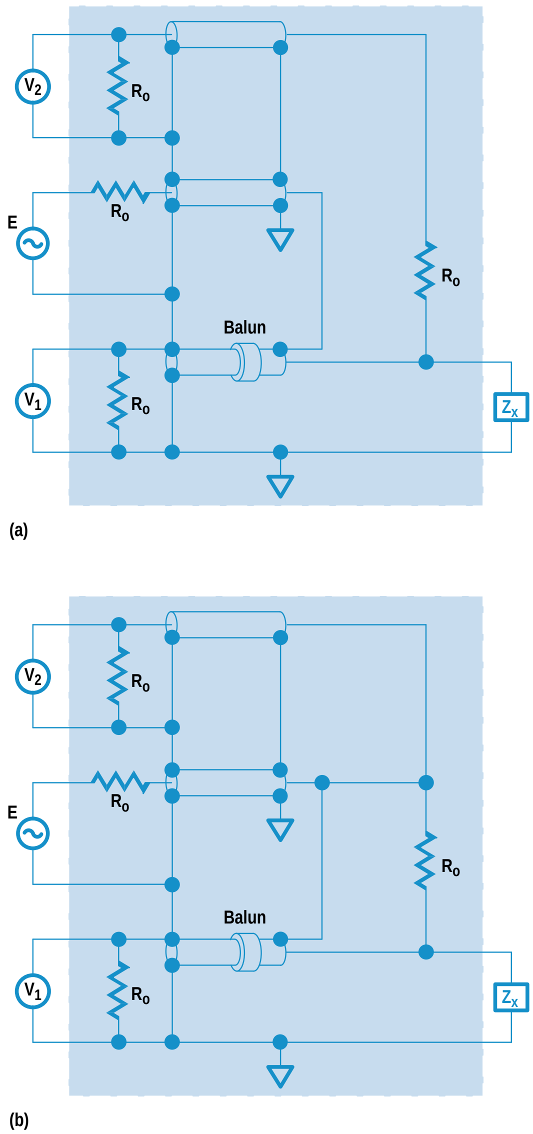

Fig. 7. (a) Actual HP 4291A transducer circuit for low-impedance measurements. (b) Circuit for high-impedance measurements.

used in the calibration, instabilities, and interpolation errors. The instabilities consist mainly of connection nonrepeatability, long-term drift, circuit nonlinearity, temperature coefficients, and noise. The instabilities and interpolation errors are small enough that the impedance phase errors can be reduced by means of the special calibration discussed next.

High-Q Measurements

Normally, the accuracy requirement for impedance phase measurements is greater than that for impedance magnitude measurements. The HP 4291A has a special, easy-to-use, calibration for measurements of devices having high Q (quality factor).

Even if the stability of the instrument is good enough, accurate Q measurements cannot be made without adequate phase calibration. For instance, if we want to measure the Q factor with 10% uncertainty for a DUT whose Q value is 100, the uncertainty in phase must be smaller than 10^{-3}. The phase accuracy of the instrument is determined almost entirely by the uncertainty of the 50-ohm load standard used in the OSL calibration. One way to improve phase measurement accuracy is to use a phase-calibrated load standard. However, it is not guaranteed that the phase uncertainty for a calibrated 50-ohm load is smaller than 10^{-3} at high frequencies, such as 1 GHz.

Another way to improve phase measurement accuracy is to use, in addition to the normal open-short-load standards, a low-loss air capacitor as a second load (LOAD2). The dissipation factor (D) of the air capacitor should be below 10^{-3} at around 1 GHz. With this method, the uncertainty in the measured phase is decreased from the phase uncertainty of the 50-ohm load (LOAD1) to the uncertainty of the dissipation factor D of the low-loss capacitor (LOAD2) for almost all DUT impedances. The next section gives the details of this method.

Modified OSL Calibration[2]

We want a calibration method that reduces the error in phase measurement in spite of the existence of phase error for the 50-ohm load. We have the 50-ohm load standard whose impedance magnitude is known but whose impedance phase is not. We add another load (LOAD2) whose impedance phase is known but whose impedance magnitude is not. We use a low-loss capacitor as the second load. There are still at most three unknown circuit parameters. However, two more unknowns related to standards are added. Let us define the problem. There are eight real unknown parameters:

- Circuit parameter K_1 (two real parameters)
- Circuit parameter K_2 (two real parameters)
- Circuit parameter K_3 (two real parameters)
- The impedance phase θ_{ls1} of the 50-ohm load (one real parameter)
- The impedance magnitude Z_{abs_ls2} of the low-loss capacitor LOAD2 (one real parameter).

We have solved this problem analytically. For the simplest case where both the open and the short standard are ideal, the three circuit parameters are found as follows:

$$K_1 = AZ_{lsi}R_o$$
$$K_2 = -Z_{sm}/R_o \qquad (5)$$
$$K_3 = -Y_{om}R_o$$

where:

R_o = characteristic impedance
$A = (1 - Z_{lmi}Y_{om})/(Z_{lmi} - Z_{sm})$
Y_{om} = measured admittance for open standard
Z_{sm} = measured impedance for short standard
Z_{lmi} = measured impedance for load standard i (i = 1 for LOAD1, i = 2 for LOAD2)
Z_{lsi} = true impedance for load standard i
$Z_{ls1} = Z_{abs\_ls1}\exp(j\theta_{ls1})$
$Z_{ls2} = Z_{abs\_ls2}\exp(j\theta_{ls2})$
$\theta_{ls1} = \theta_2 - \theta_1 + \theta_{ls2}$
$Z_{abs\_ls2} = (A_1/A_2)Z_{abs\_ls1}$
Z_{abs_ls1} = impedance magnitude for LOAD1 (50-ohm, known)
θ_{ls2} = impedance phase for LOAD2 (low-loss capacitor, known)
$\theta_1 = \arg((1 - Z_{lm1}Y_{om})/(Z_{lm1} - Z_{sm}))$
$\theta_2 = \arg((1 - Z_{lm2}Y_{om})/(Z_{lm2} - Z_{sm}))$

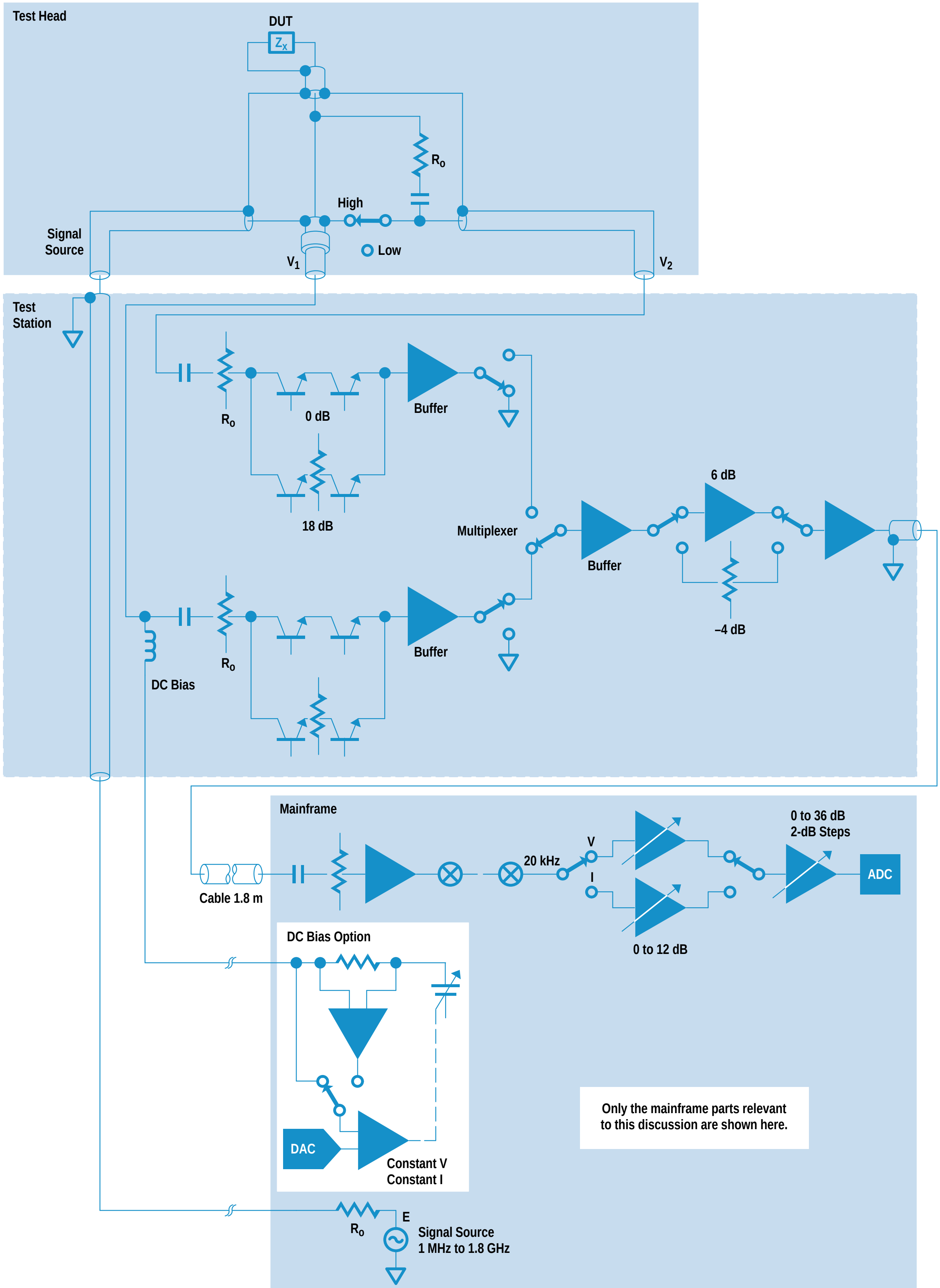

Fig. 8. Conceptual block diagram of the HP 4291A with switch (in test head) showing the difference between the transducers for high-impedance and low-impedance measurements.

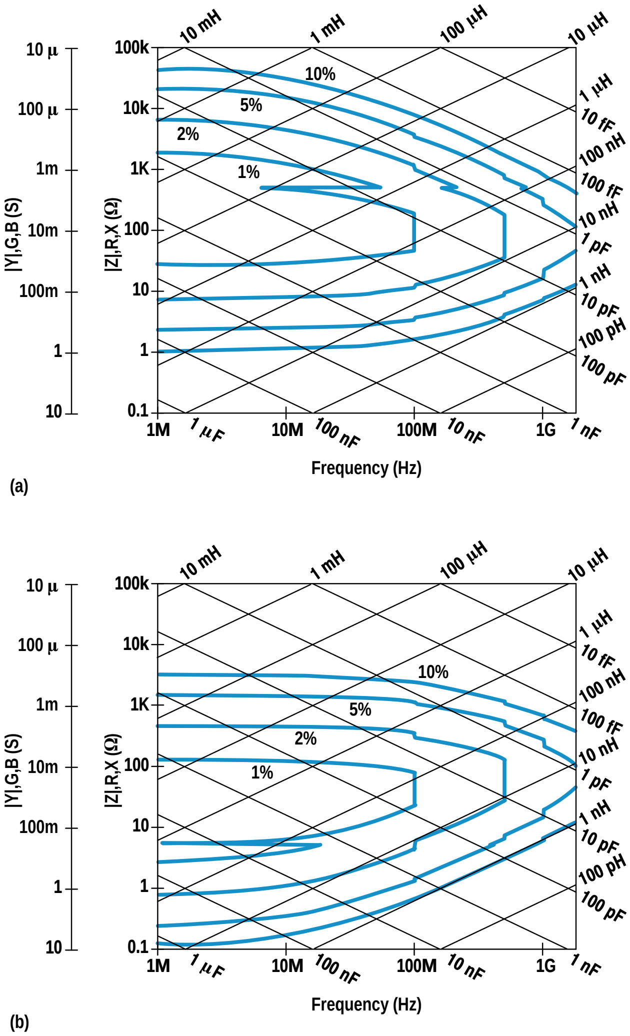

Fig. 9. (a) Errors for impedance magnitude with the transducer for high impedance. (b) Errors for impedance magnitude with the transducer for low impedance.

$$A_1 = |(1 - Z_{lm1}Y_{om})/(Z_{lm1} - Z_{sm})|$$
$$A_2 = |(1 - Z_{lm2}Y_{om})/(Z_{lm2} - Z_{sm})|$$

For the actual case these circuit parameters are expressed by far more complicated equations. Therefore, we adopted a simpler procedure consisting of two steps. Step 1 is as follows:

- Regard the impedance of the 50-ohm load as $Z_{ls1} = 50+j0$ (that is, the phase of LOAD1 is zero).
- Find the circuit parameters K_1, K_2, and K_3 by normal OSL calibration using the load value Z_{ls1}.
- Execute correction for LOAD2 and get the corrected impedance Z_{corr2}.
- Calculate the phase difference $\Delta\theta$ between the phase of Z_{corr2} and the true phase of LOAD2.

Step 2 is as follows:

- Modify the impedance of LOAD1 to Z'_{ls1} whose phase is $-\Delta\theta$ and whose impedance magnitude is still 50 ohms.
- Calculate the circuit parameters again by normal OSL calibration using the modified load impedance Z'_{ls1}.

Although this is an approximate method, it is accurate enough for our purposes. We call this method the modified OSL calibration.

Phase Measurement Errors

The following error factors affect phase measurement accuracy using the modified OSL calibration:

- Uncertainty in the impedance magnitude of LOAD1
- Impedance phase of LOAD1
- Impedance magnitude of LOAD2
- Uncertainty in the impedance phase of LOAD2
- Uncertainty in the admittance magnitude of the open standard.

Notice that the second and third factors would not cause any error if we were using the analytical solution. Computer simulations have shown that:

- The phase measurement error caused by the uncertainty in the impedance magnitude of LOAD1 is small.
- The phase measurement error caused by the impedance phase of LOAD1 is small.
- The phase measurement error caused by the impedance magnitude of LOAD2 is small.
- The uncertainty in the impedance phase of LOAD2 directly affects the phase measurement error.
- For reactive DUTs, the phase measurement error caused by the uncertainty in the admittance magnitude of the open standard ($|\Delta Y_{open}|$) is reduced to $|R_o \Delta Y_{open}|(C_{open}/C_{ls2})$ in the modified OSL calibration, where R_o is the characteristic impedance (50 ohms), C_{open} is the capacitance of the open standard, and C_{ls2} is the capacitance of LOAD2 (low-loss capacitor). In the case of resistive DUTs, the error is the same as in the normal OSL calibration.

Fig. 10 shows the relationship between the phase error $|\Delta\theta|$ and the DUT impedance when the LOAD2 phase uncertainty $|\Delta\theta_{ls2}|$ is 500×10^{-6} radian in the modified OSL calibration. The relationship between $|\Delta\theta|$ and the DUT impedance is shown in Fig. 11 when the open admittance uncertainty $|\Delta Y_{open}|$ is 5 fF in the modified OSL calibration.

In summary, the phase measurement error when using the modified OSL calibration is mainly determined by the uncertainty in the impedance phase of LOAD2 and the uncertainty in the admittance magnitude of the open standard. We now evaluate these two items. The D factor for the capacitor (3 pF) used in the calibration can be small because the capacitor's dimensions are small and the space between the inner and outer conductors is filled almost completely with air.

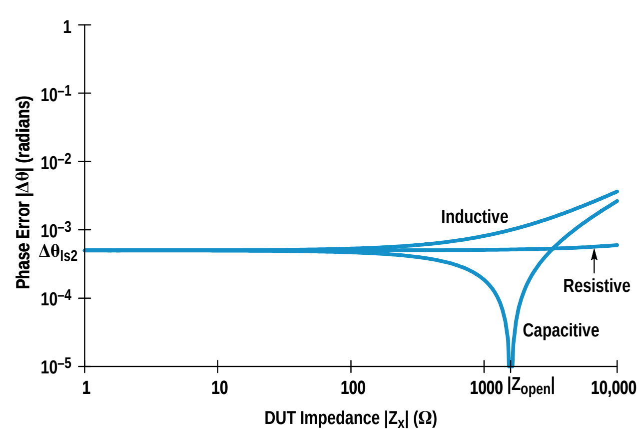

Fig. 10. Relationship between phase measurement error and the uncertainty $\Delta\theta_{ls2}$ in the impedance phase of LOAD2 (low-loss capacitor) at 1 GHz for resistive and reactive DUTs. $|Z_{open}|$ is the impedance of the open standard (about 2000 ohms).

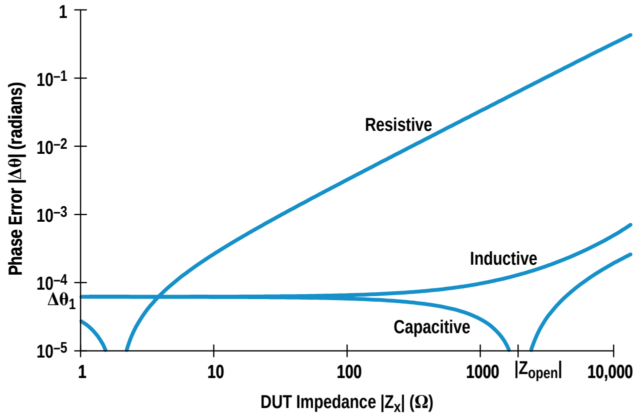

Fig. 11. Relationship between phase measurement error and the uncertainty in the admittance magnitude of the open standard at 1 GHz for resistive and reactive DUTs. $\Delta\theta_1 = |R_o \Delta Y_{open}| C_{open}/C_{ls2}$ where R_o is the characteristic impedance (approximately 50 ohms), $|\Delta Y_{open}|$ is the uncertainty in the admittance magnitude of the open standard (about 30 µS), C_{open} = 100 fF, and C_{ls2} = 3 pF. $|Z_{open}|$ is the impedance of the open standard (about 2000 ohms).

The D value has been estimated as 500×10^{-6} at 1 GHz in a residual resistance measurement at the series resonant frequency. The D factor increases with frequency f as $f^{1.5}$ because of the skin effect. By using zero as the D value for the capacitor during calibration, a phase measurement error of 500×10^{-6} is incurred at 1 GHz. The uncertainty for the open capacitance is ±5 fF at most, leading to a phase measurement error less than $\pm100\times10^{-6}$ at 1 GHz. Overall, a phase measurement uncertainty of 500×10^{-6} is incurred by using the modified OSL calibration.

Impedance Traceability

For an impedance performance check using the top-down method[4] we set up a kit traceable to U.S. national standards. This kit is calibrated annually at our in-house standards laboratory. Two major items in the kit are the 50-ohm load and a 10-cm-long, 50-ohm, beadless air line. The 50-ohm load is desirable because its frequency characteristic for impedance is very flat. The structure of the air line is very simple, so it is easy to predict its frequency characteristic and it is convenient to realize various impedances by changing frequencies with the line terminated in an open or short circuit.

The traceability path for the kit is shown in Fig. 12. The impedance characteristic of open-ended and short-ended air lines can be calculated theoretically from their dimensions and resistivity.[5] However, it is not easy to design a system to calibrate the dimensions of the air line in each individual kit. Therefore, only the dimensions of the reference air line of our standards laboratory is periodically calibrated. Calibration of the individual air line is executed by a network analyzer calibrated from the reference air line. The 50-ohm load calibration is done mainly by the quarter-wave impedance method and dc resistance measurement. The open termination is calibrated by a capacitance bridge at low frequencies and by the network analyzer at high frequencies. The short termination is treated as ideal. Uncertainties for the short termination consist of skin effect and nonrepeatabilities.

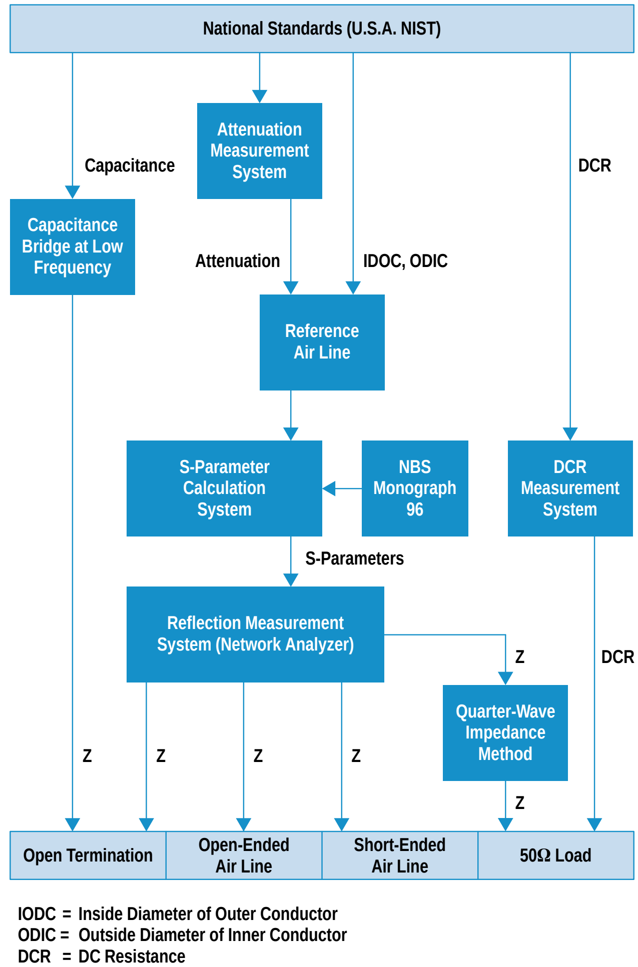

Fig. 12. Traceability path for the performance test kit.

Test Fixtures

In actual measurements, test fixtures are needed to accommodate different-shaped DUTs. As the frequency range goes up, fixtures that are able to handle smaller devices are needed. We have developed four types of fixtures:

- A fixture for surface mount devices with bottom electrodes
- A fixture for surface mount devices with side electrodes
- A fixture for very small surface mount devices
- A fixture for leaded components.

To reduce the error at the fixture terminal it is necessary to minimize the length from the reference plane of the APC-7 connector to the fixture terminal and to minimize the connection nonrepeatability. The new fixtures' repeatability is almost five times better than our old ones. The typical nonrepeatability of the surface mount device fixtures is ±50 pH and ±30 mohms for short-circuit measurements and ±5 fF and ±2 mS for open-circuit measurements.

Fixture compensation is included in the firmware corresponding to correction at the fixture terminal. This reduces the errors generated in the circuit between the reference plane and the fixture terminal. The best compensation method is the OSL method. However, it is not easy to prepare a standard load having excellent frequency characteristics. As

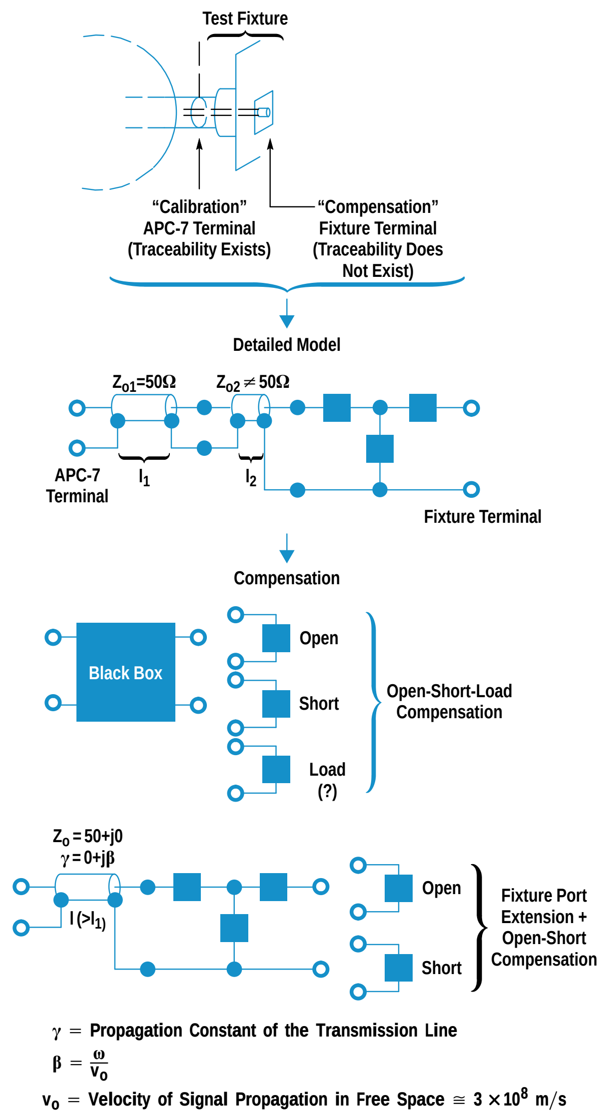

Fig. 13. Open-short-load (OSL) fixture compensation versus HP 4291A fixture port extension plus open-short compensation.

a more realistic alternative we provide another compensation function: fixture port extension combined with open-short correction at the fixture plane. This method assumes that there is a short transmission line between the APC-7 terminal of the transducer head and the DUT connector of the fixture (see Fig. 13). When the user selects one of our new fixtures on the HP 4291A display, an appropriate fixture port extension value—an equivalent length of ideal 50-ohm line previously determined for that fixture—is automatically set. The user then performs a compensation using open and short circuits.

The difference between the OSL method and the HP 4291A method is that the OSL method assumes ideal open, short, and load standards while the HP 4291A method assumes ideal open and short standards and an ideal transmission line. We feel that the assumption of an ideal transmission line is more realistic than the assumption of an ideal load for compensation over a wide frequency range.

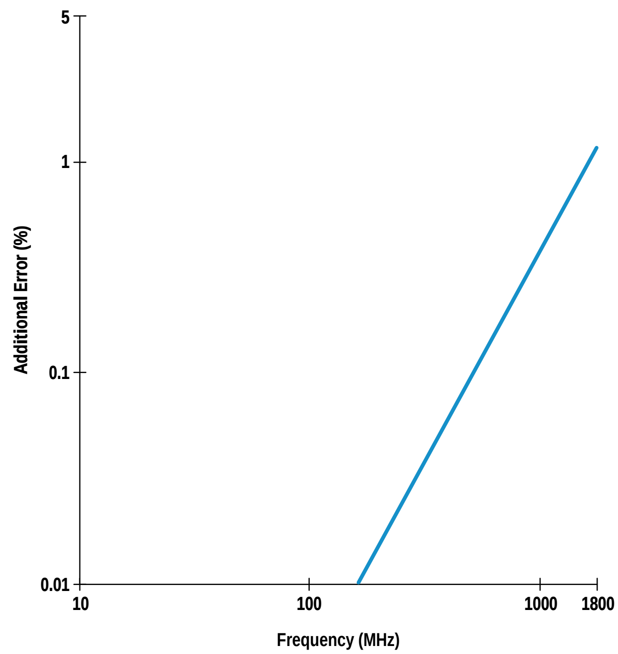

Fig. 14. Typical error contributions of the new test fixtures after fixture compensation.

Fig. 14 shows the typical test fixture error contributions when using this compensation function. These values are almost three times better than the errors for our former type of fixtures.

Conclusion

Selection of a transducer (as defined on page 68) is important for accurate impedance measurement. A new type of transducer based on the current-voltage method and having wide impedance measuring capability is used in the new HP 4291A RF impedance analyzer. A new phase calibration technique, a modified OSL calibration, has also been developed. It uses a low-loss capacitor as the second load and makes accurate Q measurements possible.

Acknowledgments

The author is grateful to K. Yagi, S. Tanimoto, and H. Wakamatsu, who provided helpful discussions.

References

1. T. Yonekura and M. Jansons, "High-Frequency Impedance Analyzer," *Proceedings of CARTS—Europe '93,* pp. 44-52.
2. T. Yonekura, "A New Calibration Method for Impedance Meters," *Proceedings of the 1994 IEEE Instrumentation and Measurement Technology Conference,* Shizuoka, Japan, May 1994, pp. 1004-1007.
3. S. Kawabata and A. Nukiyama, "A High-Performance 1.8-GHz Vector Network and Spectrum Analyzer," *Hewlett-Packard Journal,* Vol. 44, no. 5, October 1993, pp. 76-84.
4. *Traceability and the HP 8510 Network Analyzer,* Hewlett-Packard publication no. 5954-1552, November 1, 1985.
5. R.E. Nelson and M.R. Coryell, *Electrical Parameters of Precision, Coaxial, Air-Dielectric Transmission Lines,* NBS Monograph 96, June 30, 1966.