



Agilent PN 4291-1

New Technologies for Wide Impedance Range Measurements to 1.8 GHz

Product Note

Agilent 4291B RF Impedance/Material Analyzer

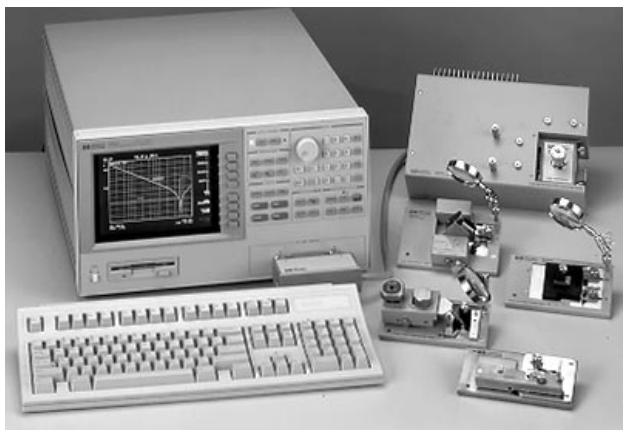
Introduction

With the current trends in the communications and data processing industries requiring higher performance, smaller physical size, lower cost, and higher reliability, accurate and efficient electronic component characterization is an increasingly important part of product design. Many of these advanced products operate at frequencies in the RF range.

Older low-frequency or indirect methods of component evaluation are often not able to provide the high quality impedance parameter information required for understanding component performance under actual operating conditions in the RF range. This product note describes new impedance measurement techniques and innovations con-

tained in the Agilent Technologies 4291B RF Impedance/Material Analyzer that allow accurate and efficient direct impedance measurement and analysis from 1 MHz to 1.8 GHz. Five general topics will be discussed:

- Limitations of traditional methods of impedance analysis
- Extending the frequency range and impedance magnitude range, while maintaining high accuracy
- Making high accuracy Quality Factor (Q) and Dissipation Factor (D) measurements at RF
- Increasing test flexibility by extending the device-under-test (DUT) location 1.8 meters from the instrument while maintaining accuracy
- Eliminating fixture errors-critical issues for accurate RF measurements



Agilent 4291B RF Impedance/Material Analyzer with SMD fixtures

Limitations of traditional impedance measurement solutions

Most conventional measuring instruments such as traditional LCR meters and impedance analyzers are limited to lower frequency analysis where the Four-Terminal-Pair method can provide very high accuracy. For higher frequencies a vector network analysis approach is often used by first measuring the reflection coefficient, then calculating impedance values. This indirect method may be useful for impedance values near 50Ω , but can be inaccurate for impedance values significantly higher or lower than 50Ω . In addition, calibration and removal of test fixturing errors can be a tedious procedure and in some cases, not possible. The issue of fixturing at RF can be a major source of



Agilent Technologies

Innovating the HP Way

error and expense. In general, impedance measurements in the RF range in the past have been very difficult and often yield widely varying results with questionable accuracy. For today's RF component or circuit designer, a new solution is needed to provide the accuracy and test efficiency required for complete RF impedance characterization of circuit elements.

Agilent now offers a new analyzer dedicated to the direct measurement of impedance and material parameters from 1 MHz to 1.8 GHz that overcomes many of the limitations of previous approaches. The 4291B provides highly accurate RF impedance measurements and offers a family of surface mount device (SMD) test fixtures. With a color display and powerful firmware, up to fifteen impedance parameters as well as equivalent circuit models and more are easily measured and displayed. The following topics describe this new analyzer's capabilities and measurement technology.

Expanding the range of impedance magnitudes

As shown in Figure 1, the 4291B provides an exceptionally wide range of impedance measurements. It is the ideal instrument for measuring very small capacitance (1 pF) and inductance (1 nH) values in the RF range.

The Agilent 4291B achieves its accuracy over these wide impedance ranges using two new techniques:

- (1) Direct impedance using the RF I-V method

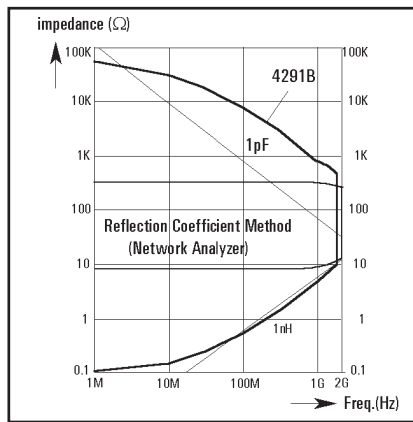


Figure 1. Ranges of impedance measurement (Accuracy of measurement: 10%)

The analyzer uses a new method to measurement impedance; the RF I-V method. Figure 2 shows the basic principles of the RF I-V method and the reflection coefficient method (conventional method using a vector network analyzer).

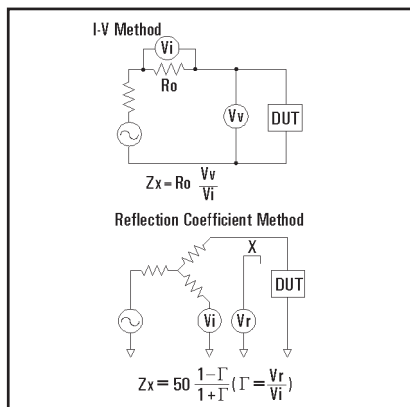


Figure 2. I-V method and reflection coefficient impedance method

As can be seen from the equations in Figure 2, the RF I-V method measures impedance directly, while the reflection coefficient method measures reflection coefficient and converts it to impedance. In both the RF I-V and reflection coefficient methods, the impedance is given by ratios of the readings of the two voltmeters. Therefore, one would expect that the accuracy of both methods would be similar. However, the

reflection coefficient method magnifies the measurement error when converting reflection coefficient to impedance. As impedance goes away from 50 Ω, Figure 3 shows that a small change in reflection coefficient value produces a large change in impedance. In other words, a small error in the reflection coefficient leads to a large error in impedance. (For example, for an impedance of 2 kΩ, a 1% error of reflection coefficient results in a 24% error in impedance.)

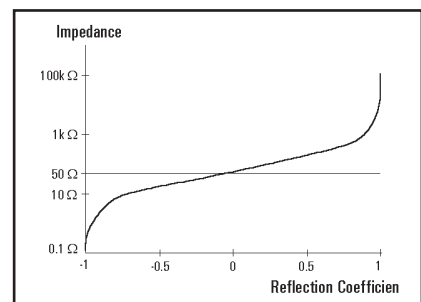


Figure 3. Relationship between impedance and reflection coefficient

The RF I-V method measures impedance directly from a ratio of voltage and current, without converting the measured data. Therefore, the RF I-V technique maintains consistent accuracy even if the impedance is significantly larger or smaller than 50 Ω. Thus, for measuring non-50 Ω components, the 4291B using the RF I-V technique is recommended.

(2) High/Low impedance circuit

The 4291B employs high-impedance and low-impedance circuits, as shown in Figure 4, to expand the range of impedance measurements. When measuring a high-impedance device, accurate measurement of the DUT current is most critical. The high-impedance circuit solves this problem by connecting the current detection circuit directly in series with the DUT, ensuring accurate DUT current measurement and not measuring current flowing in the voltage sensing circuit.

On the other hand, when measuring a low-impedance device, the voltage across the DUT is most critical. In this case, the low-impedance circuit connects the voltage detection circuit directly to the DUT, ensuring accurate voltage measurement and not measuring the voltage drop from the current sensing impedance. By using the right measuring circuit for the impedance being measured, it is possible to extend the range of impedance magnitudes measured for a given accuracy.

These two measuring circuits are implemented as two different test heads in the 4291B, so users can select and switch the circuits easily to optimize the measurement range and accuracy to best match the DUT impedance.

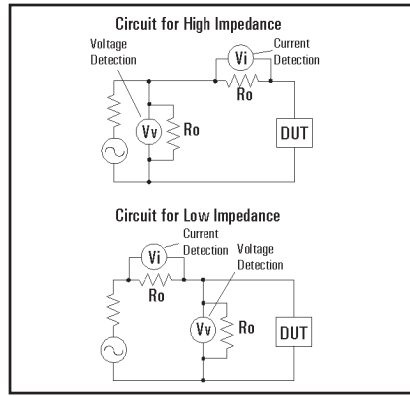


Figure 4. High-impedance and low-impedance measuring circuits

High Accuracy Q and D measurements

As shown in Figure 5, the Agilent 4291B is capable of evaluating a sample with $Q = 100$ within $\pm 15\%$ accuracy at 1 GHz. This capability is targeted for evaluating the loss of low-loss components at RF. The accuracy of Q and D measurements depends on the accuracy of phase measurement. (See Figure 6.)

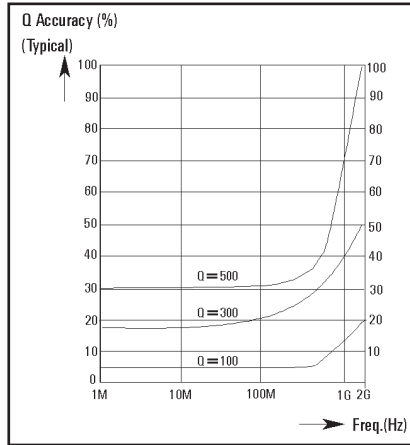


Figure 5. Q measurement accuracy

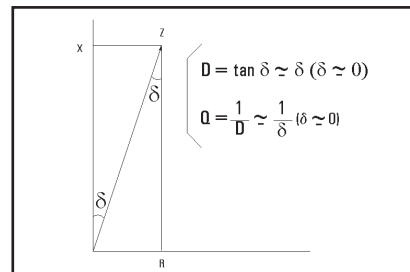


Figure 6. Q and D measurement and phase measurement

The 4291B improves the accuracy of phase measurements by requiring an additional phase calibration step. (Conventional one-port calibration uses open, short, and 50 Ω load standards.) This type of one-port calibration does not provide satisfactory accuracy for phase measurements because of the phase uncertainty of the 50 Ω standard. By using a low-loss air capacitor as a phase standard, the 4291B lowers the phase uncertainty to 1 mrad or less (corresponding to $D = 0.001$), ensuring improved accuracy for Q and D measurements.

Error-free 1.8-m cable extension

The cable connecting the main body of the 4291B to the test station has been extended to 1.8 meters without adding additional errors. (See Figure 7.) The long cable allows easy access to remote DUT locations; for example a temperature chamber, scanner/handler, or custom test setup.

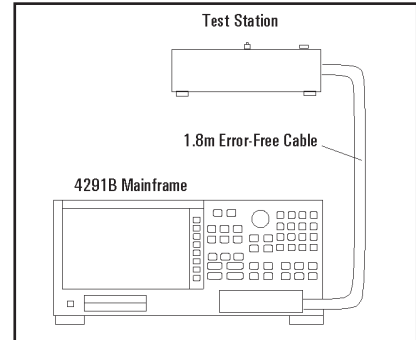


Figure 7. Extended test cable

Normally, extending the cable increases measurement error due to increased noise, temperature differentials, cable resistance, etc. However, as illustrated in Figure 8, the 4291B measures the current and voltage signals using the same circuit and alternates the measurement with fast time-division multiplexing.

Because the measurements are made at alternating intervals of several milliseconds, the same cable-induced errors occur in both the current and voltage measurement data. These errors are then canceled out when the impedance is obtained from the ratio of voltage to current.

Furthermore, by measuring current and voltage with the same circuit, errors in the measuring instrument caused by temperature changes are offset in the same manner, resulting in significant temperature characteristic improvement.

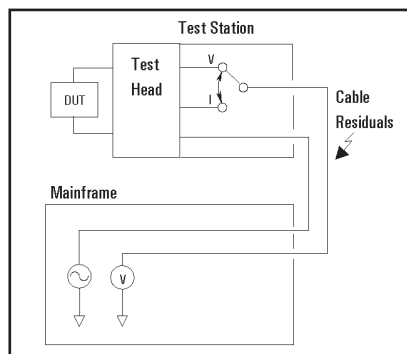


Figure 8. Time-division multiplexing of current/voltage measurement

Test fixtures have a large impact on measurement accuracy, especially at higher test frequencies. An important factor in getting accurate measurements is eliminating errors introduced by the DUT fixturing. Calibration ensures high accuracy at the plane of calibration (measurement point where the standards are applied), but in actual practice, the test fixture can add additional error terms beyond the calibration plane. This is why fixture error compensation is so important.

Electrical properties of the test fixture (which occur after the calibration point) consist of a phase rotation due to the physical length of the electrodes and other unwanted stray parasitics between electrodes.

Both of these can cause significant measurement errors in the RF band. Conventional measuring instruments often have no convenient method to eliminate them effectively. The 4291B uses electrical length compensation to remove the errors caused by phase rotation and OPEN/ SHORT compensation (at the DUT location in the fixture) for removing fixture parasitic impedance.

Conclusion

The Agilent 4291B RF impedance and material analyzer provides highly accurate impedance and material measurements by incorporating new technologies and offering an integrated package including a family of SMD and material fixtures. The analyzer overcomes many limitations of conventional impedance analysis and, for the first time, provides an efficient and accurate measurement solution for passive component analysis over the RF range.

For detailed technical information on the Agilent 4291B RF impedance and material analyzer, refer to the Agilent 4291B Technical Information beginning on page 5.

For more information, request following literatures from your local Agilent representative:

Agilent 4291B 1.8 GHz Impedance/Material Analyzer, Product Overview, P/N 5966-1501E

Agilent 4291B, Data Sheet, P/N 5966-1543E

Highly Accurate Evaluation of Chip Capacitors using the Agilent 4291B, Application Note 1300-1, P/N 5966-1850E

Evaluating Chip Inductors using the 4291B, Application Note 1300-2, P/N 5966-1848E

Permittivity Measurements of PC Board and Substrate Materials using the Agilent 4291B and 16453A, Application Note 1300-3, P/N 5966-1847E

Permeability Measurements using the Agilent 4291B and 16454A, Application Note 1300-4, P/N 5966-1844E

Electronic Characterization of IC Package, Application Note 1300-5, P/N 5966-1849E

Impedance Characterization of Magneto-Resistive Disk Heads Using the Agilent 4291B Impedance/Material Analyzer, Application Note 1300-6, P/N 5966-1096E

On-Chip Semiconductor Device Impedance Measurements Using the Agilent 4291B, Application Note 1300-7, P/N 5966-1845E

Evaluating Temperature Characteristics using a Temperature Chamber and the Agilent 4291B, Product Note 4291-2, P/N 5966-1927E

Impedance Measurements Using the Agilent 4291B and the Cascade Microtech Prober, Product Note 4291-3, P/N 5966-1928E

Dielectric Constant Evaluation of Rough-Surfaced Materials, Product Note 4291-5, P/N 5966-1926E

APPENDIX

New Technologies used in the High Frequency Impedance Analyzer Agilent 4291B RF Impedance/ Material Analyzer Technical Information

Abstract

A new one-port impedance analyzer has been developed for analysis of high frequency devices and materials up to 1.8 GHz.

Traditionally, impedances near 50 Ω have been measured accurately by the null method using a directional bridge. However, this new analyzer uses a voltmeter/ammeter method and offers precise measurement capability over a wide impedance range. Furthermore, a special calibration method using a low-loss capacitor realizes an accurate high-Q device measurement.

This paper describes the advantages of these techniques. Impedance traceability of the instrument will also be discussed. Finally, many types of test fixtures are introduced, because they are a key element in any test system.

1. Introduction

A general impedance measurement schematic using two vector voltmeters is shown in Fig. 1. In this case, the true impedance (Z_x) of a device under test (DUT) is determined by measuring the voltages of any two different sets of points in a linear circuit.

$$Z_x = K1 \times \frac{K2 + V_r}{1 + K3 \times V_r} \quad (1)$$

where:

$K1, K2, K3$ = complex constant
 V_r = voltage ratio

In the process of deriving the equation above, linearity is assumed but reciprocity is not assumed. Therefore, the existence of active devices in the circuitry is not prohibited.

There are, at most, three unknown parameters related to the circuit in Equation 1. Once we know these parameters, we can calculate any impedance of the DUT from the measured voltage ratio (V_r).

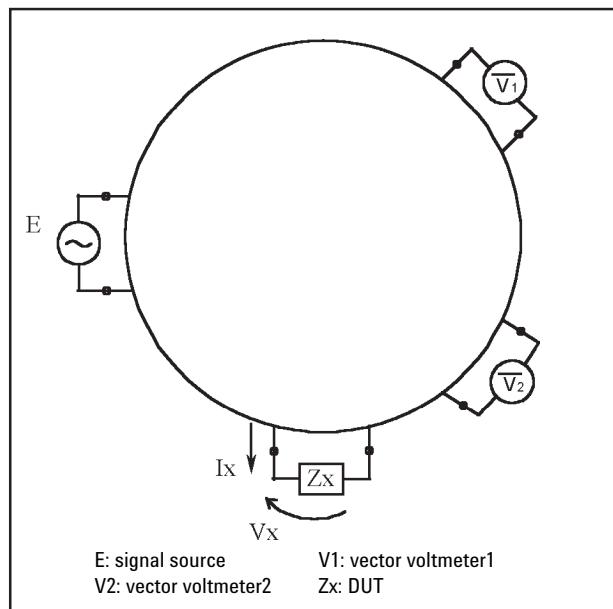


Figure 1. General schematic for impedance measurement using two vector voltmeters

The procedure that estimates these circuit 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.”

2. Transducer

We call a linear circuit that relates a signal source, two vector voltmeters, and a DUT (see Figure 1) a “transducer.” Transducers are the key element in impedance measurement. For example, two types of transducers, the “directional bridge” and the transducer in a “voltmeter/ammeter (V-I) method,” are compared in terms of sensitivity to the gain variance.

2-1. Directional bridge type

Directional bridges (see Fig. 2-1) are used in many network analyzers.

In this case, the bilinear transformation is

$$Z_x = -R_o \times \frac{1+\Gamma}{1-\Gamma} \quad (2)$$

where:

$$\Gamma = \frac{(Z_x - R_o)}{(Z_x + R_o)} : \text{reflection coefficient}$$

$$V_r = \frac{V_2}{V_1} = \left(-\frac{1}{8}\right) \times \Gamma$$

$R_o = 50 \Omega$: characteristic impedance

Parameters corresponding to those in Equation 1 are

$$K1 = -R_o \quad K2 = 1 \quad K3 = -1$$

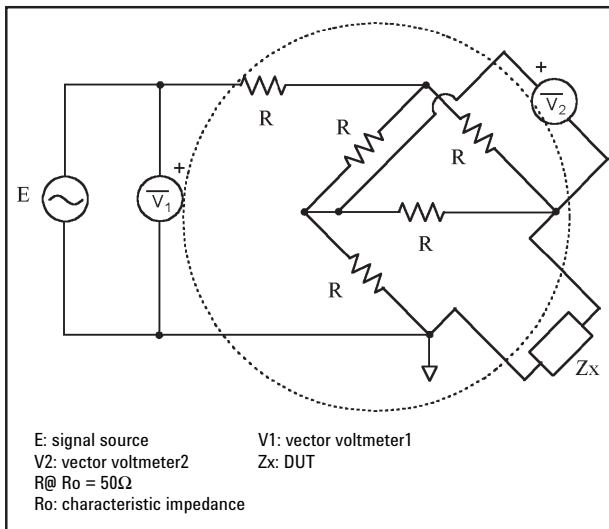


Figure 2-1. Directional bridge circuit

2-2. V-I method

Fig. 2-2 shows the simplest transducer in the V-I method. The bilinear transformation in this case is

$$Z_x = R_o \times V_r \quad (3)$$

where:

$R_o = 50 \Omega$: resistor that converts DUT current to a voltage

$$V_r = \frac{V_2}{V_1}$$

Parameters corresponding to those in Equation 1 are

$$K1 = R \quad K2 = 0 \quad K3 = 0$$

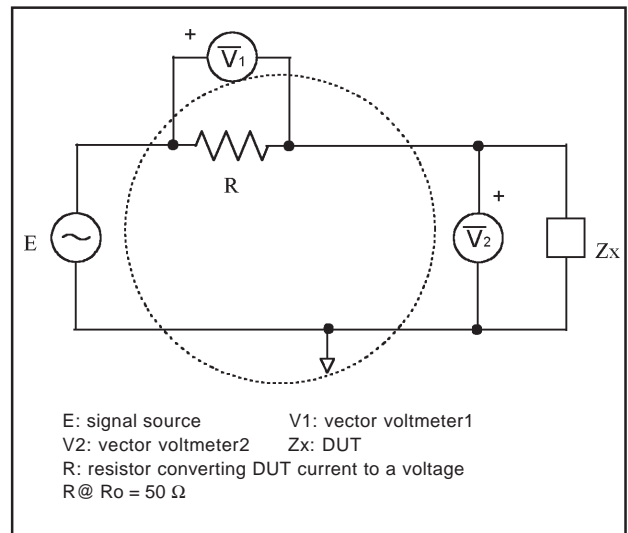


Figure 2-2. V-I method

2-3. Sensitivity to gain variance

In this section we discuss the relationship between the impedance measurement error and the gain variance of the vector voltmeters.

2-3-1. Directional bridge type

We assume that the vector voltmeters in Fig. 2-1 are not ideal but have some gain variance. In this situation, the measured voltages (V_1 and V_2) and the calculated impedance (Z_x) are:

$$V_1 = E \times \alpha_1$$

$$V_2 = \left(-\frac{1}{8}\right) \times E \times \Gamma \times \alpha_2$$

$$Z_x = -R_o \times \frac{1+\Gamma}{1-\Gamma}$$

where:

α_1 = gain of the vector voltmeter1

α_2 = gain of the vector voltmeter2

$\Gamma = -8 \times V_r \times \alpha_r$: measured reflection coefficient

$$V_r = \frac{V_2}{V_1} : \text{voltage ratio}$$

$$\alpha_2 = \frac{\alpha_2}{\alpha_1} : \text{ratio of voltmeter gains}$$

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

$$S = Z \frac{\frac{\delta Z_x}{Z_x}}{\frac{\delta \alpha_r}{a_r}} \quad (4)$$

This sensitivity can be considered as the inverse of the "magnification on gain variance." As S becomes smaller the error of the calculated impedance also becomes smaller. In this case, Equation 4 is given as follows:

$$S = \frac{\delta Z_x}{\delta \Gamma} \times \frac{\delta \Gamma}{\delta \alpha_r} \times \frac{\alpha_r}{Z_x}$$

$$= \left(\frac{1}{2}\right) \times \frac{Z_x^2 - R_o^2}{Z_x \times R_o}$$

$$\left(-\frac{1}{2}\right) \times \left(\frac{R_o}{Z_x}\right) \text{ for } |Z_x| \ll R_o$$

$$= 0 \quad \text{for } |Z_x| = R_o$$

$$\left(\frac{1}{2}\right) \times \left(\frac{Z_x}{R_o}\right) \text{ for } |Z_x| \gg R_o$$

This implies

- 1) This type of transducer has little sensitivity to the voltmeter gain variance when the DUT impedance is near R_o (50 Ω).
- 2) The gain variance of the voltmeter behaves as the offset impedance with a magnitude of $(1/2) \cdot R_o \cdot |\Delta \alpha_r / \alpha_r|$ when the DUT impedance is far smaller than R_o .
- 3) The gain variance of the voltmeters behaves as the offset admittance with a magnitude of $(1/2) \cdot G_o \cdot |\Delta \alpha_r / \alpha_r|$ when the DUT impedance is far larger than R_o .

where:

$$\Delta \alpha_r = \text{change in gain ratio } \alpha_r$$

$$G_o = 1/R_o$$

Fig. 2-3 shows this characteristic.

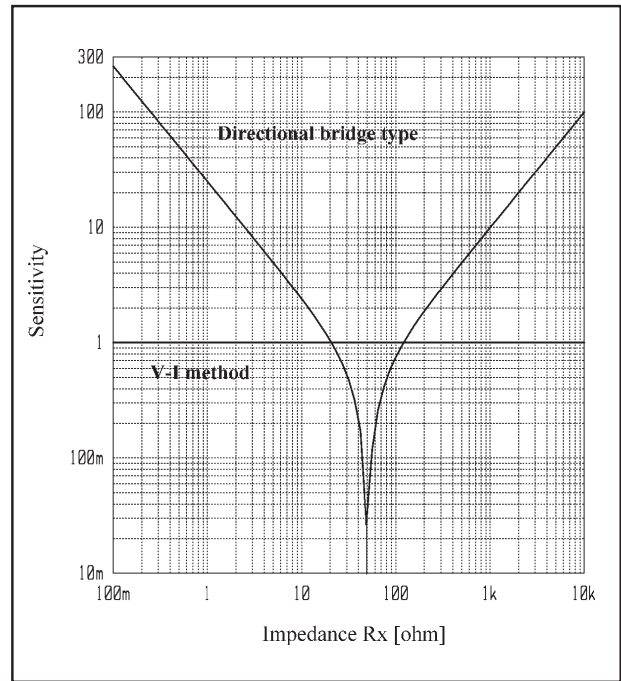


Figure 2-3. Sensitivity to the voltmeter gain variance

2-3-2. V-I method

We also assume that there is some gain variance in the vector voltmeters in Fig. 2-2. The relationship among measured voltages (V_1 and V_2) and the calculated impedance (Z_x) are:

$$V_1 = E \times \frac{R_o}{Z_x + R_o} \times \alpha_1$$

$$V_2 = E \times \frac{Z_x}{Z_x + R_o} \times \alpha_2$$

$$Z_x = R_o \times V_r \times \alpha_r$$

where:

$$\alpha_1 = \text{gain of the vector voltmeter 1}$$

$$\alpha_2 = \text{gain of the vector voltmeter 2}$$

$$V_r = \frac{V_2}{V_1} : \text{voltage ratio}$$

$$\alpha_r = \frac{\alpha_1}{\alpha_2} : \text{ratio of voltmeter gains}$$

In this case, the sensitivity is given by the following:

The error ratio ($\Delta Z_x / Z_x$) is always constant and equal to unity. For example, if the voltmeter gain

ratio ar changes 1%, an impedance error of 1% is caused for any DUT.

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

3. Schematic of new RF impedance analyzer

From the discussion in Section 2, we find that the voltmeter gain variance is neither suppressed nor magnified for all DUT impedances in a “V-I method type transducer.” This characteristic is desirable for wide impedance measuring capability. Therefore we adopted this type of transducer for our new one-port RF impedance meter.

3-1. Basic transducer circuit

Fig. 3-1 shows the basic circuit of the transducer in New RF Impedance Meter, which is modified from Fig. 2-2. The high impedance configuration realizes perfect OPEN and imperfect SHORT conditions. On the other hand, the low impedance configuration realizes imperfect OPEN and perfect SHORT conditions.

Moreover, for this circuit the output impedance at the DUT port is always R_o in either switch position.

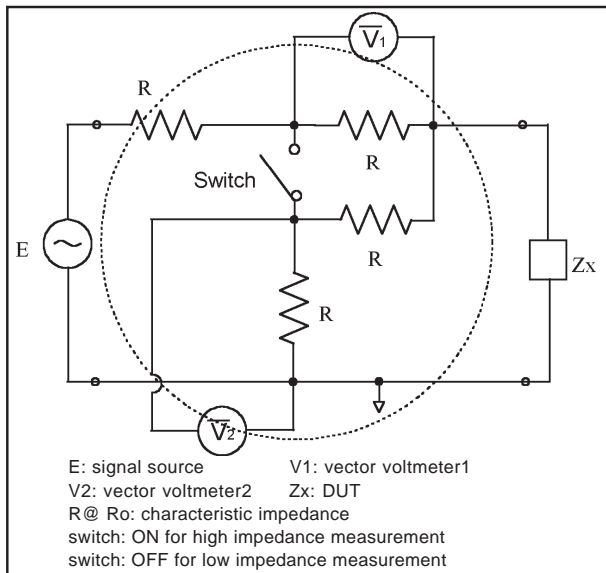


Figure 3-1. Basic transducer circuit in New RF Impedance Analyzer

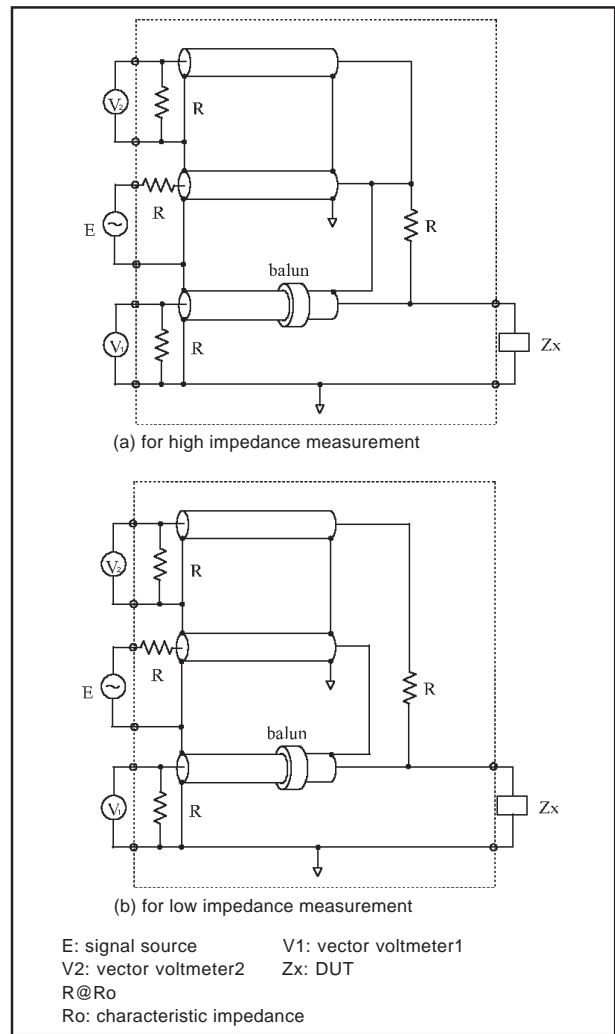


Figure 3-2. The actual transducer circuit

3-2. Actual transducer circuit

Actual transducer circuits are shown in Fig. 3-2 (a) and Fig. 3-2 (b).

We took the following into account when designing the actual transducer circuits:

- 1) Because a wideband switch with small non-linearity and with small transients over a wide signal range is not easily realized, we divided the circuit in Fig. 3-1 into two separate circuits.
- 2) Because the minimum frequency of the new meter is 1 MHz, the floating voltmeter (V1),

which corresponds to the current meter, is easily realized by using a balun.

- 3) In order to prevent the error caused by by-pass current we adopted a circuit in which the exiting impedance of the balun is parallel connected not to the current meter (V1) but to the signal source (E).

3-3. Simplified block diagram of new impedance analyzer

The followings are other key features of the instrument:

- 1) Time division multiplex—
Two voltmeters are obtained by time division multiplexing one voltmeter. The multiplexing period is 2 msec. This ensures that the slow drift of the voltmeter gain does not affect the impedance measurement. The signal path after the multiplexer can be extended. This instrument 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. Good temperature characteristics are also derived, even with an extended cable, by the single path configuration.
- (2) Impedance ranging—
At frequencies below 200 MHz there exists an “expand range.” In the expand range there is a gain difference between the voltage channel and the current channel in the earlier stage of the multiplexer. This impedance ranging offers stable measurements for DUTs with impedances that differ greatly from 50 Ω.

Fig. 3-3 (a) and Fig. 3-3 (b) show the typical measurement errors for impedance of the instrument. The errors consist mainly of the uncertainties for standards (STDs) used in the calibration, non-repeatabilities in connections, temperature coefficients, noise, and errors in the interpolations. The impedance phase errors can be reduced by using a special calibration. (See Section 4.)

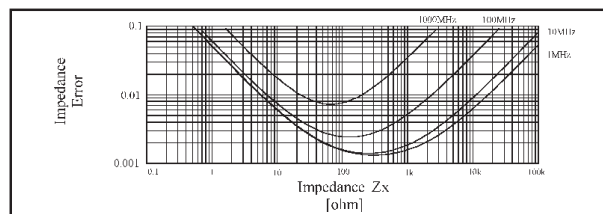


Figure 3-5. (a) Typical errors for impedance magnitude with the transducer for high impedance

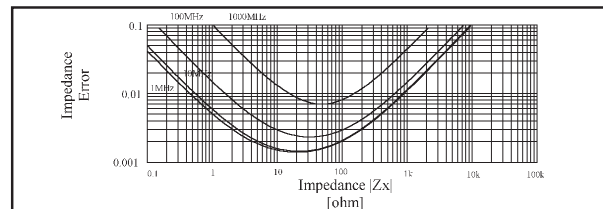


Figure 3-5. (b) Typical errors for impedance magnitude with the transducer for low impedance

4. A special calibration for high Q measurement

Normally the accuracy requirement for the impedance phase is greater than that of the impedance magnitude. Our new impedance meter has a special, but easy-to-use, calibration for high Q (quality factor) device measurements. We discuss this calibration technique in this section.

4.1 Outline of the special calibration

Even if the stability of the instrument is good enough, accurate Q measurements are not performed without correct markings on the phase scale of the instrument. For instance, if we want to measure the Q factor with 10% uncertainty for a DUT whose Q value is almost 100, the uncertainty for phase scaling must be smaller than 1E-3. The phase accuracy of the instrument is determined almost entirely by the uncertainty of the 50 Ω LOAD STD used in the OSL calibration. One method to improve phase measurement accuracy is to use a phase calibrated LOAD STD. However, it is not ensured that phase uncertainty for a calibrated 50 Ω LOAD is smaller than 1E-3 at high frequencies (such as 1 GHz).

In addition to the normal OPEN-SHORT-LOAD STDs, using a low-loss air-capacitor as the second LOAD (LOAD2), whose dissipation factor (D) is kept below 1E-3 at around 1 GHz, also offers another feature; the uncertainty for the measured

phase is decreased from the phase uncertainty of the 50 Ω LOAD (LOAD1) to the uncertainty of D of the low-loss capacitor (LOAD2) for almost all DUT impedances.

4.2 Details of the modified OSL calibration using an additional load

We want to have the calibration method that reduces the error in phase measurement in spite of the existence of phase error for the 50 Ω LOAD. Consider a case where we have the 50 Ω LOAD STD whose impedance phase is not known but whose impedance magnitude is known. In this situation how about adding another LOAD (LOAD2) whose impedance magnitude is not known but whose impedance phase is known? We use a low-loss capacitor as the second LOAD. The number of unknown circuit parameters are still three at most. However, two more unknowns related to STDs are added. Let us define the problem. There are eight real unknown parameters:

- circuit parameter K1 (two real parameters)
- circuit parameter K2 (two real parameters)
- circuit parameter K3 (two real parameters)
- impedance phase θ_{ls1} of the 50 Ω LOAD (one real parameter)
- impedance magnitude Z_{abs_ls2} of the low-loss capacitor LOAD2 (one real parameter)

We solved this problem analytically. For the simplest case where both the OPEN and the SHORT STD are ideal, the three circuit parameters are found as follows

$$K1 = A \times Z_{ls} \times R_o \quad (5)$$

$$K2 = \frac{-Z_{sm}}{R_o}$$

$$K3 = -Y_{sm} \times R_o$$

where:

R_o = characteristic impedance

$A = (1 - Z_{lm_i} * Y_{om}) / (Z_{lm_i} - Z_{sm})$

Y_{om} = measured admittance for OPEN STD

Z_{sm} = measured impedance for SHORT STD

Z_{lm_i} = measured impedance for LOAD STD

(i = 1: LOAD 1, i = 2 : LOAD2)

Z_{ls_i} = true impedance for LOAD STD

(i = 1: LOAD 1, i = 2 : LOAD2)

$Z_{ls1} = Z_{abs_ls1} * \text{EXP}(j * \theta_{ls1})$

$Z_{ls2} = Z_{abs_ls2} * \text{EXP}(j * \theta_{ls2})$

$\theta_{ls1} = \theta_2 - \theta_1 + \theta_{ls2}$

$Z_{abs_ls2} = A1/A2 * Z_{abs_ls1}$

Z_{abs_ls1} = impedance magnitude for LOAD 1 (50 Ω) : known

θ_{ls2} : impedance phase for LOAD2 (low-loss capacitor) : known

$\theta_1 = \arg((1 - Z_{lm_1} * Y_{om}) / (Z_{lm_1} - Z_{sm}))$

$\theta_2 = \arg((1 - Z_{lm_2} * Y_{om}) / (Z_{lm_2} - Z_{sm}))$

$A1 = |(1 - Z_{lm_1} * Y_{om}) / (Z_{lm_1} - Z_{sm})|$

$A2 = |(1 - Z_{lm_2} * Y_{om}) / (Z_{lm_2} - Z_{sm})|$

For actual cases, these circuit parameters are expressed by far more complicated equations. Therefore, we adopted a simpler procedure consisting of two steps:

Step 1–

- Regard the impedance of the 50 Ω LOAD as $Z_{ls1} = 50 + j * 0$ (that is, the phase of LOAD1 is zero).
- Find the circuit parameters K1, K2, and K3 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–

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

Although this is an approximate method, actually just performing the two steps is accurate enough for our purpose. We call this method the “modified OSL calibration.”

4.3 Phase measurement error using modified OSL calibration

We considered the following error factors during phase measurement in the modified OSL calibration:

- 1) uncertainty for impedance magnitude of LOAD 1
- 2) impedance phase of LOAD1
- 3) impedance magnitude of LOAD2
- 4) uncertainty for impedance phase of LOAD2
- 5) uncertainty for admittance magnitude of OPEN

Notice that Factor 2 and Factor 3 do not cause errors if we use the analytical solution. With computer simulations the phase error is shown to behave as follows:

- 1) Sensitivity of phase measurement error due to the uncertainty of impedance magnitude for LOAD1 is small.
- 2) Sensitivity of phase measurement error due to the impedance phase of LOAD1 is small.
- 3) Sensitivity of phase measurement error due to the impedance magnitude of LOAD2 is small.
- 4) Uncertainty for the impedance phase of LOAD2 directly affects the phase measurement error.
- 5) In the case of reactive DUT sensitivity of the phase measurement error due to the uncertainty for admittance magnitude of OPEN ($|\Delta Y_{open}|$) is reduced to $|R_o \cdot \Delta Y_{open}| \cdot (C_{open}/C_{ls2})$. In the case of resistive DUTs, the sensitivity is the same as in the normal OSL calibration.

where:

R_o = characteristic impedance = 50 Ω

C_{open} = capacitance of OPEN STD

C_{ls2} = capacitance of LOAD2 (low-loss capacitor)

For example, the relationship between the phase measurement error and the uncertainty for the impedance phase of LOAD2 ($\Delta\theta_{ls2}$) is shown in Fig. 4-1. Fig. 4-2 shows the relationship between the phase measurement error and the uncertainty for admittance magnitude of OPEN (ΔY_{open}).

From the above discussion, the phase measurement error, when using the modified OSL calibration, is mainly determined by

- uncertainty for the impedance phase of LOAD2
- uncertainty for the admittance magnitude of OPEN

Now we evaluate these two items. The D factor for the capacitor (3pF) used in the calibration can be small because its dimensions are small and the space between the inner and outer conductors is filled almost completely with air. The D value is verified as 500E-6 at 1 GHz from the residual resistance measurement at the series resonant frequency. The D factor is in proportion to $\text{Freq}^{1.5}$ due to the skin effect. By using zero as the D value for the capacitor during calibration, a phase measurement error of 500E-6 is realized at 1 GHz. On the other hand, the uncertainty for the OPEN capacitance is +/-5fF at most.

This leads to a phase measurement error less than +/-100E-6 at 1 GHz. In all, a phase measurement uncertainty of 500E-6 is realized by using the modified OSL calibration.

5. Impedance traceability

For an impedance performance check using the top-down method, we set up a kit traceable to the National Standards. This kit is calibrated annually at our standards laboratory. Two major items in the kit are the 50 Ω LOAD and 10 cm long, 50 Ω beadless air line. The 50 Ω LOAD is desirable because its frequency characteristic for impedance is very flat. The structure of the air line is very simple. Therefore, it is easy to predict its frequency characteristic and it is convenient to realize various impedances by changing frequencies with the OPEN or SHORT terminated.

The traceability path for the kit is shown in Fig. 5. The impedance characteristic of OPEN ended and SHORT ended air lines can be calculated theoretically from their dimensions and resistivity.

However, it is not easy to design a system offering calibrated dimensions with the individual kit.

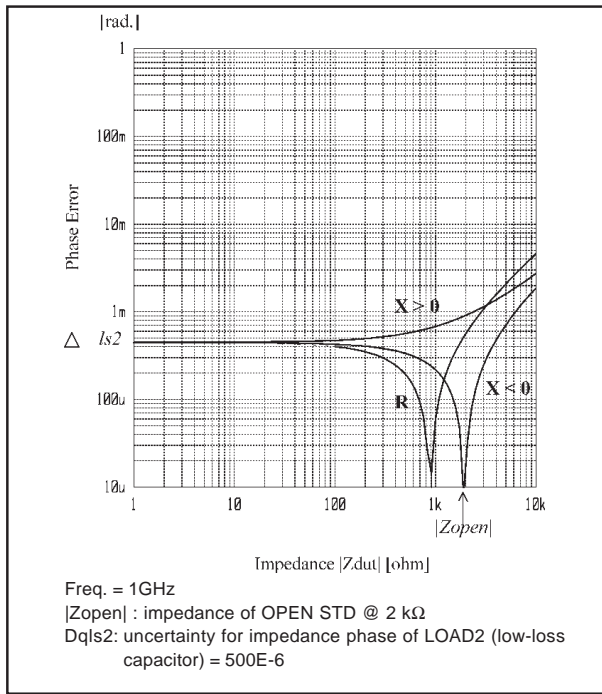


Figure 4-1. Relationship between the phase measurement error and the uncertainty for the impedance phase of LOAD2

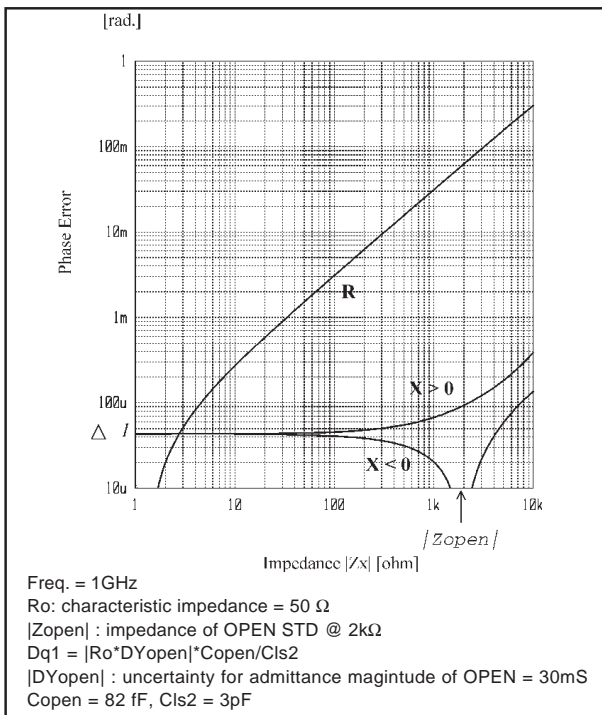
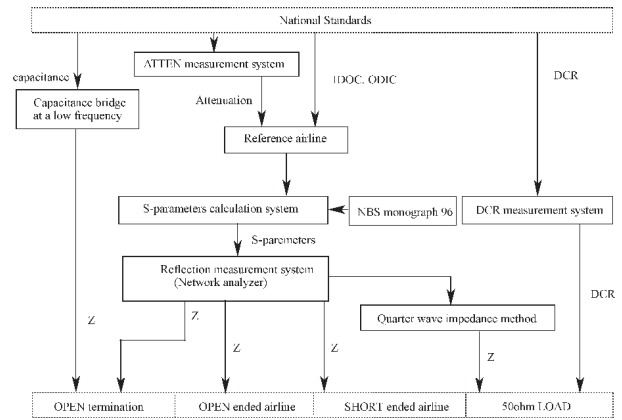


Figure 4-2. Relationship between the phase measurement error



Therefore, the periodical calibration of dimensions is performed only on the reference air line of the standard laboratory. Calibration for the individual air line is executed by the network analyzer calibrated from the reference air line. The 50 Ω LOAD calibration is done mainly by the quarter wave impedance method and DCR 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 an ideal one. Its uncertainties are the skin effect and non-repeatabilities.

6. Test fixtures

In the actual measurement test, fixtures corresponding to different shaped DUTs are needed. As the frequency range goes higher, fixtures are needed that are able to handle smaller devices. We have developed four such fixtures:

- a fixture for surface mount devices (SMDs) with bottom electrodes
- a fixture for SMDs with side electrodes
- a fixture for very small SMDs
- a fixture for lead components

To reduce the error at the fixture terminal it is necessary to minimize the length from the reference 7 mm plane to the fixture terminal and to minimize the connection non-repeatability.

The new fixtures improved repeatability by almost five times compared with our old ones. The typical non-repeatability of the SMD fixtures is

- +/- 50 pH, +/- 30 mΩ for SHORT measurement
- +/- 5 fF, +/- 2 μS for OPEN measurement

Furthermore, we installed “fixture compensations” in the firmware that correspond to the corrections at the reference plane. This reduces the errors generated in the circuit between the reference plane and the fixture terminal. The best choice for the compensation is the OSL. However, it is not easy to prepare a LOAD having excellent frequency characteristics. We prepare the following more realistic compensation functions:

- fixture port extension
- OPEN-SHORT correction at the fixture plane

Fig. 6 shows the typical additional error when using the above corrections at the same time. These values are almost three times better than the error values for our former types of fixtures.

7. Conclusion

Selection of “transducers” is important for accurate impedance measurement. A new type of transducer, based on the voltmeter/ammeter method and with wide impedance measuring capability, is proposed. By adopting these transducers, the New RF Impedance Analyzer has been developed. We also proposed a new phase calibration technique modified from the OSL calibration. It utilizes a low-loss capacitor as the second LOAD. This calibration enables accurate Q measurements.

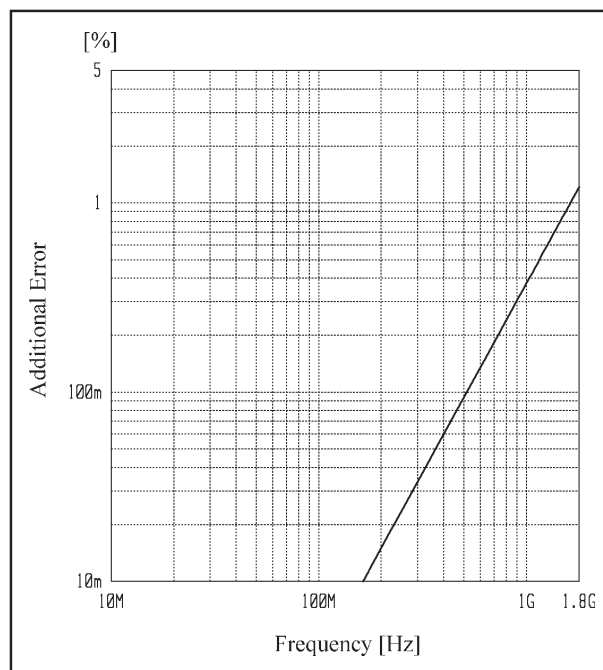


Figure 6. Typical additional errors after fixture compensations

References

Agilent Technologies, *Traceability and the 8510 Network Analyzer*, Nov. 1 1985

Robert E. Nelson, Marlene R. Coryell, *Electrical Parameters of Precision, Coaxial, Air-Dielectric Transmission Lines*, NBS Monograph96, June 30 1966

Agilent Technologies' Test and Measurement Support, Services, and Assistance

Agilent Technologies aims to maximize the value you receive, while minimizing your risk and problems. We strive to ensure that you get the test and measurement capabilities you paid for and obtain the support you need. Our extensive support resources and services can help you choose the right Agilent products for your applications and apply them successfully. Every instrument and system we sell has a global warranty. Support is available for at least five years beyond the production life of the product. Two concepts underlie Agilent's overall support policy: "Our Promise" and "Your Advantage."

Our Promise

"Our Promise" means your Agilent test and measurement equipment will meet its advertised performance and functionality. When you are choosing new equipment, we will help you with product information, including realistic performance specifications and practical recommendations from experienced test engineers. When you use Agilent equipment, we can verify that it works properly, help with product operation, and provide basic measurement assistance for the use of specified capabilities, at no extra cost upon request. Many self-help tools are available.

Your Advantage

"Your Advantage" means that Agilent offers a wide range of additional expert test and measurement services, which you can purchase according to your unique technical and business needs. Solve problems efficiently and gain a competitive edge by contracting with us for calibration, extra-cost upgrades, out-of-warranty repairs, and on-site education and training, as well as design, system integration, project management, and other professional services. Experienced Agilent engineers and technicians worldwide can help you maximize your productivity, optimize the return on investment of your Agilent instruments and systems, and obtain dependable measurement accuracy for the life of those products.

By internet, phone, or fax, get assistance with all your test and measurement needs.

Online Assistance

www.agilent.com/find/assist

Phone or Fax

United States:
(tel) 1 800 452 4844

Canada:
(tel) 1 877 894 4414
(fax) (905) 282 6495

Europe:
(tel) (31 20) 547 2323
(fax) (31 20) 547 2390

Japan:
(tel) (81) 426 56 7832
(fax) (81) 426 56 7840

Latin America:
(tel) (305) 269 7500
(fax) (305) 269 7599

Australia:
(tel) 1 800 629 485
(fax) (61 3) 9210 5947

New Zealand:
(tel) 0 800 738 378
(fax) (64 4) 495 8950

Asia Pacific:
(tel) (852) 3197 7777
(fax) (852) 2506 9284

Product specifications and descriptions in this document subject to change without notice.

Copyright © 1998, 2000 Agilent Technologies
Printed in U.S.A. 11/00
5966-2046E



Agilent Technologies

Innovating the HP Way