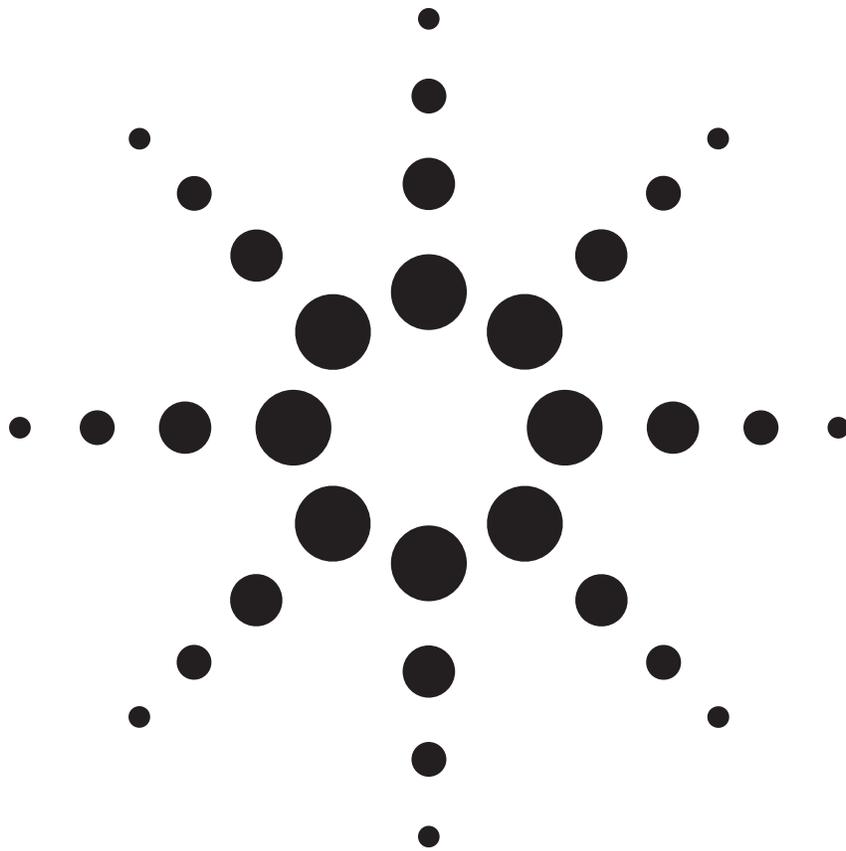# Agilent
How To Accurately Evaluate
Low ESR, High Q RF Chip Devices

Application Note 1369-6

Agilent Technologies

Contents

The Changing Requirements of RF Component Testing

Several trends in wireless communications are changing the requirements for RF component testing. Today's hottest new applications—including 3G UMTS (W-CDMA), *Bluetooth™*, and wireless LAN applications—are being offered at the 2 – 3 GHz frequency bands. Design engineers therefore must evaluate the RF components used in wireless communication devices at these higher operating frequencies, under realistic conditions, to ensure the quality and capability of their final products.

While the operating frequencies of wireless devices are moving higher, device size is shrinking and so is the size of the device's RF components. For example, chip inductors as small as 0603 (0.6 x 0.3 mm or 0.2 x 0.1 inch) are being used today in mobile phones and PDAs.

Smaller, higher frequency devices require higher performing RF circuits and more efficient battery operation. To meet increasingly stringent design requirements, design engineers use passive components such as chip inductors and capacitors with low equivalent series resistance (ESR) and high quality factor (high Q). Because component selection is so important in determining the final performance of a product, component parameters must be characterized as completely and accurately as possible.

Proper measurement techniques, the right test instrument, and fixtures that can accommodate the minute size of today's RF chip devices are all necessary to enable thorough, accurate RF characterization and testing. This application note discusses how these needs are met with a new impedance measurement solution.

Measurement challenges

As the operating frequency of an RF component increases, so the self-resonant frequency (SRF) gets higher. Furthermore, impedance at the SRF becomes large. To characterize the component, we must be able to measure across a broad impedance range. The common practice of using a vector network analyzer and the reflection coefficient measurement technique gives very accurate results at 50 or 75 ohms. However, it does not cover the higher and lower impedance areas. This situation is illustrated in Figure 1.
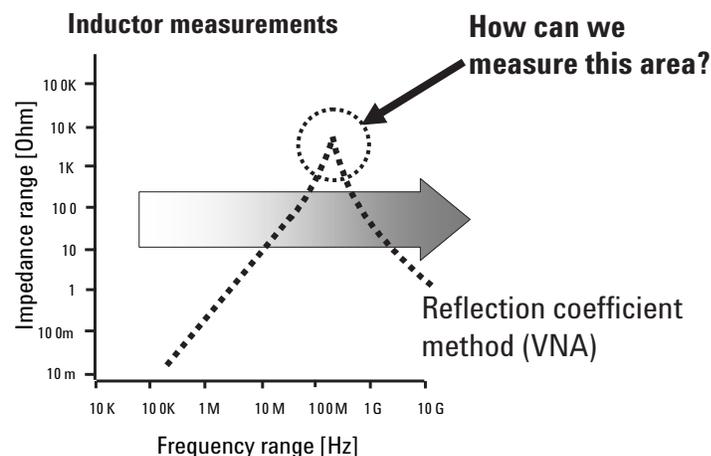
Figure 1. The reflection coefficient method does not adequately measure the impedance range of a modern chip inductor.

Better results can be obtained with the Agilent E4991A RF impedance/ material analyzer, which uses a proprietary RF I-V technique for highly accurate impedance measurement to 3 GHz. Unlike the reflection coefficient method, RF I-V directly measures current and voltage (I-V) through the device under test, achieving an impedance measurement accuracy of ±0.8% over a wide impedance and frequency range. The area of coverage is shown in Figure 2.
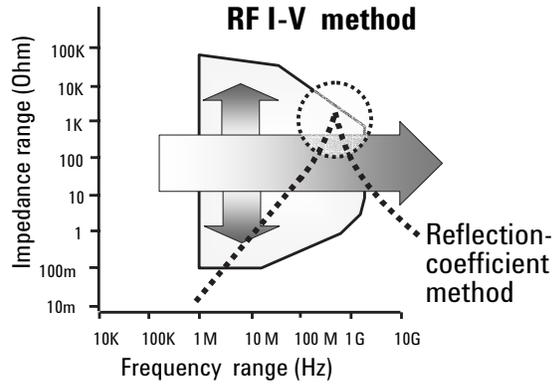
Figure 2. The RF I-V method used by the Agilent E4991A RF impedance analyzer extends the possibility of accurate impedance measurement across the RF range.

The E4991A is calibrated using a low-loss capacitor standard for improved phase accuracy. As a result, it can measure components with small capacitance and inductance values over the frequency range, and offers unique advantages in characterizing high Q devices and evaluating temperature characteristics.

A variety of test fixtures for different components and materials are available for the E4991A, with options for material measurement, temperature characteristic evaluation, and on-wafer measurement.

Advantages of the RF I-V Method

To test today's higher frequency RF chip devices, the RF I-V method employs a simple operating principle, illustrated in Figure 3. A test signal is applied from the signal source to the device under test (DUT). A vector voltage meter measures the voltage (V_1) across the DUT. At the same time, current (I) flowing through the DUT is picked up by a balun transformer, converted into a voltage, and measured by a second vector voltage meter (V_2). The impedance of the DUT (Z) is then obtained from the values of V_1, V_2, and the resistance (R).

$$V_2 = I \times R$$

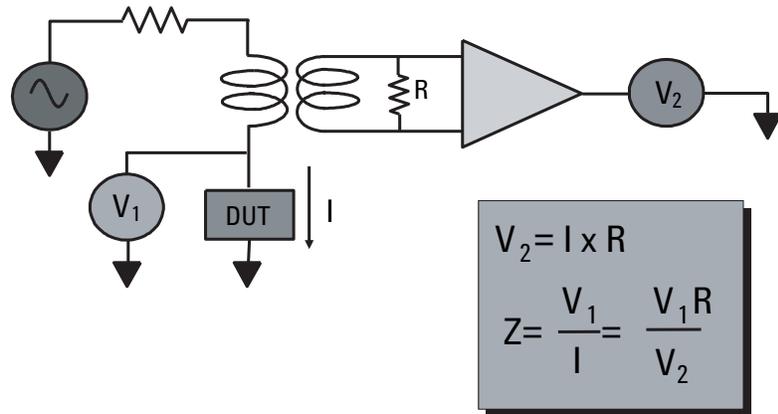$$Z = \frac{V_1}{I} = \frac{V_1 R}{V_2}$$

Figure 3. The RF I-V technique is based on a simple operating principle.

The simple operating principle of the RF I-V method enables measurement over a wider frequency range than was previously possible using a single instrument. The E4991A, for example, has a frequency range of 1 MHz to 3 GHz. The analyzer's basic measurement accuracy of 0.8% is a dramatic improvement over the accuracy achieved using the more traditional reflection-coefficient technique and a vector network analyzer.

Another advantage provided by the RF I-V method is improved measurement stability after calibration. Generally, changes in environmental temperature significantly influence the measurement of vector impedance at high frequencies. In a network analyzer, the reference and test channels of the voltage ratio detector (VRD) independently measure the input signals, as shown in Figure 4. Tracking errors, which occur when the relative gain and phase of the two channels vary after calibration is performed, are often the result of temperature variance. Because tracking errors cause the impedance measurement values to drift, these errors must be eliminated to improve measurement stability.

In the VRD section of the E4991A, also shown in Figure 4, the input signals V_v and V_i are multiplexed by the input switch. Each signal is measured alternately with two-channel VRD circuits in a measurement cycle. As a result of the vector voltage ratio calculation, tracking errors are canceled, which makes it possible to obtain stable impedance measurement results.
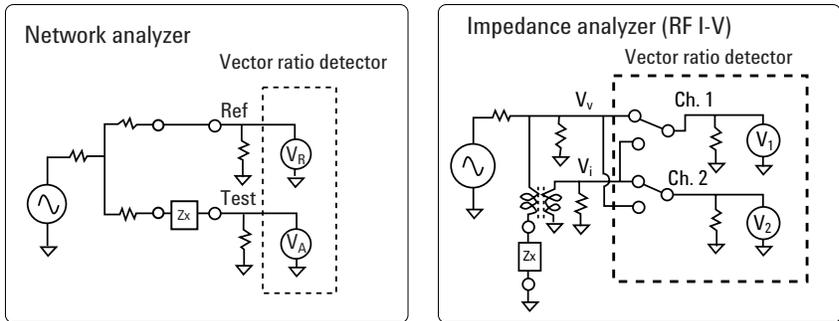

Figure 4. Block diagrams of the vector ratio detector in a network analyzer and the E4991A.

An additional advantage of the E4991A and RF I-V method is in the evaluation of temperature characteristics. The E4991A is not limited by the temperature dependencies that are evident in measurements using a network analyzer.

For example, if we use a modern vector network analyzer and typical reflection coefficient (Γ) and S-parameter (S_{11} and S_{12}) configurations to measure ESR, we see changes in the test results—illustrated in Figure 5—at 100 MHz and 1 GHz for temperature settings of 18°C, 23°C, and 28°C. (The temperature range is satisfied with instrument specifications).

In the same figure, however, we see that measurements made using the RF I-V method are stable across the temperature range, regardless of the frequency and the DUT's impedance.
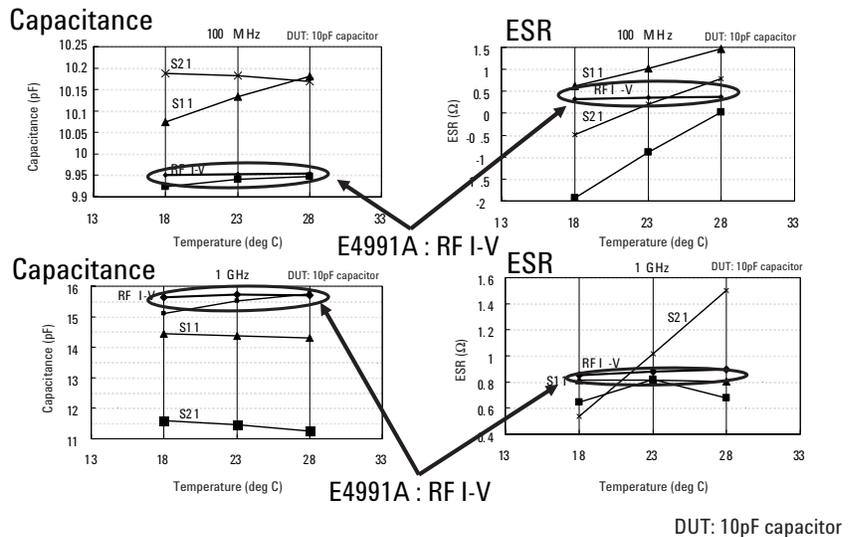

Figure 5. Temperature characteristics of a 10 pF capacitor. The measurements made using the E4991A and RF I-V method are stable across the 18° to 28° temperature range.

Calibration and Compensation in the RF Range

Conventional calibration methodology

Instrument calibration and compensation are processes that play a critical role in RF component testing. The conventional way to calibrate a network analyzer is with *open*, *short*, and *load* calibration. To perform this method of calibration, the open, short, and load reference terminations are connected to the test port and then each termination is measured. The calibration data is then stored in instrument memory and used to calculate and remove instrument errors. Impedance values for the reference terminations are indicated in two ways: with vector impedance coordinates and with a Smith chart, as shown in Figure 6.

Note: OPEN impedance is infinite, so it is not shown in the graph.

(a) Vector impedance plane   (b) Smith chart
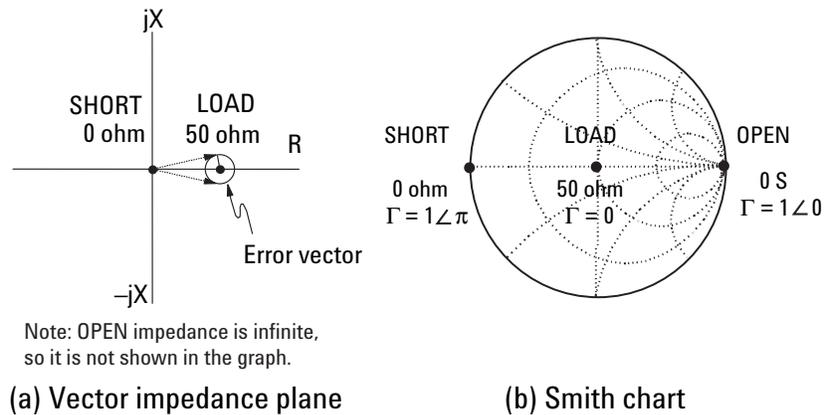
Figure 6. Impedance values for reference terminations are graphed on the vector impedance plane and on the Smith chart.

Although all three terminations are indispensable for calibration, the 50 Ω LOAD termination impedance is particularly important for ensuring precise calibration and has a large influence on the resulting measurement accuracy. The uncertainty of the LOAD termination impedance is represented by a circle that encloses the error vector, as we see in Figure 6a. The uncertainty of its phase angle increases with the frequency and becomes a considerable error factor, especially in measurements of high-Q devices at high frequencies.

Advanced calibration methodology

To achieve highly accurate measurement results using the RF I-V method, the E4991A employs an advanced calibration method that incorporates a low-loss capacitor (LLC) calibration standard. This calibration method is superior to a network analyzer's typical 50-ohm calibration, and it enables the E4991A to measure parameters such as high-Q with great accuracy. To do this, Agilent provides a new standard, a low-loss capacitor, to define phase—the absolute X axis as shown in Figure 7. The R and X axes are also defined, which allows us to accurately measure high Q.

When we calibrate the E4991A using the LLC standard, phase is determined as follows: First, we assume that Z_0 phase is equivalent to 0 degrees. Then we measure the LLC phase in order to get the difference from 0 degrees. Finally, we subtract this difference from Z_0 to define the phase standard. See Figure 7.

Note that the LLC calibration method is effective for measurements above 300 MHz.

X
Z
?
R

$Q = \tan\theta$

SHORT   OPEN   LOAD   Low loss (air) capacitor

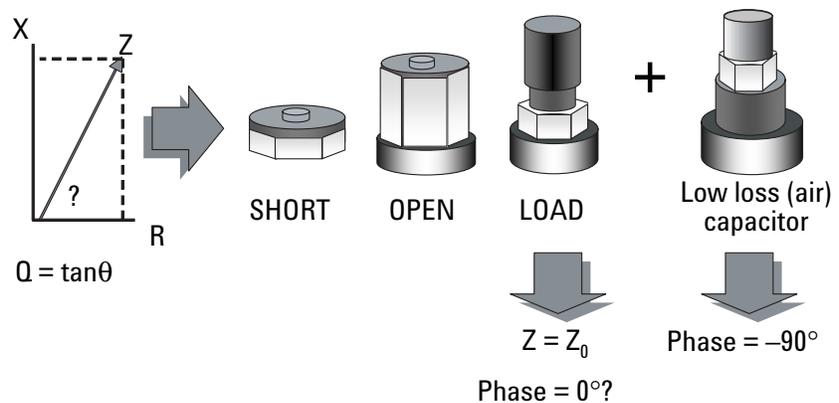$Z = Z_0$ Phase = $-90°$

Phase = $0°$?

Figure 7. Instruments that use the RF I-V method of impedance measurement employ an advanced calibration technique to obtain high accuracy.

High-Q measurement

High-Q (or low-D) measurement is currently a hot topic in RF component testing. Quality factor (Q) is defined as the ratio between reactance (X) and resistance (R) and is commonly used for inductors. The reactance is the product of $2\pi fL$. A low-Q inductor (that is, one with high R) absorbs more energy than a high-Q inductor. Therefore, high Q is a desirable characteristic in mobile equipment, which must conserve battery energy.

Why is it so difficult to measure a high-Q or low-D component accurately? To make this measurement, the test instrument must be able to measure a large reactance and a very small resistance at the same time. If, for example, Q = 100, the resistance will be 100 times smaller than the reactance, as depicted in Figure 8. The instrument's inaccuracy, shown as a small circle at the end of measured impedance vector, makes it difficult to measure the small amount of resistance. Even a tiny fluctuation can create a large error in the Q measurement result. Overcoming this difficulty requires a highly accuracy instrument.

Error vector

$$Q = \frac{X}{R}$$

Q=100

R value is 1/100 of X value
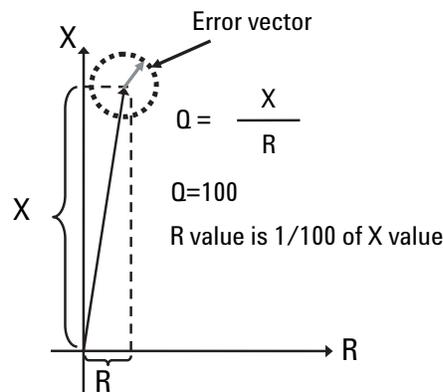
Figure 8. Instrument inaccuracy can add significant error to high-Q measurements.

If we use a conventional vector network analyzer to measure impedance at Qx=100, the measurement error will be greater than 65% and could go as high as 100%. Even if the value of Q is below 10, the measurement error will be approximately 10%.

We can obtain better results using the E4991A's RF I-V method with LLC calibration. As the chart in Figure 9 shows, at Q=100 the E4991A can measure up to 3 GHz with an error of 30% or less. At Qx=50, the measurement error is 15%.
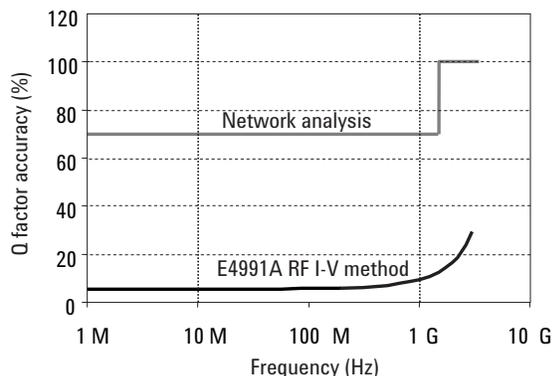
Note: Q accuracy is compared at a Q factor of 100 at 50 Ω impedance.

Figure 9. The RF I-V method with LLC calibration provides better Q measurements to 3 GHz.

Fixture compensation

In addition to performing calibration to remove instrument inaccuracies, we must compensate for the electrical characteristics of the test fixture and cable. To get the real characteristics of the DUT, two sources of error that exist between the calibration plane and the DUT must be removed: residual impedance and electrical length. See Figure 10.
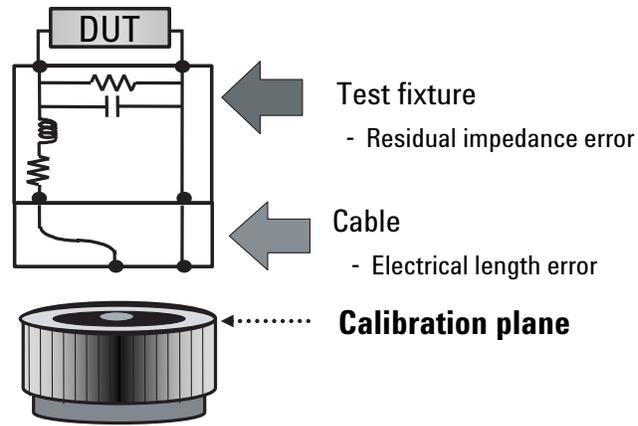
Figure 10. Two kinds of error—residual impedance and electrical length—can be introduced between the DUT and the calibration plane.

Residual impedance error

Small amounts of impedance, inductance, conductance, and capacitance exist in a test fixture, and these can lead to an impedance measurement error. With the E4991A, we can remove the impedance error by using open/short compensation at the test fixture port.

Electrical length error

Electrical length error is caused by a phase shift of the test signal at a 50-ohm electric line between the calibration plane and the DUT. It can cause a phase measurement error when we measure complex impedance parameters. We can compensate for this electrical length error by using the port extension feature of the E4991A.

Using compensation values of Yo and Zs and the measured value of Zm, we can calculate the real impedance of the DUT (Zdut) with the equation in Figure 11.
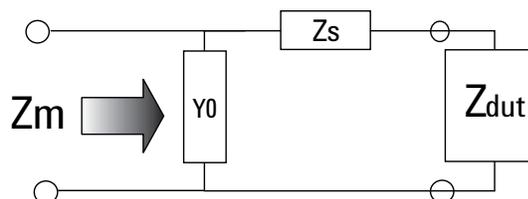
$$Zdut = \frac{Zm - Zs}{1 - (Zm - Zs)Yo}$$

Figure 11. The DUT impedance can be obtained with proper compensation.

Measurement Correlation and Repeatability

It is possible to obtain different measurement results for the same device when the same instrument and test fixture are used. There are many possible causes of such measurement discrepancies, as well as residuals. Typical factors that contribute to the measurement discrepancies in RF impedance measurements are described as follows.

Residual parameters at high frequencies

It is important to consider the effects that residual parameters have on measurement results at high frequencies.

Assume, for example, that a residual inductance of 0.1 nH and a stray capacitance of 0.1 pF exist around the measurement terminals of the test fixture. The effects of these small residuals differ depending on the frequency, as depicted graphically in Figure 12.
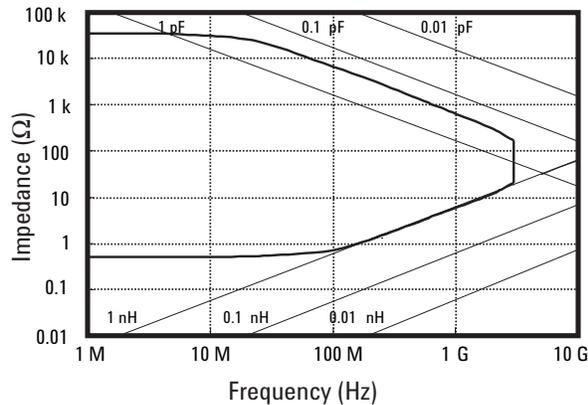
Figure 12. Relationships between the residual parameter values and the typical impedance measurement range using the RF I-V method.

In the low-frequency range, the residual parameter values are much smaller than the values of normally measured devices. This is because the capacitors and inductors, which are designed for use in low-frequency electronic equipment, possess large values compared to small residuals. Those devices used for higher-frequency circuits and equipment, however, have lower values. In the frequency range above 100 MHz, the majority of DUTs are low-value devices; that is, their values are in the low nH and low pF, approaching the values of the residuals.

Consequently, the residual parameters have a greater effect on higher-frequency measurements, and they become a primary factor in measurement error. The accuracy of the measurement results after compensation depends on whether the open/short measurements have been properly performed. To obtain measurement results with good correlation and repeatability, the compensation must be performed each time under identical conditions. Any differences in the compensation method will result in differences in the measured values, leading to problems in correlating the measurement results.

Variance in residual parameter values

The effective residual impedance and stray capacitance will vary depending on the position at which the DUT is connected to the measurement terminals. Residual inductance increases when the DUT is connected to the tips of the terminals rather than the bottom. Stray capacitance also varies with the DUT's relative position. See Figure 13.
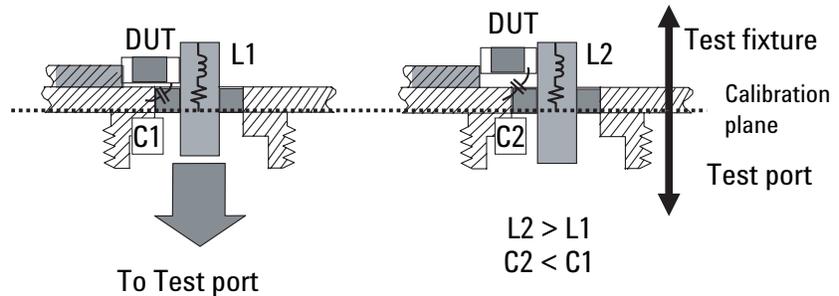
Figure 13. Difference in residual parameter values due to DUT positioning.

A difference in contact condition

Changes in the orientation of the DUT to the terminals are another sources of measurement discrepancy. When the DUT makes contact evenly across the measurement terminals, the distance of the current flow between the contact points is minimized, thus providing the lowest impedance measurement value. If the DUT tilts or slants, however, the distance of current flow increases, yielding an additional inductance between the contact points. See Figure 14.

Residual resistance also varies in this scenario, producing changes in the measured D, Q, and R values. Such positioning errors affect the measurement of low-value inductors and degrade the repeatability of measured values.
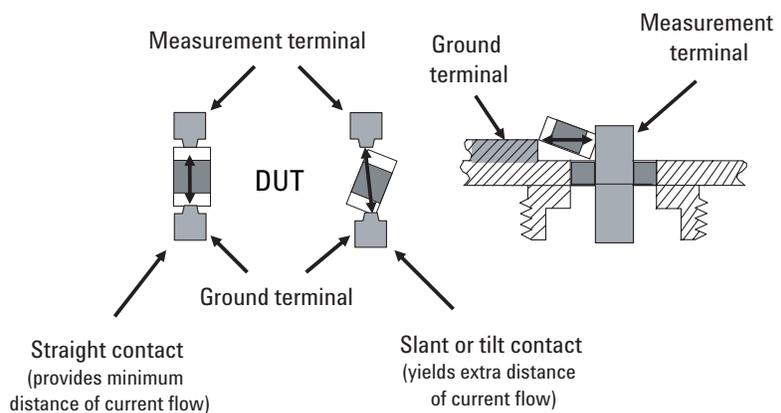
Figure 14. Altering the contact conditions changes the measured values of inductance and residual resistance.

A difference in open/short compensation conditions

Improper open/short measurements similarly decrease the accuracy of compensated measurement results. If the open/short measurement conditions vary, again we get inconsistent measurement values.

In particular, each short device has an inherent impedance (or inductance) value that, if not properly defined, can introduce error. The effective impedance of the short device will vary depending on how the DUT makes contact with the measurement terminals. As we see in Figure 14, the contact points for a bottom electrode test fixture differ from those for a parallel electrode fixture.

Contact points   Contact points

Shorting chip

Bottom electrode fixture   Parallel electrode test fixture
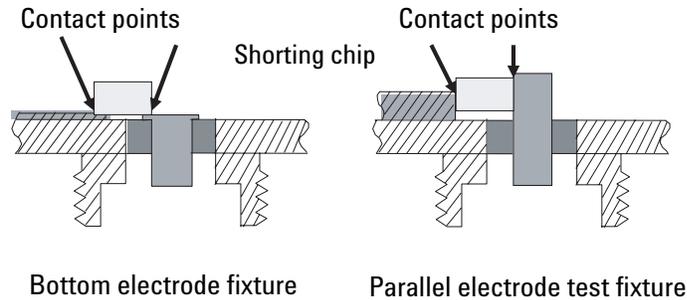
Figure 15. Different contact points can affect the results of open/short measurements.

Furthermore, if the short device is not straight—that is, if it curves slightly— the measured impedance will vary depending on which side of the device is facing up. These effects are usually small, but they should be taken into consideration, especially when we make very low inductance measurements, typically below 10 nH.

13

Electromagnetic coupling with a conductor near the DUT

Electromagnetic coupling between a DUT and a metallic object near the DUT varies with mutual distance and causes variance in measured values. Leakage flux generated around an inductive DUT induces an eddy current in a closely located metallic object. The eddy current suppresses the flux, decreasing the measured inductance and Q factor values. The distance of the metallic object from the DUT is a factor of the eddy current strength, as shown in Figure 16a.

Because test fixtures contain metallic objects, they can be important causes of measurement discrepancies. Open-flux-path inductors usually gene rate leakage flux with directivity. As a result, measured values will vary depending on the direction of the DUT. Differences in the eddy current due to leakage flux directivity are illustrated in Figure 16b, c, and d. If a parasitic capacitance exists between the DUT and an external conductor, it is difficult to remove the effect on the measurement because the guarding technique will be invalid. Thus, the DUT should be separated from the conductor with enough distance to minimize measurement errors.

(a) Metallic object near the leakage flux of inductor will cause an eddy current effect. This effect increases as the distance of the metallic object from the inductor decreases.

(b) Inductor with leakage flux directivity. Turning the inductor by 90 degrees reduces the eddy current as shown in figure (c).

(c) Metallic object in parallel with the leakage flux causes less eddy current.

(d) Inductor with less leakage flux directivity. The eddy current effect is almost independent from the direction of inductor's side faces.
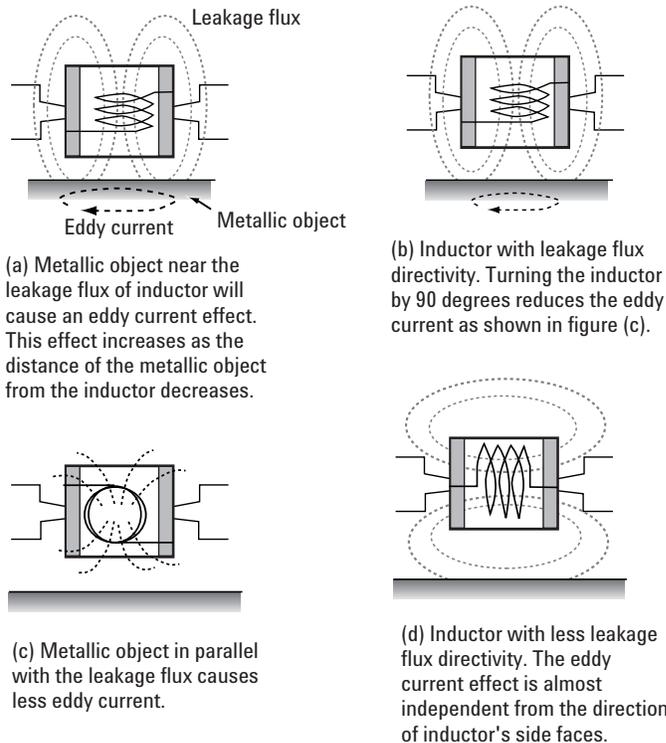
Figure 16. Eddy current effects and the magnetic flux directivity of a DUT are important test considerations.

Defining short-bar residual inductance

Surface mount device (SMD) test fixtures for the E4991A are furnished with short bars that can be used for various-sized devices. Optional short bars are available for standard industrial sizes (EIA, IEC, EIAJ). The 16197A comes furnished with only the industrial standard sizes.

Because the residual impedances of the short bars differ according to short-bar size, Agilent recommends using the same sized short bar whenever compensation is performed.

Two short bars of different sizes were used to make the 2 nH SMD inductor measurements in Figure 15. There is a discrepancy of approximately 200 pH between the measurements that could have been prevented by using short bars of the same size.

The residual impedance of the short bar is defined not simply by the existing residuals in the short bar itself. It also depends on how the short bar is placed in the surrounding conditions. Factors that can affect residual impedance include the permittivity, thickness, micro-stripline, and ground conditions of the printed circuit board, as well as other environmental factors. Unless variations in the surrounding conditions are minimized, it is very difficult to simulate the residuals of the short bar. When we measure extremely small inductance values, there are two methods of defining the residual inductance of the short bar.

One definition is to let short equal 0 H. With this method, the measurement result is the relative value of the SMD inductor's impedance to the short bar's impedance. The short bar's residual inductance based on its size and shape is not estimated. Traditionally, this has been the primary method of compensating the test fixture's residual inductance.

The second definition is to let SHORT equal xH. With this method, the measurement result is the absolute value of the SMD inductor's impedance. The short bar's residual inductance based on its size and shape is estimated and is used as a reference value.

Both measurement results are correct, and the difference in results is due only to the difference in the definition methods. However, we may still have a problem in obtaining a good correlation.

For example, if we use the same short bar to measure a 10 nH inductor, and the short bar's residual inductance is defined as 0 nH in one instance and 0.4 nH in another, the measurement result will differ by 4%.

Changing the short bar's residual parameters has a greater effect on higher-frequency measurements. In Figure 17, different reference residual inductance values are used to measure a 100 pF capacitor. We see that the resonant frequency shifts by more than 200 MHz.

If short equals 0H, resulting in the short bar's residual equaling 0.4 nH, the SRF is calculated to be 762 MHz—a higher value than the 550 MHz resonant frequency obtained when SHORT equals 0.4 nH.
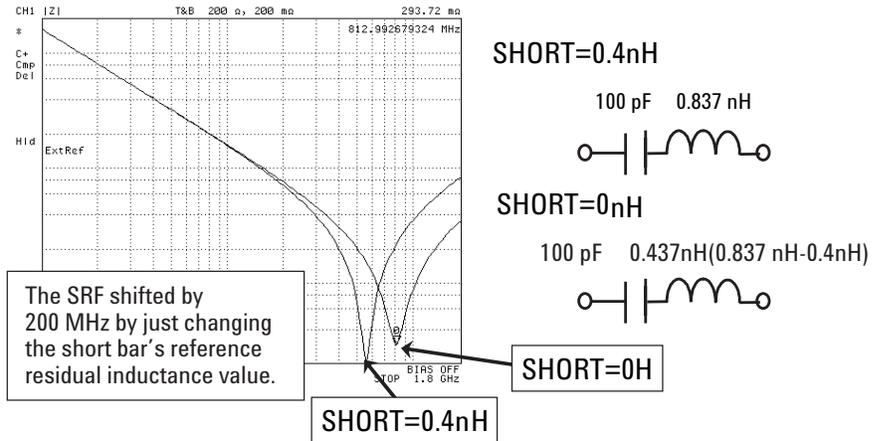
Figure 17. Using different reference residual inductance values results in different SRFs.

Although the short bar's residual inductance can be defined as short = 0H or short = xH, and either definition will give correct results, we have to use the same method of definition as well as the same measurement conditions (test fixture, compensation method, frequency and test signal) when we correlate measurement results, especially those of inductors with low values.

Agilent generally chooses to use short = 0 H, but for the 16196A/B/C family of test fixtures that are used with the E4991A, the short bar's residual inductance values have been provided in the form of theoretical values, so either method of definition can be used and residuals can be calculated without error.

Measurement repeatability

Table 1 compares the major Agilent RF SMD test fixtures. Short repeatability, which becomes critical when we measure low nH inductors, is smallest with the 16196x and greatest with the 16191A/2A/3A fixtures. Proportional error, which becomes critical when we measure impedance at high frequencies, is again smallest with the 16196x and greatest with the 16191A/2A fixtures.

Table 1. To maintain data correlation, we need to use a test fixture that has high repeatability and low proportional error.

Key Items	16196x	16197A	16191A/2A
Test Frequency Range	Dc to 3GHz	Dc to 3GHz	Dc to 2GHz
Short Repeatability [mΩ]	30+125 x f [GHz]	30+ 150 x f [GHz]	30+250 x f [GHz]
Open Repeatability [μS]	5+40 x f [GHz]	2+30 x f [GHz]	2+30 x f [GHz]
Proportional Error [%]	1.0 x f^2 [GHz]	1.2 x f^2 [GHz]	1.5 x f^2 [GHz]

Table 2 summarizes the importance of calibration and compensation methodologies that are used to measure low-ESR/high-Q components.

Table 2. This table illustrates the main features of each calibration and compensation method.

	Theory
Calibration	Eliminates instrument system errors Defines the "Calibration Plane″ using the CAL standard Low Loss C Calibration improves high Q measurement in RF
Cable correction	Eliminates the effects of cable error Extends the "Calibration Plane" to the end of the cable
Compensation	Eliminates the effects of error sources existing between the "Calibration Plane" and the DUT
OPEN/SHORT Compensation	Eliminates the effects of simple fixture residuals Note: Use the same short bar size and definition value to keep the data correlation
OPEN/SHORT/LOAD Compensation	Eliminates the effects of complicated fixture residuals Note: Use the same short bar size and definition value to keep the data correlation

Summary

The RF I-V method of impedance measurement, implemented in the E4991A impedance analyzer, provides accurate and stable results across a wide RF frequency range. An advanced method of calibration using a low loss capacitor enables highly accurate measurement of high Q.

Test fixture compensation should be made using one short-bar size and the same definition of short each time. Also, keep in mind that the position of the connection and the orientation of the short bar in relation to the test fixture terminals will influence the impedance measurement results. Ensuring consistency in the contact environment is important in maintaining data correlation.

Appendix:
Special Calibration Method for High Q Measurement

Agilent has developed a special, easy-to-perform method of calibrating the E4991A RF impedance/material analyzer that provides the impedance phase accuracy needed for high-Q device measurements. This calibration technique is described below.

Outline of the special calibration

Although a measurement instrument may be stable, high phase accuracy is still required to measure Q factors accurately. 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 standard used in the OSL calibration. One method of improving the accuracy of the phase measurement is to use a phase-calibrated load standard. However, we cannot be sure that the phase uncertainty of the calibrated 50 Ω load will be smaller than 1E-3 at high frequencies (such as 1 GHz).

An additional technique is to use, in addition to the normal open-short- load standards, a low loss capacitor as the second load (Load 2), keeping the dissipation factor (D) below 1E-3 at 1 GHz. In this case the uncertainty of the measured phase is decreased from the phase uncertainty of the 50 Ω load (Load 1) to the uncertainty of D of the low loss capacitor (Load 2) for almost all DUT impedances.

Details of the low loss capacitor (LCC) calibration

We want a calibration method that will reduce the error in phase measurement in spite of the phase error associated with the 50 Ω load. Consider a case in which we have a 50 Ω load standard whose impedance phase is unknown, but whose impedance magnitude is known. We add another load (Load 2), and this time the impedance magnitude is unknown and the impedance phase is known. Load 2 is a low loss capacitor. The number of unknown circuit parameters is still three at most. However, two more unknowns related to standards have been added. There are now eight real unknown parameters:

- Circuit parameter K1 (two real parameters)
- Circuit parameter K2 (two real parameters)
- Circuit parameter K3 (two real parameters)
- Impedance phase $\theta ls1$ of the 50 Ω LOAD (one real parameter)
- Impedance magnitude Zabs_ls2 of the low loss capacitor LOAD2 (one real parameter)

We solve this problem analytically. For the simplest case in which both the open and the short standard are ideal, the circuit parameters are found as follows.

$K1 = A \times Z_{ls} \times R_0$

$K2 = -Z_{sm} / R_0$

$K3 = -Y_{sm} \times R_0$

where:

R_0 = characteristic impedance

A = $(1 - Zlm_i \, {}^*Yom)/(Zlm_i - Zsm)$

Yom = measured admittance for OPEN STD

Zsm = measured impedance for SHORT STD

Zlm_i = measured impedance for LOAD STD
(i = 1:LOAD1, i = 2:LOAD2)

Zls_i = true impedance for LOAD STD
(i = 1:LOAD1, i = 2:LOAD2)

Zls_1 = $Zabs\_ls1*EXP(j*\theta ls1)$

Zls_2 = $Zabs\_ls2*EXP(j*\theta ls2)$

$\theta ls1$ = $\theta 2 - \theta 1 + \theta ls2$

$Zabs\_ls2 = A1/A2*Zabs\_ls1$

Zabs_ls1 = impedance magnitude for LOAD1 (50 Ω): known
$\theta ls2$: impedance phase for LOAD2 (low loss capacitor): known

$\theta 1$ = $arg((1-Zlm_1 \, {}^*Yom)/(Zlm_1 - Zsm))$

$\theta 2$ = $arg((1-Zlm_2 \, {}^*Yom)/(Zlm_2 - Zsm))$

$A1$ = $|(1-Zlm_1 \, {}^*Yom)/(Zlm_1 - Zsm)|$

$A2$ = $|(1-Zlm_2 \, {}^*Yom)/(Zlm_2 - Zsm)|$

In 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 Zls1 = 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 (Zls1).
- Execute correction for LOAD2 and get the corrected impedance (Zcorr2).
- Calculate the phase difference (Δθ) between the phase of Zcorr2 and the true phase of LOAD2.

Step 2

- Modify the impedance of LOAD1 to Zls1' whose phase is −Δθ and whose impedance magnitude is still 50 Ω.
- Calculate the circuit parameters again by normal OSL calibration using modified LOAD impedance Zls1'.

Although this method gives an approximation, performing just these two steps provides enough accuracy for our purposes.

Phase measurement error factors

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

1) Uncertainty for impedance magnitude of LOAD1

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 $|R_0 * \Delta Y open| * (Copen/Cls2)$. In the case of resistive DUTs, the sensitivity is the same as in the normal OSL calibration.

where:

R_0 = characteristic impedance = 50 Ω,

Copen = capacitance of OPEN STD, and

Cls2 = capacitance of LOAD2 (low loss capacitor).

Based on the discussion above, the phase measurement error when we use the low- loss- capacitor calibration method is determined mainly by

• Uncertainty for the impedance phase of LOAD2

• Uncertainty for the admittance magnitude of OPEN

Now we evaluate two other items. The D factor for the capacitor (3 pF) 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 proportional to the frequency, 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 ±5 fF at most.

This gives us 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 low loss capacitor calibration.

For More Information

www.agilent.com/find/impedance

Related Literature

Literature Number 5952-1530E
Agilent LCR Meters, Impedance Analyzers, and Test Fixtures

Literature Number 5950-3000
Impedance Measurement Handbook

Literature Number 5965-4792E
Accessories Selection Guide For Impedance Measurements

Literature Number 5980-1233E
Agilent E4991A RF Impedance/Material Analyzer Technical Specifications

Literature Number 5980-1234E
Agilent E4991A RF Impedance/Material Analyzer

Literature Number 5988-0728EN
Advanced impedance measurement capability of the RF I-V method compared to the network analysis method AN 1369-2

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