Measuring Frequency Response with the Keysight E5061B LF-RF Network Analyzer

Application Note

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TECHNOLOGIES
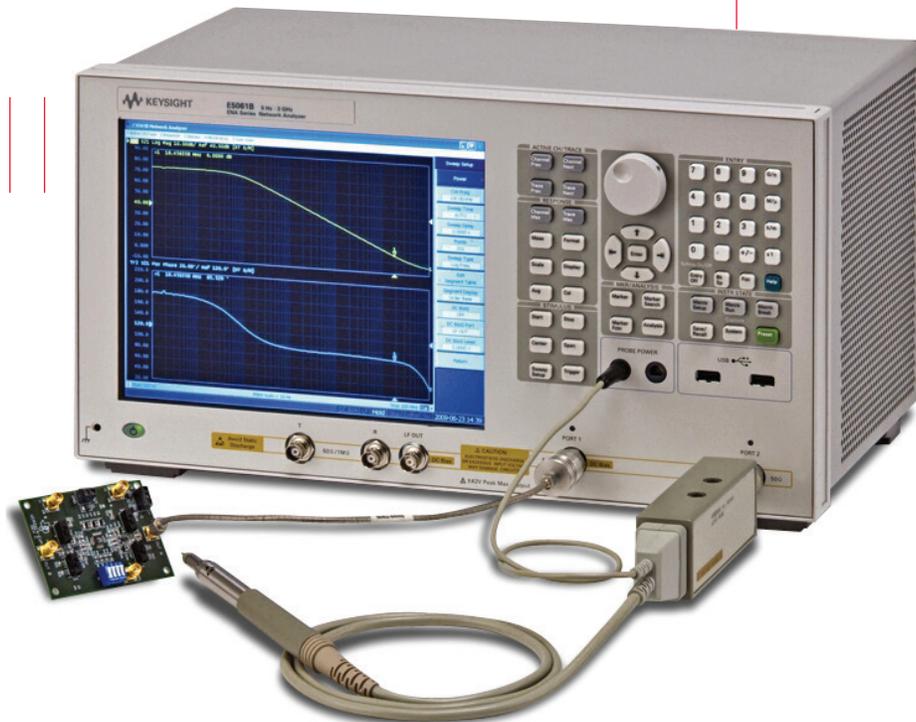
Introduction

Evaluating frequency responses of components and circuits is essential for ensuring the performance of electronic equipment. Especially in the case of the high-reliability electronic equipment used in automotive, medical, and aerospace and defense industries, it is necessary to evaluate a wide variety of components and circuits in the low- to high-frequency ranges. Among these applications, the low-frequency network analyzer plays an important role in ensuring the stable and reliable operation of low-frequency analog circuits such as sensor systems and power supplies. You need a better understanding of low-frequency network analysis (gain-phase measurements) as well as RF network analysis (S-parameter measurements).

This application note describes fundamentals of low-frequency network analysis using the E5061B LF-RF network analyzer. Here we mainly discuss simple low-frequency 2-port device measurements and associated topics such as high-impedance probing techniques and high-attenuation measurements.

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Table 1. Guideline for selecting test ports

Test ports	Situation	Application
S-parameter test port	Transmission and reflection measurements in the 50 Ω system impedance	Passive filters, antennas, cables, RF amplifiers, etc.
	Transmission measurements with high-impedance probing over 30 MHz using the 41800A active probe	High-frequency op amps
Gain-phase test port	Transmission measurements with high-impedance probing in the low frequency range	OP amp circuits
	Feedback loop measurements	DC-DC converter loop gain
	High-attenuation measurements in the low-frequency range	CMRR and PSRR of OP amps

E5061B-3L5 LF-RF Network Analyzer

The E5061B with the Option 3L5 vector network analyzer covers a broad test frequency range from 5 Hz to 3 GHz in a single instrument. The E5061B-3L5 includes an S-parameter test port (5 Hz to 3 GHz, Z_{in} = 50 Ω) and a gain-phase test port (5 Hz to 30 MHz, Z_{in} = 1 MΩ/50 Ω). Both test ports can be used for low-frequency applications, depending on your measurement needs. Table 1 shows an example of selecting the test ports.
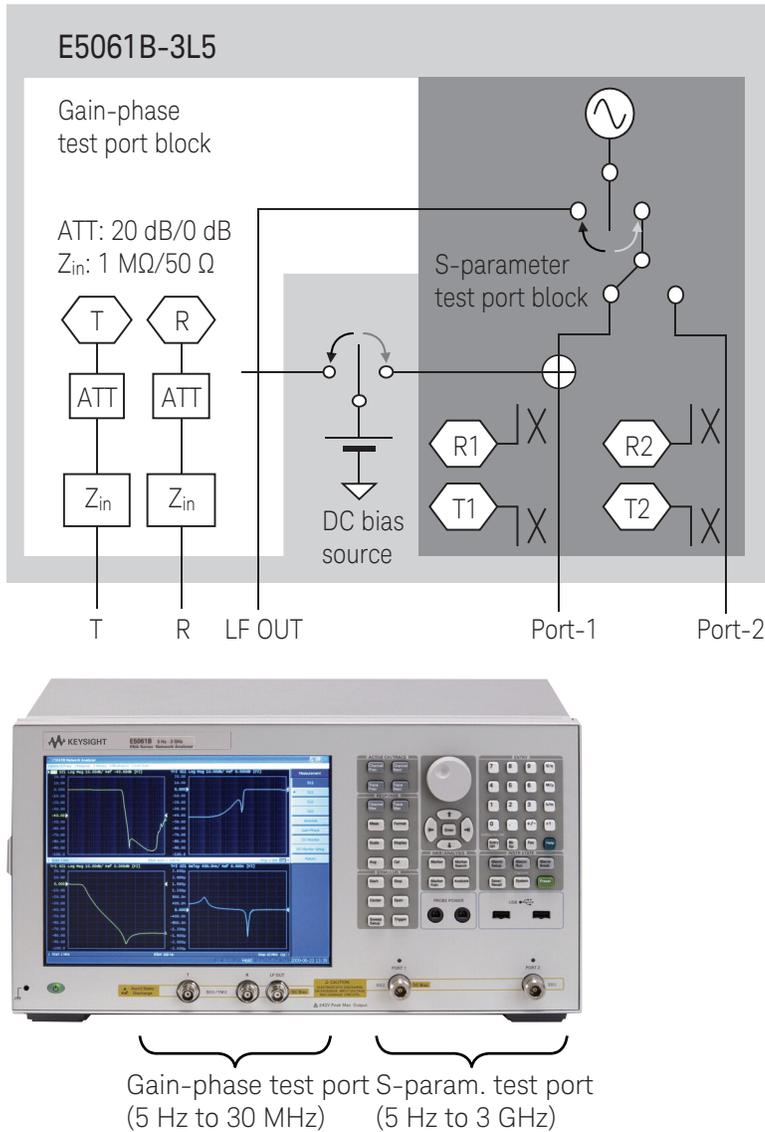

Figure 1. E5061B-3L5 simplified block diagram

Basic Measurement Configurations

50 Ω DUTs

First let's summarize how to connect DUTs in typical low-frequency network measurement applications. Here the focus is on configurations for 2-port transmission measurements. The first case is a transmission measurement for 50 Ω devices, such as filters and cables. Figure 2 shows a configuration using the gain-phase test port. The R-ch receiver VR monitors the source output voltage applied to the 50 Ω system impedance (incident voltage to the 50 Ω transmission line), and the T-ch receiver VT monitors the transmitted voltage. Then the analyzer measures the voltage ratio VT/VR which indicates the transmission coefficient S21.

Figure 2 shows a configuration using the S-parameter test port. The S-parameter test set has built-in directional bridges, and an external power splitter is not required. In most cases, the S-parameter test port for 50 Ω transmission measurements is used.

Most 50 Ω transmission measurement can be covered with the S-parameter test port. For high-attenuation devices such as, mΩ impedance measurement for DC-DC converters and large bypass capacitors using the shunt-thru method, the 50 Ω transmission measurement should be performed using the gain-phase test port rather than the S-parameter test port. In this case, the semi-floating receiver architecture of the gain-phase receiver ports eliminate the measurement error in the low-frequency range, which is caused by the test cable ground loop between the source and receiver (discussed later).
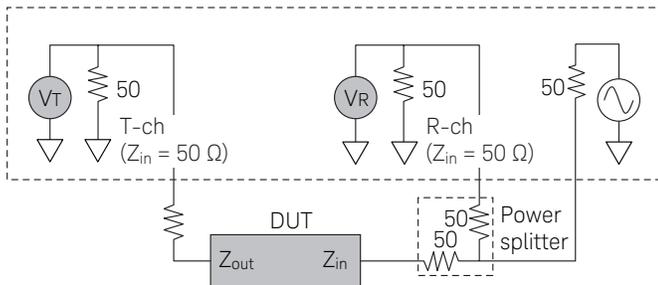
E5061B-3L5

Figure 2. Configuration for measuring transmission coefficient of 50 Ω DUTs with the gain-phase test port
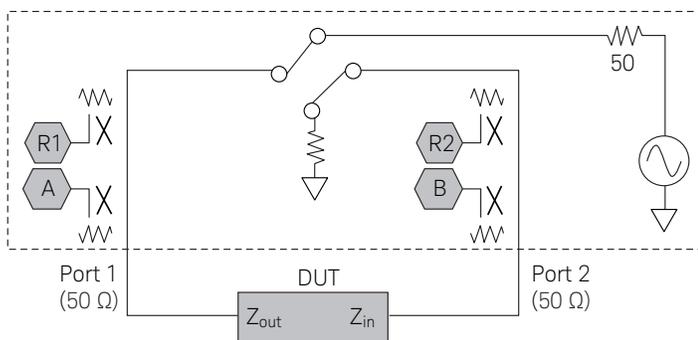
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Figure 3. Configuration for measuring transmission coefficient of 50 Ω DUTs with the S-parameter test port

Basic Measurement Configurations

Non-50 Ω DUTs, example 1

Low-frequency 2-port devices often have non-50 Ω impedances. The most typical examples are low-frequency amplifier circuits. Figure 4 shows a configuration example of measuring the frequency response of amplifiers with the gain-phase test port. The DUT has high input impedance and the output port is terminated with a non-50 Ω load ZL. The load impedance ZL depends on requirements of the targeted application. The load ZL can be either a resistive load or a reactive load.

The parameter to be measured is the voltage transfer function from the DUT's input port to the output port, V_{out}/V_{in}. The difference from the 50 Ω transmission coefficient measurements which were shown in Figures 2 and 3 is that the R-ch receiver VR directly monitors the AC voltage across the DUT's input impedance Z_{in} with high-impedance probing, instead of monitoring the voltage across the 50 Ω system impedance. The output voltage V_{out} can be monitored using the high-impedance probing, without affecting the DUT's load condition.

The analyzer's high-impedance receivers and the DUT should be connected with coaxial test cables or 10:1 passive probes, depending on the requirements of the maximum test frequency, the probing input impedance, the probing input capacitance, and so on (discussed later). When you use the coaxial test cables, a T-connector can be used at the R-ch probing point. To compensate the frequency response and phase errors between two probes/test cables, the response thru calibration should be performed. For example, by contacting the T-ch probe to the point TP1.
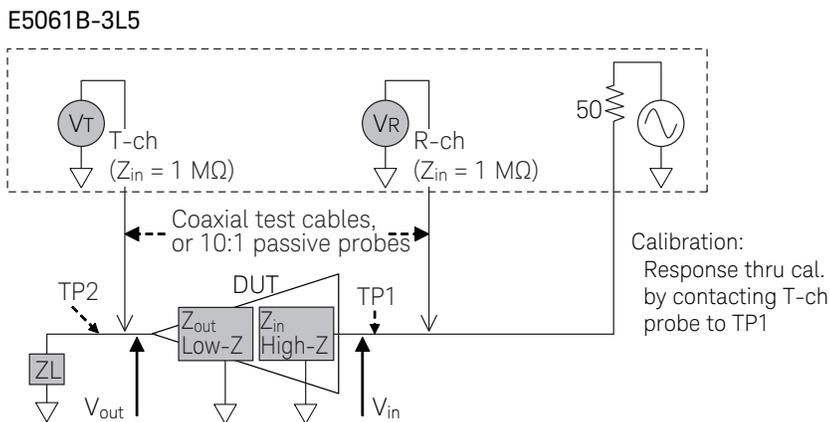

Figure 4. Configuration for measuring amplifiers with the gain-phase port (up to 30 MHz)

Basic Measurement Configurations

Non-50 Ω DUTs, example 1 *(continued)*

If you need to measure the frequency response of an amplifier up to more than 30 MHz, or if you need to probe the amplifier with a very small probing capacitance, the solution is to use the active probe at the S-parameter test port.

Figure 5 shows a configuration example. Unlike the configuration that was shown in Figure 4, the ratio measurement is referenced to the 50 Ω impedance of the built-in R1 receiver, and the response thru calibration must be performed by probing TP1 in order to correctly measure the voltage transfer function Vout/Vin. If the response thru calibration is not performed (and if a feed thru is not connected), the measured gain will be about 6 dB higher than the correct value because the AC voltage measured at the internal 50 Ω reference receiver will be about half of Vin.

For measurements in high frequencies over tens of MHz, connecting the 50 Ω feed thru to the DUT's input port can prevent the standing wave that may be caused by the impedance mismatch between the analyzer's 50 Ω source and the DUT's high input impedance. However, it must be noted that connecting the feed through will form a shunt signal path from the center conductor to the ground of the test cable, and this may cause measurement errors associated with the ground loop in the high-attenuation measurements such as CMRR and PSRR. If this is of concern, it is better not to connect the feed thu.
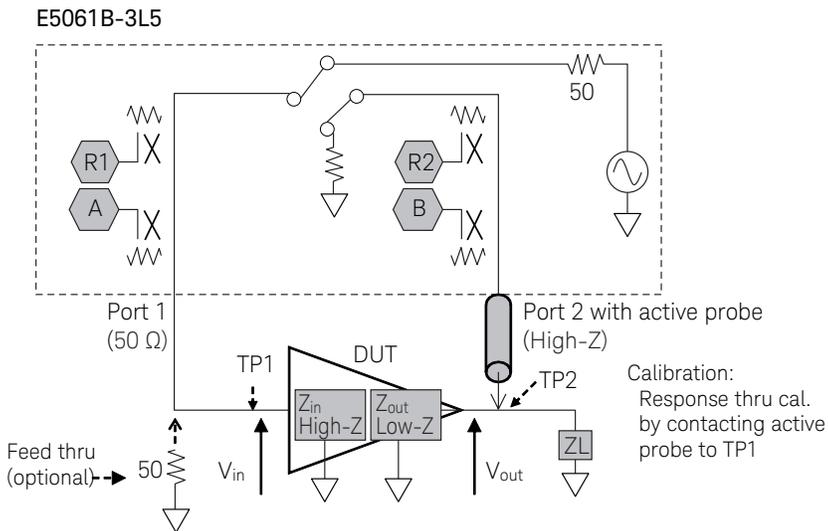
E5061B-3L5

Figure 5. Configuration for measuring amplifiers with the S-parameter test port and active probe (up to 30 MHz)

Basic Measurement Configurations

Non-50 Ω DUTs, example 2

Figure 6 and 7 shows configuration examples for measuring 2-port devices whose input and output impedances are several hundreds of Ωs to 1 or 2 kΩ. Typical applications are low-frequency passive filters, such as ceramic filters and LC filters. In these examples, impedance matching is implemented by simply connecting a series resistor. The configuration of Figure 6 uses the gain-phase test port. The ratio VT/VR indicates the transmission coefficient for the 1 kΩ system impedance.

Some types of filters need to be tested by connecting a load capacitor CL in parallel with the load resistor. The input capacitance of the high-impedance probe must be small enough not to affect the filter's characteristics. So the high-impedance T-ch receiver should be connected with the 10:1 passive probe which has the input capacitance around 10 pF. Or, if the DUT is very sensitive to the capacitive loading, use the S-parameter test port with the active probe, see amplifier measurement configuration shown in Figure 5.

The equivalent measurement can be achieved by using the 50 Ω input instead of using high-impedance probing at the T-channel and connecting another matching resistor as shown in Figure 7. This configuration is simpler and has an advantage that no probe capacitance is applied at the T-ch. However, it is not suitable for testing high-rejection filters because the measurement dynamic range is degraded by the series matching resistor. The degradation is 20*Log (50/1000) = 26 dB, in this case.
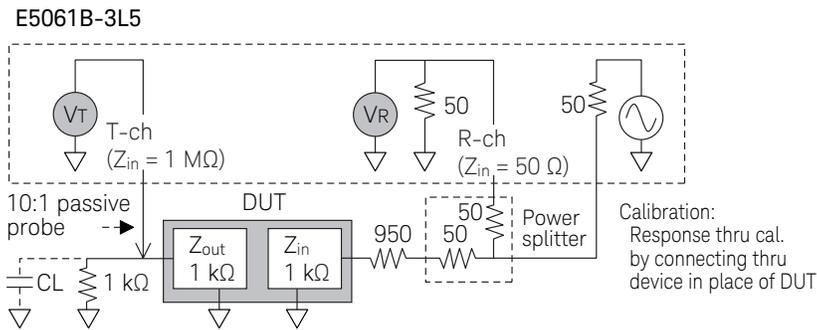
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Figure 6. Configuration for measuring passive IF filters with high-impedance probing
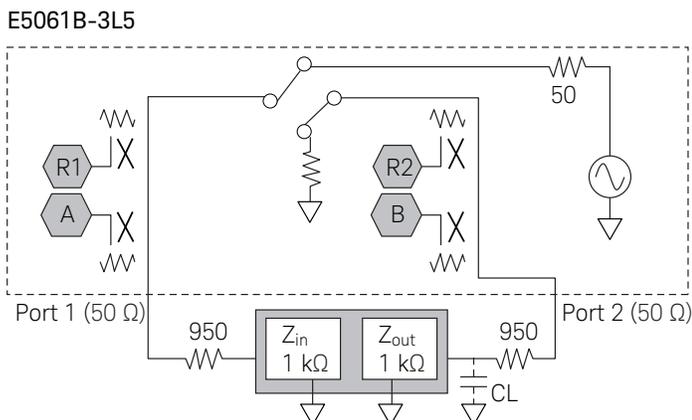(for DUTs not extremely sensitive to capacitive loading)

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Figure 7. Configuration for measuring passive IF filters with 50 Ω input

Basic Measurement Configurations

In-circuit probing measurements

The next application example is an in-circuit probing measurement, in which we measure the frequency response between two test points in the circuit under test. Figure 8 shows how to measure the frequency response of block-2 with the gain-phase test port. The frequency response of the circuit block-2 can be directly measured by probing TP1 and TP2 with dual high-impedance probing.

Similarly to the amplifier measurement configuration shown in Figure 4, connection between the analyzer's high-impedance receivers and the DUT should be appropriately selected from either the coaxial test cables or 10:1 passive probes, depending on the requirements on maximum test frequency, probing input impedance, probing input capacitance, and so on.
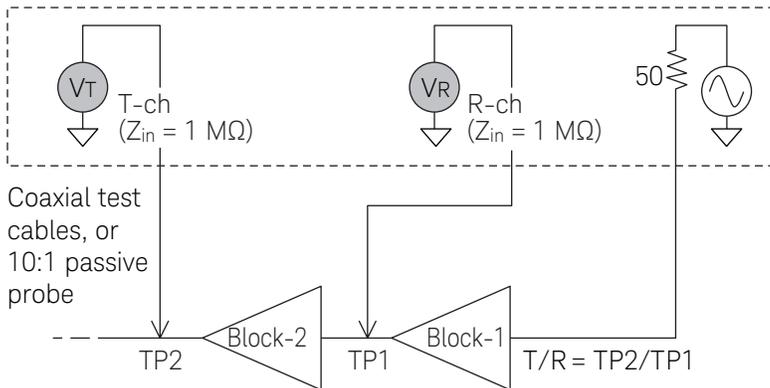
E5061B-3L5

Figure 8. In-circuit measurement using dual high-impedance probing with the gain-phase test port (up to 30 MHz)

Basic Measurement Configurations

In-circuit probing measurements *(continued)*

The maximum test frequency range of the E5061B's gain-phase test port is 30 MHz. If you want to perform incircuit measurements up to more than 30 MHz, the solution is to connect a single active probe to the S-parameter test port, and perform the two step measurement sequence as illustrated in Figure 9.

First we measure the response of the block-1 by connecting the active probe to TP1 and save the measured data into the memory trace. And then we measure the entire response of the block-1 plus block-2 by probing TP2. The measured data is stored into the data trace. Then we can obtain the frequency response of the block-2 using the data/memory trace math function of the analyzer.

The equivalent measurement is possible if we performing the response thru calibration by probing TP1 and then performing the measurement by probing TP2. This will directly give the response of the block-2 referenced to TP1 without using the trace math function.

If the DUT's output characteristic at TP2 is sensitive to the capacitance at TP1, the DUT's condition in the step 2 will slightly differ from that of the step 1, and the measurement result obtained by combining these two measurement results may contain errors. To minimize errors, connect a dummy capacitor C2 whose capacitance is about the same as the input capacitance of the active probe only when making the measurement of step 2 as shown in Figure 9. For example, this capacitance compensation method is required for measuring the phase margin of high-speed OP amps using this dual-step measurement technique. (A measurement example will be shown later.)
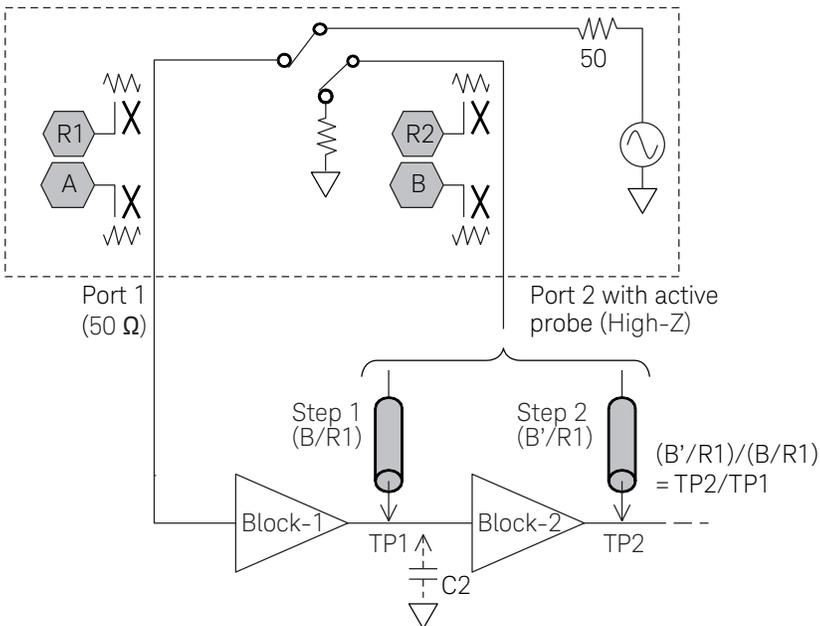
E5061B-3L5

Figure 9. In-circuit measurement with a single high-impedance probe (up to 30 MHz)

IFBW Setting in Low-Frequency Measurements

The IFBW (IF bandwidth) setting is one of the most common questions that many LF network analyzer users may first encounter. In high-frequency measurements, it is possible to use a wide IFBW for faster sweep speed, but for low-frequency measurements we need to set the IFBW to a narrow value to avoid measurement errors mainly caused by the LO feed through. For example, let's assume the case of measuring a high-attenuation device where start frequency = 1 kHz and IFBW = 3 kHz. The small signal attenuated by the DUT is up-converted to an intermediate frequency (IF) and passes through the IF filter of the receiver. Here the problem is that the leakage signal from the local oscillator (LO feed through) also passes through the IF filter because its frequency is very close to the IF frequency as shown in Figure 10, and this causes unwanted large measurement response.

Figure 11 shows an example of measuring a 60 dB attenuator with the E5061B's gain-phase test port under the conditions of source level = –10 dBm, start frequency = 1 kHz, and IFBW = 3 kHz, the attenuator setting of T and R ports is 20 dB. As you can see, incorrect measurement response appears around the start frequency due to the LO feed through. A similar problem also occurs even when the measured RF signal level is high (e.g. in a low-pass filter measurement).

In this case, the measured trace around the start frequency will be unstable due to the interference caused by the LO feed through that exists in the very close frequency to the RF signal. To avoid these problems, set the IFBW to a sufficiently narrower value than the start frequency (e.g. 5 times smaller), or use the IFBW AUTO mode in which the analyzer automatically selects narrow to wide IFBW settings depending on the frequency decade in the logarithmic sweep, so that the total sweep time won't be very long. The E5061B's IFBW AUTO mode sets the IFBW to one fifth of the start frequency of each decade.
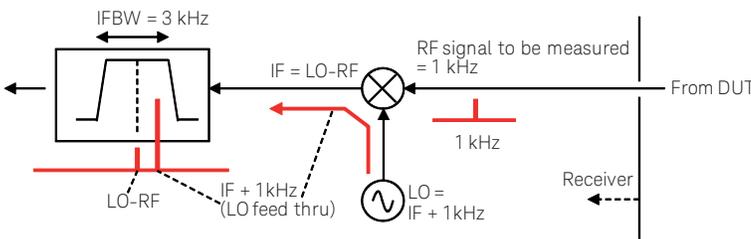

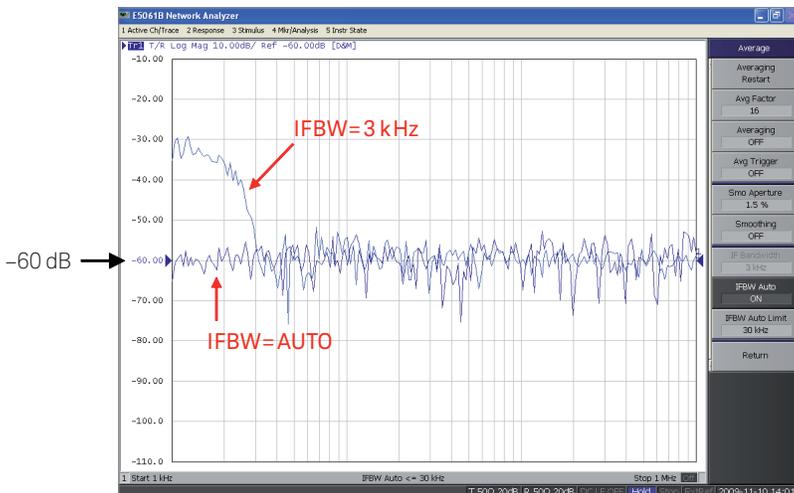Figure 10. Measurement error caused by LO feed through

Figure 11. Example of 60 dB attenuator measurement (Start = 1 kHz, IFBW = 3 kHz and AUTO)

High-Impedance Probing Methods

Using an appropriate probing method is important for making accurate high-impedance probing measurements. Special attention needs to be made to the probe input capacitance. The large input capacitance reduces the probe input impedance at high frequencies. For example, if the input capacitance at the probe end (= C_{in}) is 100 pF, the input impedance (= $1/(2*pi*f*C_{in})$) is 15.9 kΩ at 100 kHz, which is still high impedance. If the frequency goes up to 10 MHz, the input impedance is 159 Ω, which is not high enough for many applications. Also, the large input capacitance affects measurements which are sensitive to capacitive loading, such as passive IF filters, resonant circuits, and some amplifier parameters which depend on the capacitance condition (e.g. phase margin measurement). For these applications, it is necessary to use probing methods which provide small input capacitances if the network analyzer has a high-impedance input port like the E5061B. The easiest way for accessing the DUT is to use a coaxial test cable, such as a BNC to test clip lead, or a 1:1 passive probe to the high-impedance input port as shown in Figure 12.

If the measurement frequency range is lower than 1 MHz and if the probing capacitance of capacitive loading is not a problem for the DUT, this method is a good solution. Unlike a 10:1 passive probe, the measurement dynamic range is not degraded by the probe and small signals can be measured with a good signal-to-noise ratio. The drawback of this method is that the input capacitance of the probe will be large, because the test cable capacitance is added to the capacitance of the high-impedance input port. The input capacitance at the cable end will be more than several tens of picofarads even if using a short cable. Therefore, this method is not suitable for measurements in the high-frequency range of over 1 MHz. Also, it is not suitable for the measurements which are sensitive to capacitive loading.
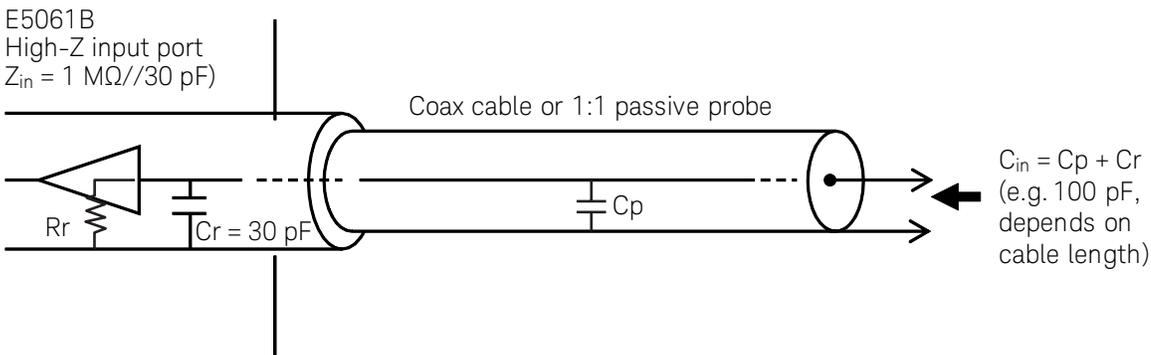

Figure 12. Coaxial test cable or 1:1 passive probe

High-Impedance Probing Methods *(continued)*

The probe input capacitance can be reduced using a 10:1 passive probe for oscillo-scopes, which is designed for use with the high-impedance input port, as shown in Figure 13. The 10:1 passive probe generally gives small input capacitance around 10 pF at the probe end, which enables high-impedance probing up to higher frequencies. Similarly to general oscilloscope applications, using the 10:1 passive probe is an ortho-dox way for high-impedance probing if the analyzer has built-in high-impedance inputs. The drawback is that the measurement dynamic range is degraded by 20 dB due to the 10:1 attenuation of the probe. So this method is not suitable for applications where very small signals need to be measured.

The active probe provides a high input resistance and a very small input capacitance without attenuating measured signals due to the active circuit integrated in the probe end, as shown in Figure 14. For example, the input resistance//capacitance of the 41800A active probe (DC to 500 MHz) is 100 kΩ//3 pF. Moreover, by adding the 10:1 adapter at the probe end, we can achieve 1 MΩ//1.5 pF, although the dynamic range is degraded by 20 dB in this case. If you need to measure over 30 MHz, or if the DUT is very sensitive to the capacitive loading, it is recommended to use active probe.
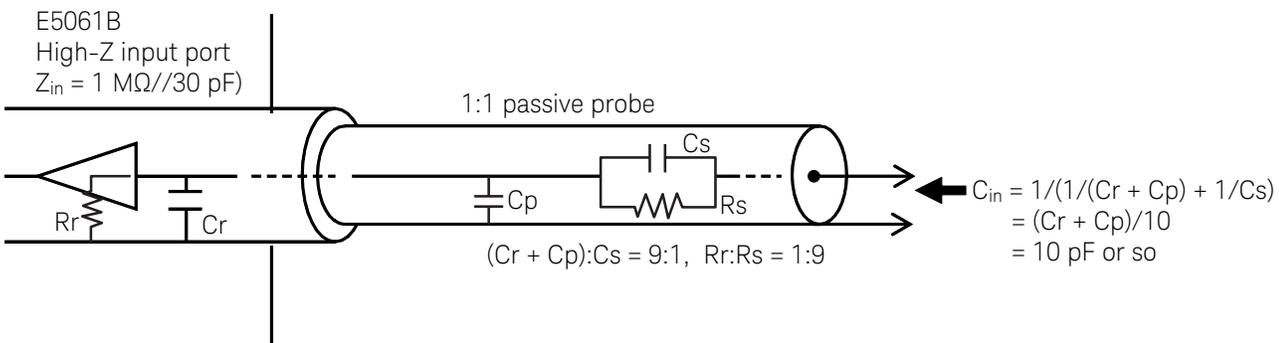

E5061B
High-Z input port
Z_{in} = 1 MΩ//30 pF)

1:1 passive probe

Cs
Cp
Rs
Rr
Cr

(Cr + Cp):Cs = 9:1, Rr:Rs = 1:9

$C_{in} = 1/(1/(Cr + Cp) + 1/Cs)$
$= (Cr + Cp)/10$
$= 10$ pF or so

Figure 13. 10:1 passive probe

E5061B
50 Ω S-parameter
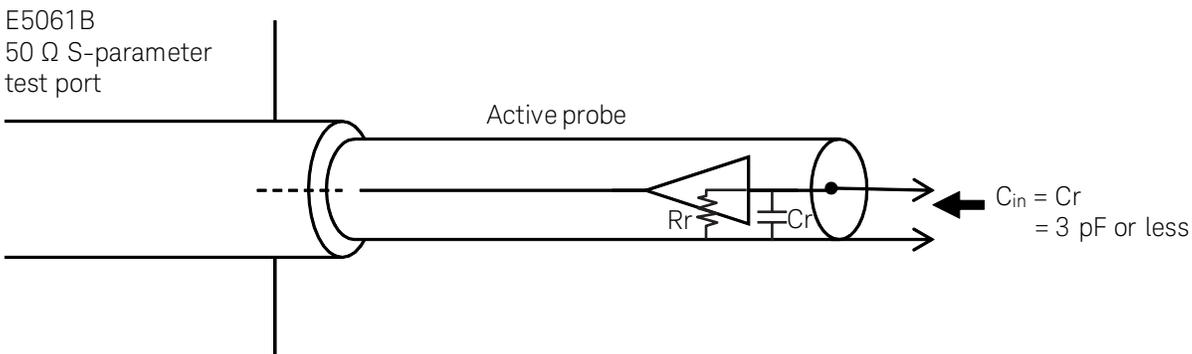test port

Active probe

Rr
Cr

$C_{in} = Cr$
$= 3$ pF or less

Figure 14. Active probe

Signal Separation for Ratio Measurements

To measure the transmission coefficient for 50 Ω devices such as passive filters in the system impedance Z0 = 50 Ω (or for devices with other Z0 values by converting the system impedance with matching circuits), the source output signal must be separated into the 50 Ω R-ch receiver and the DUT's input port. If using a source output port which does not have a built-in signal separation device, such as a built-in power splitter or a built-in directional bridge, it is necessary to separate the signal externally using an appropriate separation device.

The E5061B-3L5 has a S-parameter test port and most of 50 Ω transmission measure-ments can be covered without using the external signal separation device. But in some transmission coefficient measurements that should be covered by the gain-phase test port, such as output impedance measurements of DC-DC converters with the shunt-thru technique, you need the external signal separation device.

In general network analysis targeting linear devices, the most important requirement for the separation device is that it provides the 50 Ω source output impedance (source matching) when making the ratio measurement. The most common and recommended separation device is a two-resistor type power splitter which covers a broad frequency range from DC to GHz and provides an excellent source output impedance for the ratio measurement.

The ratio measurement using the power splitter shown in Figure 15-a is equivalent to making two measurements shown in Figure 15-b by considering the AC voltage Vo at the branch point as a virtual source voltage. As shown in this figure, the equivalent source output impedance in both R-ch and T-ch measurements will be precisely 50 Ω, which is generally an ideal source matching condition for 50 Ω network measurements.

Note that the two-resistor type power splitter is just applicable to ratio measurements and not suitable for absolute voltage measurements in the 50 Ω system impedance because the splitter's physical output impedance seen from the DUT is not 50 Ω but 83.3 Ω.
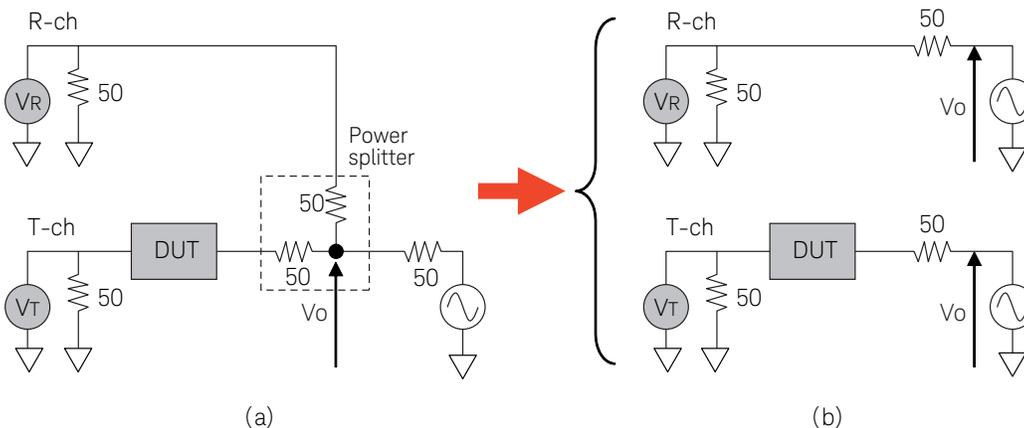

Figure 15. 50 Ω ratio measurement with a power splitter

Signal Separation for Ratio Measurements *(continued)*

Alternative separation devices to the power splitter are low-frequency directional couplers, or reactive power dividers (AC-coupled with a transformer) that have a high isolation between the two output ports (more than 25 or 30 dB). Examples are mini-circuits (**www.minicircuits.com**) ZFDC-15-6 directional coupler (0.03 to 35 MHz, BNC) or ZFSC power divider (0.002 to 60 MHz, BNC). Although their frequency range is just three or four decades and the lower frequency coverage is just down to several kHz or several tens of kHz, they are reasonable solutions if their frequency ranges meet the application needs. Due to the high isolation between two output ports, the reflected signal at the DUT's input will not directly go to the R-ch receiver and the R-ch measurement will not be affected.

Since the equivalent source matching for ratio measurements is not as good as that of two-resistor type power splitters, an attenuator pad (6 dB or so) should be connected between the output port and the DUT to improve the source matching, if necessary. The superiority of these separation devices over the power splitter is that the absolute source output impedance (port matching) is 50 Ω. This enables you to perform the absolute voltage measurements in the 50 Ω environment, although this may not be so significant in general low-frequency applications in contrast to RF applications.

A three-resistor type resistive power divider which has resistors of Z0/3 in its three arms is not applicable to the ratio measurement. Its equivalent source output impedance is not 50 Ω but 50/3 = 16.7 Ω if we consider its branch point as a virtual signal source (similar to the two-resistor type power splitter), and the isolation between output ports is small (= 6 dB). Using the three resistor type power divider in the ratio measurement will give significant measurement errors unless the DUT's input impedance is exactly 50 Ω.
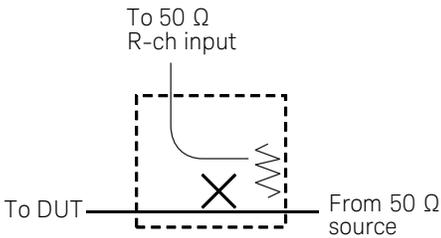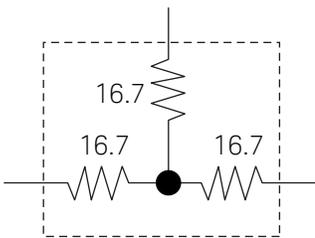

Figure 16. Directional coupler/bridge

Figure 17. Resistive power divider (Not applicable for ratio measurements)

High-Attenuation Measurement at Low Frequencies

Measurement error

Measuring high attenuation with conventional low-frequency network analyzers is likely to be affected by the errors associated with the test cable ground loop in the low frequency below 100 kHz. These errors are critical in applications such as CMRR and PSRR measurements of low-frequency amplifiers. The most significant problem is the error caused by the cable shield resistance (braid resistance), which is not negligible in the low-frequency range below 100 kHz.

Figure 18 illustrates a high-attenuation measurement with the network analyzer. As the DUT's attenuation is very high, the DUT's output voltage Vo will be very small. Ideally, the measured AC voltage at the receiver VT should be also Vo.

However, in the low-frequency range, external common mode noise is likely to flow into the test cable ground loop between the source and receiver as shown in Figure 18, and voltage Vc2 appears across the resistance of the cable outer shield, Rc2. The voltage Vc2 will cause a voltage measurement error at the receiver VT, because the voltage to be measured, Vo, is very small. As a result, the measured attenuation will be erroneous.

Depending on the phase relationship between Vo and Vc2, the measured attenuation can be higher, or lower than the DUT's true attenuation value. Or, sometimes the error appears as a dip in the trace.
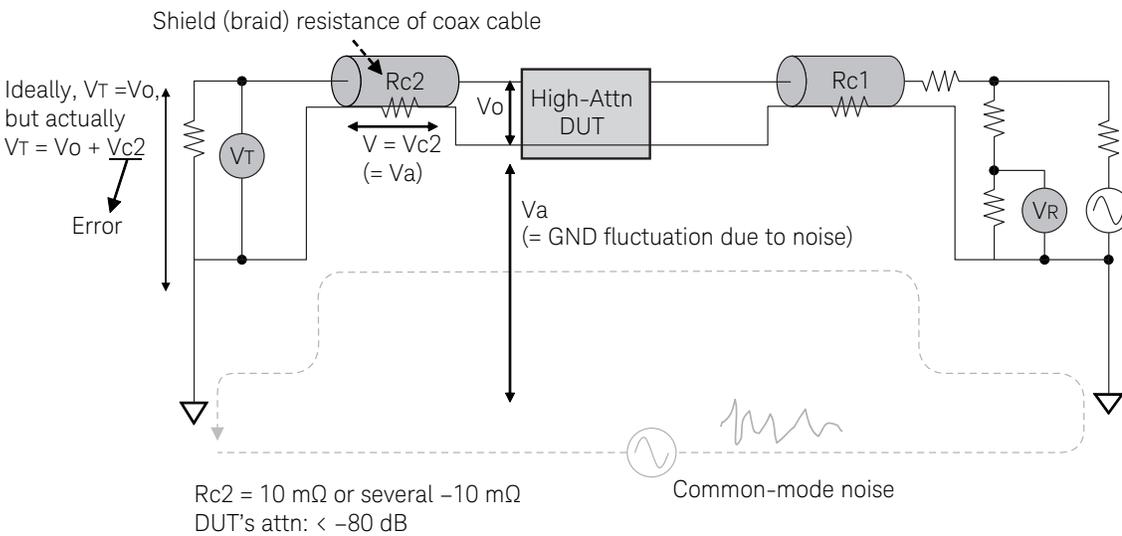

Figure 18. Measurement error due to cable shield resistance (1)

High–Attenuation Measurement at Low Frequencies *(continued)*

Measurement error *(continued)*

The test cable ground loop cause another measurement error in the low-frequency range. You can assume the DUT has a shunt signal path with a small impedance Zsh. Typical examples are mΩ impedance measurements for the low-frequency PDN components such as DC-DC converters and large bypass capacitors using the shunt-thru technique.

Ideally, the source signal flowing through the DUT should return to the source side through the outer shield of the cable.

However, the source current also flows into the test cable shield of the T-ch receiver-side in the low-frequency range. Similarly to the case of the common-mode noise, the source current flowing into the T-ch cable shield will cause the voltage Vc2 across cable outer shield resistance Rc2, and this will cause the error in the measurement at the receiver VT. In this case, the measured attenuation will be larger than the DUT's true attenuation value.

Note that these measurement errors associated with the test cable ground loop occurs only in the low frequency below 100 kHz. In the higher frequency range, the coaxial test cable's inductance act like a common-mode choke (balun) described in the next section, and the error current does not flow through the shield of the VT receiver side.
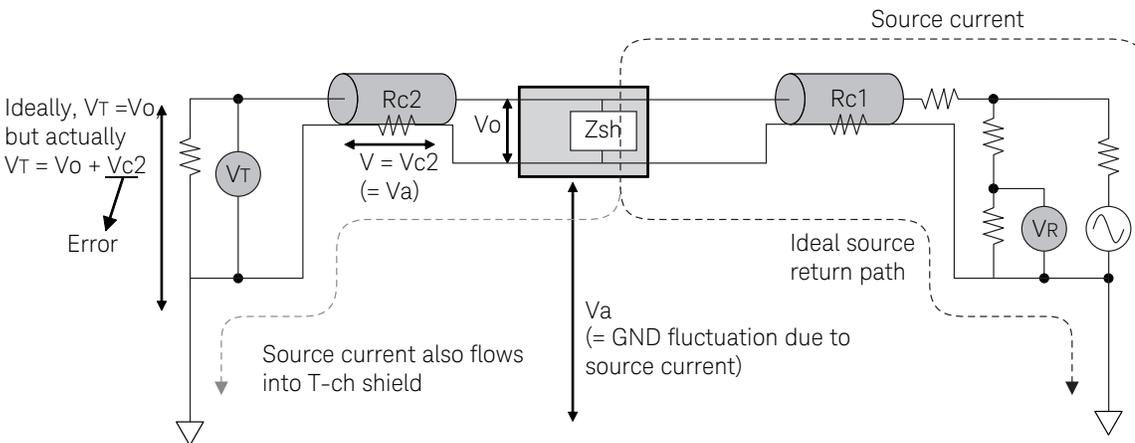

Figure 19. Measurement error due to cable shield resistance (2)

High–Attenuation Measurement at Low Frequencies *(continued)*

Conventional solution

There are several techniques to minimize these measurement errors. The most common approach is to clamp magnetic cores to the test cables or wrap the test cables several times around magnetic cores. The equivalent circuit of using magnetic cores is shown in Figure 20. The magnetic core increases the shield impedance and suppresses the current from flowing through the cable shield, while not affecting the signal that flows in the center conductor and returns in the cable shield.

The shield impedance caused by the core's self inductance reduces the common mode noise current flowing through the ground loop and the source current flowing into the shield of the VT receiver side. Also, the core attached at the source side forces the source current to return through the shield, back to the source side.

However, it is not easy to implement this solution, because we need to find a good magnetic core that has very large inductance (high permeability) to fully eliminate the errors down to the low-frequency range. Also, sometimes it is difficult to judge whether the core is sufficiently working or not, especially if the DUT's attenuation characteristic is not flat.

An example toroidal core recommended for this application is Metglas Finemet F7555G (Φ 79 mm) **www.metglas.com**.
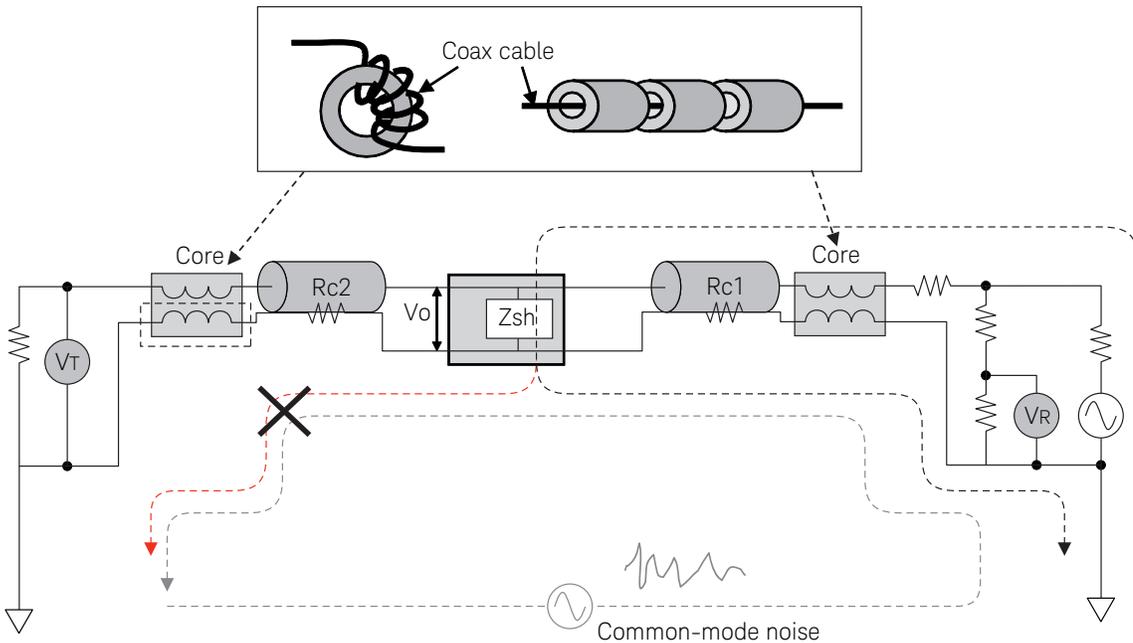

Figure 20. Solution with magnetic cores

High-Attenuation Measurement at Low Frequencies *(continued)*

Solution with E5061B-3L5

The gain-phase test port (5 Hz to 30 MHz) of the E5061B-3L5 has a unique hardware architecture that enables you to eliminate the measurement error associated with the source-to-receiver test cable ground loop. Figure 21 shows a simplified block diagram when using the gain-phase test port. The receivers are semi-floated with the impedance |Zg|, which is about 30 Ω in the low-frequency range below 100 kHz. Similarly to the case of using the magnetic core, we can intuitively understand that the shield current is blocked with the impedance |Zg|. Or, if you assume that the voltage swing at the DUT's ground side is Va as shown in Figure 21, since Rshield is much smaller than the receiver input impedance 50 Ω, VT is approximately derived as follows,

$$VT = Vc2 + Vo$$
$$= Va \times Rc2/(Rc2+Zg) + Vo$$

Since Rc2 << |Zg|, the first term of the above equation is negligible. So VT will be almost Vo that we want to measure. Thus the DUT's high attenuation or mΩ shunt impedance can be correctly measured by minimizing the effect of the shield resistance. The E5061B's gain-phase test port enables easy and accurate high-attenuation measurements in the low-frequency range.

On the other hand, the S-parameter test port of the E5061B-3L5 has a normal grounded receiver architecture, like most of other existing low-frequency network analyzers. If you want to measure the low-frequency high-attenuation devices with the S-parameter test port (for example, if you want to measure them up to more than 30 MHz, that cannot be covered with the gain-phase test port), it is necessary to use the magnetic cores to eliminate the errors caused by the test cable ground loop.
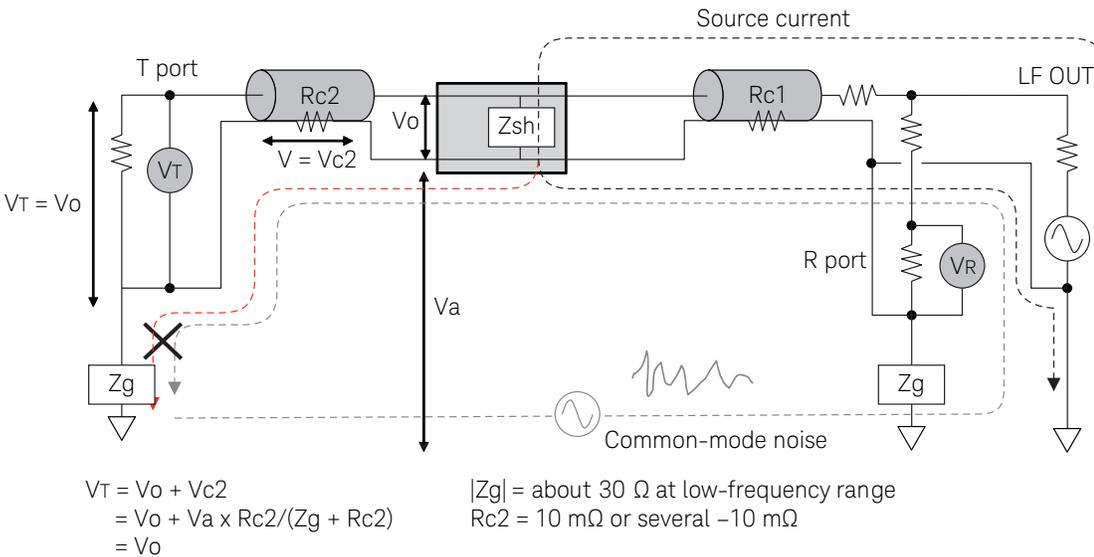

$$VT = Vo + Vc2$$
$$= Vo + Va \times Rc2/(Zg + Rc2)$$
$$= Vo$$

$$|Zg| = \text{about } 30 \ \Omega \text{ at low-frequency range}$$
$$Rc2 = 10 \text{ m}\Omega \text{ or several } -10 \text{ m}\Omega$$

Figure 21. Solution with E5061B-3L5 gain-phase test ports

High-Attenuation Measurement at Low Frequencies (continued)

Effectiveness of gain-phase test port

Figure 22 shows the transmission measurement results of a 90 dB coaxial attenuator with E5061B's Sparameter test port and the gain-phase test port. The test frequency range is 100 Hz to 10 MHz. The traces in the channel-1 (left side) are the measurement results with the S-parameter test port. As you can see, the measurement result without a magnetic core indicates incorrectly higher values in the low-frequency range, which is the error caused by the test cable ground loop between the source and receiver. Another trace in the same graph is the measurement result by attaching a clamp-on type magnetic core to the test cable. The measurement in the low-frequency range is slightly improved, but is still not accurate in the very low-frequency range.

On the other hand, the trace in the channel-2 (right side) is the measurement result using the gain-phase test port. As you can see, –90 dB attenuation is correctly measured down to 100 Hz, without being affected by the test cable ground loop.
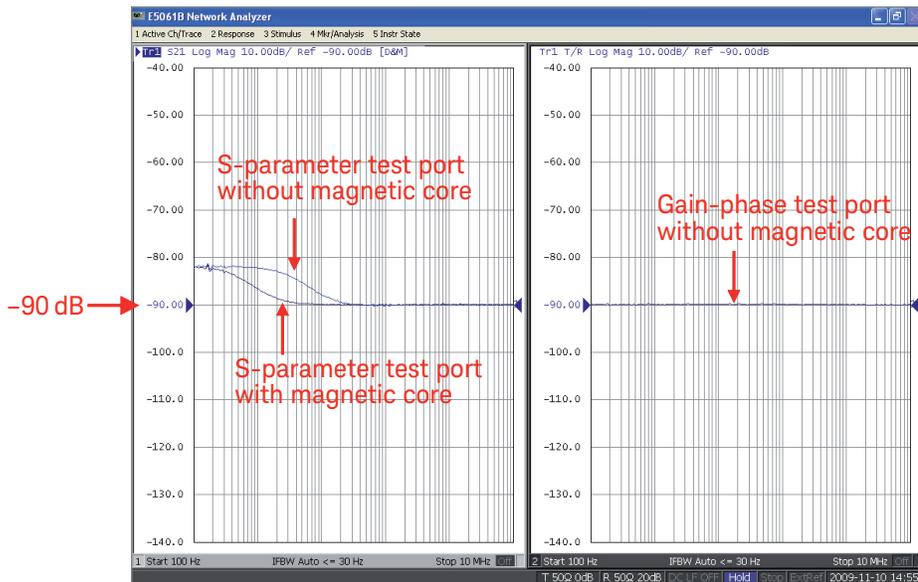

Figure 22. Comparison of the measurement results among 3 different configurations

OP Amp Measurement Example

Closed-loop gain

The following sections show measurement examples of various frequency response characteristics of operational amplifiers.

Figure 23 shows a configuration example of measuring the closed-loop gain of a simple inverting amplifier (Av = –1) using the gain-phase test port (up to 30 MHz).

To minimize the influence of the probing capacitance to the amplifier's loading condition, it is recommended to use the 10:1 probes, which give relatively small input capacitances.

To accurately measure the gain and phase characteristics, the response thru calibration should be performed by connecting the T-ch probe to the point TP1, so that the gain and phase errors between two probes are eliminated.
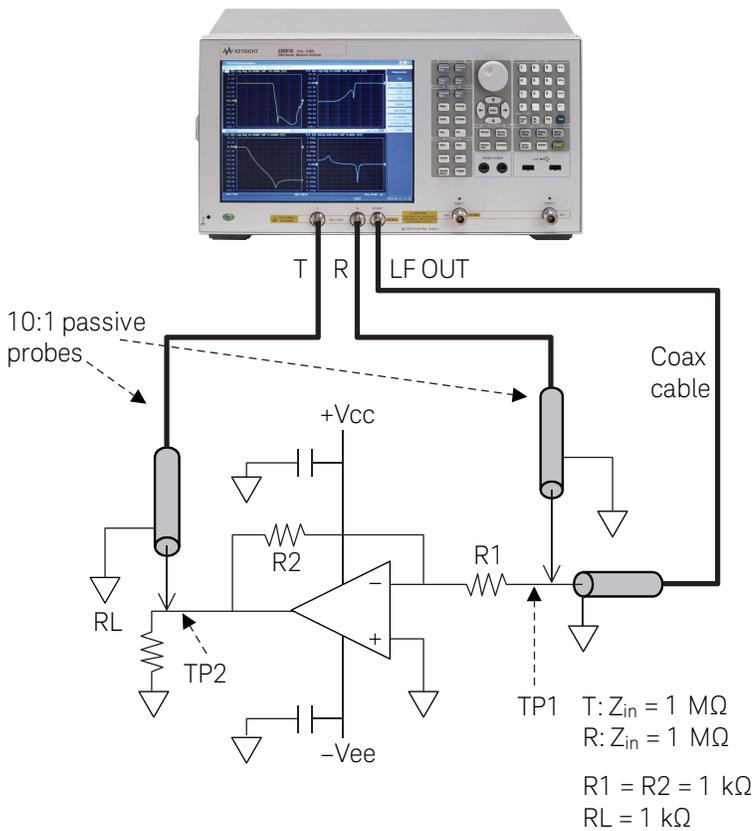

Figure 23. Configuration example of closed-loop gain measurement with the gain-phase test port

OP Amp Measurement Example *(continued)*

Closed-loop gain *(continued)*

If you want to measure the amplifier's frequency response up to more than 30 MHz, the solution is to use the S-parameter test port and the active probe. Figure 24 shows a configuration example. The response thru calibration by probing the point TP1 is necessary, because the R-ch receiver has 50 Ω input and we need to set the reference to TP1 so that the voltage transfer function from the DUT's intput and output ports can be measured.

Figure 25 shows a closed-loop gain measurement example of a high-speed operational amplifier by using the E5061B's S-parameter test port and the 41800A active probe. The marker is put on the –3 dB cutoff frequency, which indicates the bandwidth of this amplifier circuit is approximately 20 MHz.
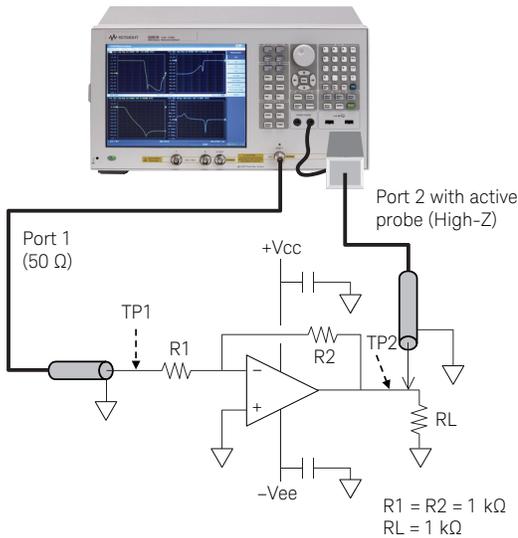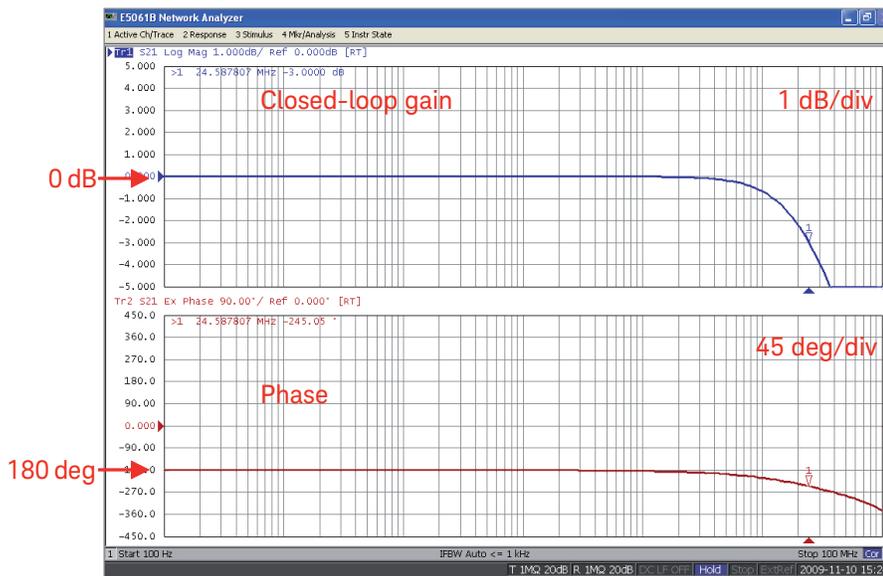

Figure 24. Configuration example of closed-loop gain measurement with the S-parameter test port

Frequency = 100 Hz to 100 MHz
Source level = 0 dBm
IFBW = AUTO (Upperlimit = 1kHz)

Figure 25. Closed-loop gain measurement example

OP Amp Measurement Example *(continued)*

Open-loop gain

There are several methods for measuring the open-loop gain of OP amps. The most common method is to measure the voltage ratio V_T/V_R in the circuit, Figure 26. Assuming that the open-loop gain of the OP amp is A, if we look at the current $Ir2$, the following equation can be derived:

$$(V_T-V_R)/R2 = \{V_T-(-A \times V_R)\}/Z_{out}$$

If Zout << R2, the voltage ratio V_T/V_R can be calculated as follows:

$$V_T/V_R = (-A-Z_{out}/R2)/(1-(Z_{out}/R2))$$
$$= -A$$

In the case of high-gain OP amps, if the closed-loop gain Av is small (e.g. Av = −R2/R1 = −1), the voltage V_R will be too small to be accurately measured, especially in the low-frequency range where the open-loop gain is very high.

In the linear operating region, if the closed-loop gain Av is increased, the voltage V_R will also be increased proportionally and the measurement will be easier for the analyzers. For example, if |Av| = R2/R1 = 10, V_R will be 10 times (= 20 dB) higher than the case of |Av| = 1. Here it should be noted that V_T will also be 20 dB higher and you'll need to avoid the receiver from overloading when measuring V_T. Also the measurement in the higher frequency range will be inaccurate, because the linear region of the amplifier circuit is narrower if Av is high. The open-loop gain measurement can be implemented by using either the dual probing method, or single probing method.
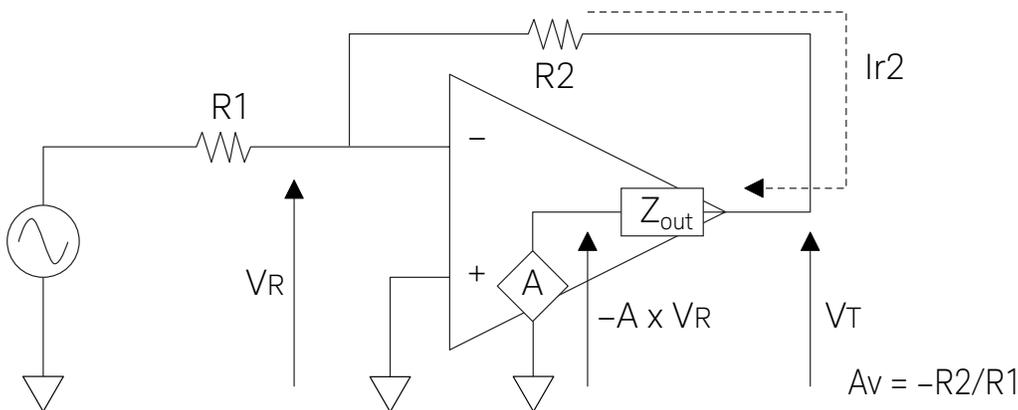

Figure 26. Configuration example of open-loop gain measurement

OP Amp Measurement Example *(continued)*

Open-loop gain *(continued)*

Figure 27 shows a measurement configuration with the gain-phase port. The ratio measurement T/R will directly indicate the open-loop gain A. To accurately measure the phase response without affecting the loading condition due to the large probing capacitance, 10:1 passive probes should be used, rather than coaxial test cables.
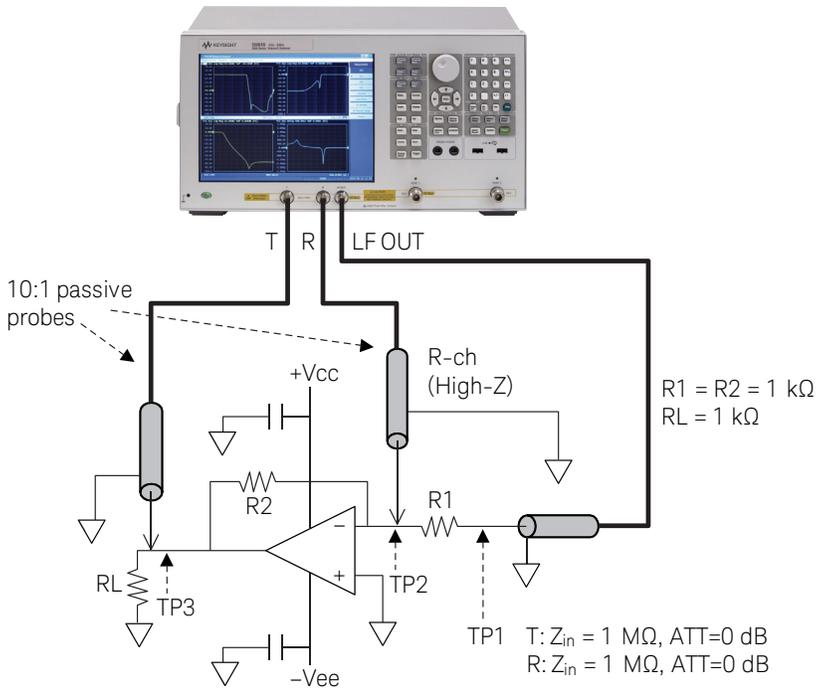

Figure 27. Configuration example of open-loop gain measurement with the gain-phase test port
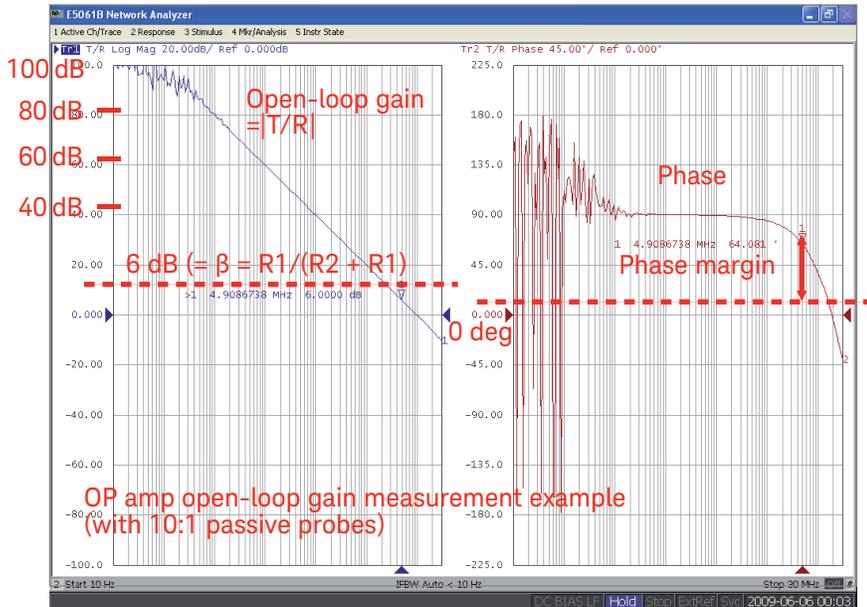
OP Amp Measurement Example *(continued)*

Open-loop gain *(continued)*

Figure 28 shows a open-loop gain measurement example of an OP amp in the unity gain condition (R1 = R2 = 1 kΩ) using the gain-phase configuration shown in Figure 27. The test frequency range is from 10 Hz to 30 MHz. The phase margin can be derived from these measurements. By simply calculating the transfer function of the feedback path as β = R1/(R1 + R2) = 1/2 = –6 dB (assuming no phase shift), crossover point where the loop gain |–A x **β**|=0 dB can be found by placing the marker at the +6 dB point. And the phase margin can be directly given by the marker on the phase trace, as we are looking at the loop transfer function –A x **β** which includes the 180 degree inversion at the OP amp input port.

The trace fluctuation in the high gain area is due to the dynamic range degradation caused by the 20 dB loss of the passive probe. Since we are measuring the open-loop gain with the amplifier of unity gain, the measured AC voltage at the R-ch receiver will be extremely small in the high gain area and this causes the trace fluctuation. The trace fluctuation in the high gain area is not a problem for evaluating the phase margin from the measurement data in the lower gain area.

But, if you also want to measure the very high gain in the low-frequency range, separately perform another open-loop gain measurement using coaxial test cables instead of 10:1 passive probes. The receiver attenuator setting should be 0 dB at the R-port, and 20 dB at the T-port, so that the very small voltage at the R-ch receiver can be measured with good SNR. Note that this measurement configuration is applicable only in the low to middle frequency range where the open-loop gain is high and the voltage appearing at the R-ch receiver will not exceed the receiver's maximum input level with the attenuator setting of 0 dB.

Frequency = 100 Hz to 100 MHz
Source level = 0 dBm
IFBW = AUTO (100 Hz limit)

Figure 28. Open-loop gain and phase measurement example with the gain-phase test port

OP Amp Measurement Example *(continued)*

Open-loop gain *(continued)*

If the OP amp's open-loop gain must be measured up to more than 30 MHz, the solution is the S-parameter test port with the active probe. Since the S-parameter test port allows us to use just a single active probe, you'll need to use a two step measurement technique. The procedure is as follows:

1. Perform the response through calibration by probing TP1.
2. Measure S21 by probing TP2 and memorize the trace data with DATA -> MEM function (Step 1).
3. Connect a dummy capacitor to TP2, and measure S21 by probing TP3 (Step 2).
4. Calculate Data/Memory with the data math function to obtain the open-loop gain.

The dummy capacitor connected in step 2 duplicates the probing capacitance of step 1 which affects the open-loop phase measurement in the high-frequency range. Its capacitance should be about the same capacitance as the active probe's input capacitance.

If you need to measure a very high open-loop gain, it might be better to attach magnetic cores to the test cable to eliminate the measurement errors associated with the ground loop, which may affect the small signal measurement of the step 1.
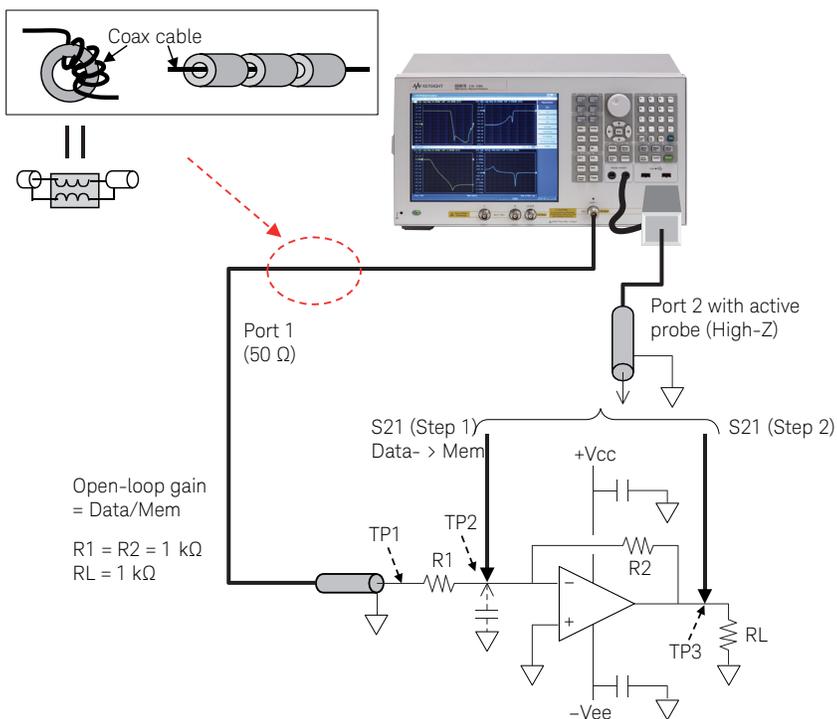

Figure 29. Configuration example of open-loop gain measurement with a single active probe

OP Amp Measurement Example *(continued)*

Open-loop gain *(continued)*

Figure 30 shows a measurement example of the open-loop gain and phase with the configuration of Figure 29. Trace 1 is the measured response by probing TP2, which indicate the ratio of the input voltage and the attenuated voltage at TP2. Trace 2 is the measured response by probing TP3, which is the closed-loop gain and phase. And, trace 3 is the open-loop gain and phase calculated from these measurement results. The results are calculated by using the trace math function (data/memory).

As previously described, the phase margin is indicated by the phase measurement value at the 6 dB open-loop gain point, where the loop gain is 0 dB. In this example, the phase margin is about 86 degrees.

Trace 3: Open-loop gain (data/memory)
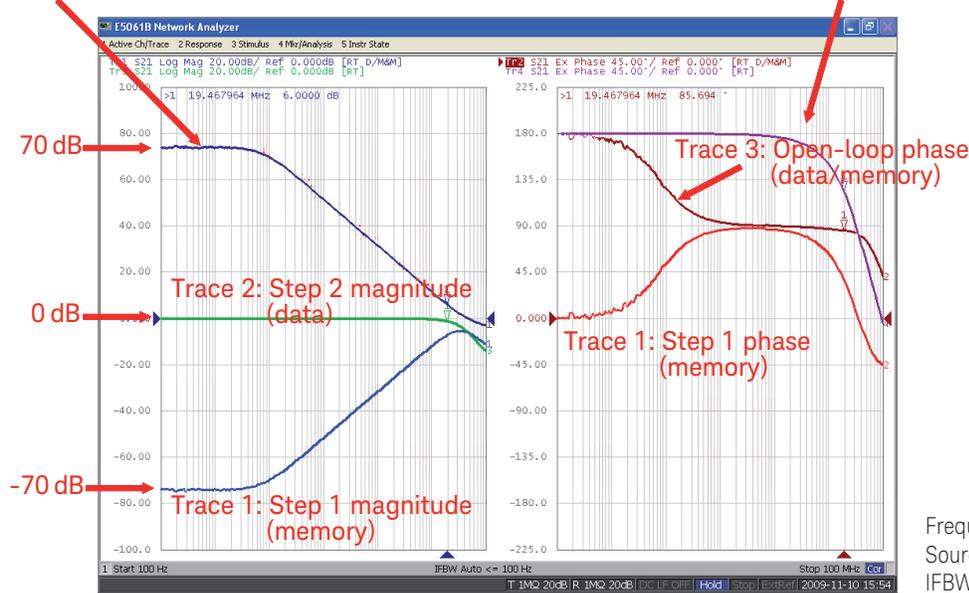
Trace 2: Step 2 phase (data)

70 dB

0 dB

-70 dB

Trace 2: Step 2 magnitude (data)

Trace 1: Step 1 magnitude (memory)

Trace 3: Open-loop phase (data/memory)

Trace 1: Step 1 phase (memory)

Frequency = 100 Hz to 100 MHz
Source level = 0 dBm
IFBW = AUTO (100 Hz limit)

Figure 30. Open-loop gain and phase measurement example with a single active probe

OP Amp Measurement Example *(continued)*

CMRR

The CMRR (Common-mode Rejection Ratio) of OP amps and other low-frequency differential amplifiers is generally difficult to measure, because you need to measure very high-attenuation for the common-mode input. The CMRR is defined as CMRR = Ad/Ac, where Ad is the differential-mode gain and Ac is the common-mode gain. Figure 31 shows the configuration using the gain-phase test port. To measure high attenuation, the coaxial test cables are used for connecting the receivers to the DUT, instead of 10:1 passive probes which have 20 dB loss.

The common-mode gain (attenuation) Ac is measured by turning SW1 to the A-position. The differential gain Ad is measured by turning the switch SW1 to the B-position. Then the CMRR is calculated as Ad/Ac (= 20 x Log (Ad/Ac) in dB). The differential gain of this circuit is |Ad| = R2/R1 = 10. Accordingly, the common-mode gain Ac is 10 times (20 dB) larger than the case of |Ad|=1. This allows the analyzer to measure high CMRR over 100 dB.

Due to the semi-floating receiver architecture of the gain-phase test port, you can accurately measure high CMRR by eliminating the error associated with the test cable ground loop.
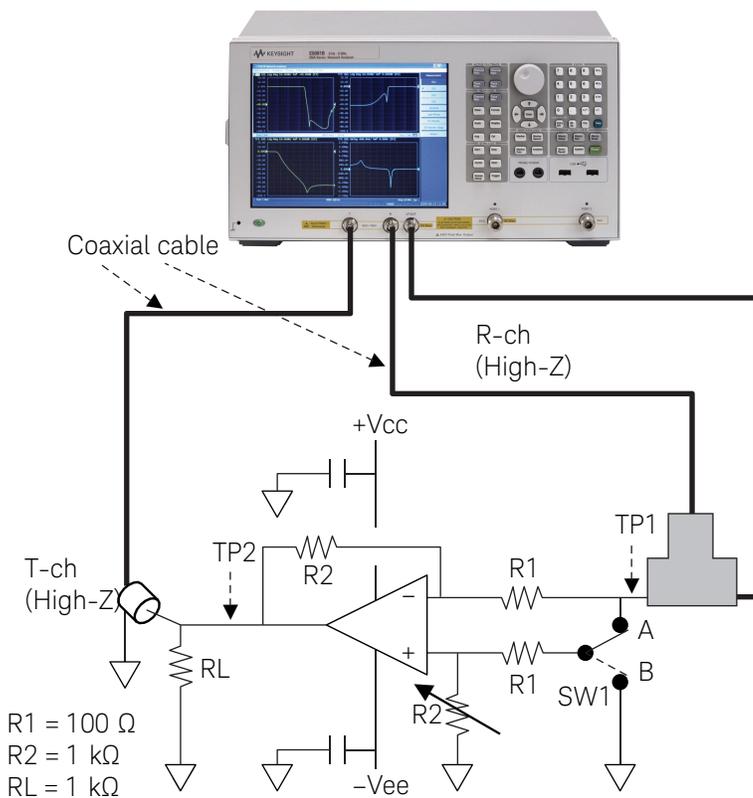

Figure 31. Configuration example of CMRR measurement with the gain-phase test port

OP Amp Measurement Example (continued)

CMRR (continued)

CMRR measurement higher than 30 MHz can be performed by using the S-parameter test port and the active probe. In this case, magnetic cores should be attached to the test cable as shown in Figure 32 to eliminate the measurement error caused by the common mode noise. One example of magnetic core is Metglas Finemet F7555G (Φ 79 mm: www.metglas.com)

Figure 33 shows the measurement example with gain-phase test port. Trace 1 is the common-mode gain Ac, and Trace 2 is the differential-mode gain Ad (= 20 dB). The common-mode gain Ac of about –90 dB is accurately measured by eliminating the ground loop effects. Trace 3 is the CMRR calculated from these measurement results. The marker indicates that the CMRR at 100 kHz is about 80 dB. In the lower frequency range, the CMRR is more than 100 dB.
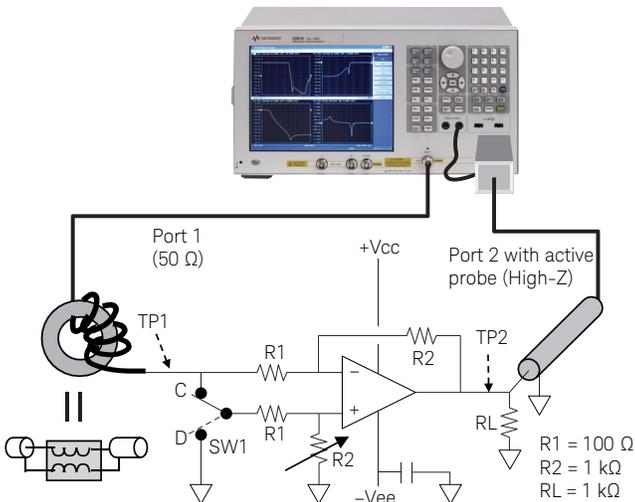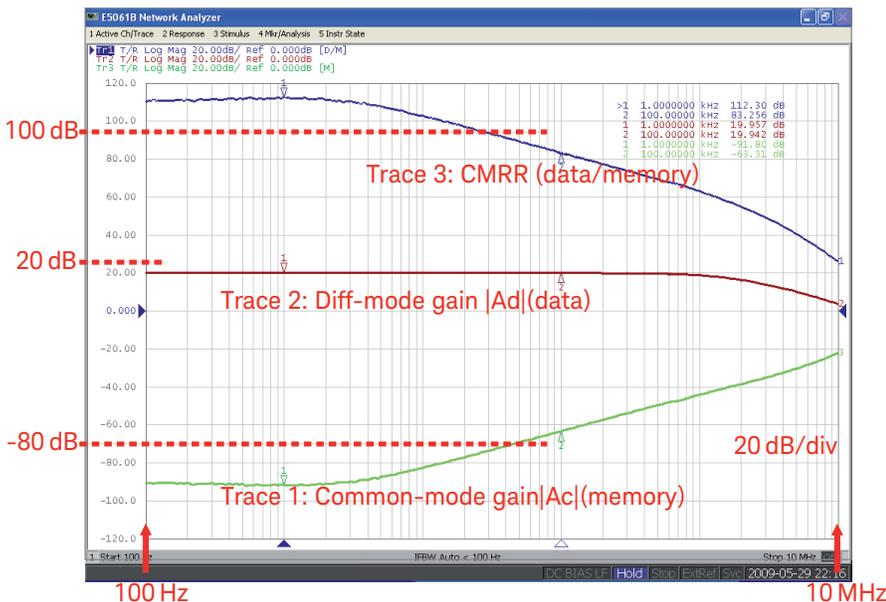

Figure 32. Configuration example of CMRR measurement with the S-parameter test port

Frequency = 100 Hz to 100 MHz
Source level
    for Ac measurement): 0 dBm
    for Ac measurement): –15 dBm
IFBW = Auto (100 Hz limit)
Receiver ATT setting
    Ac measurement: 20 dB (R-ch)
              0 dB (T-ch)
    Ad measurement: 20 dB (R-ch and T-ch)
The balance of R1 and R2 is not fully optimized in this measurement example.

Figure 33. CMRR measurement example with the gain-phase test port

OP Amp Measurement Example *(continued)*

PSRR

The PSRR (Power Supply Rejection Ratio) of amplifiers is another difficult parameter to measure as it requires high-attenuation measurements. Here we consider the definition of PSRR = Av/Ap, where Av is the closed-loop gain of the amplifier circuit and Ap is the gain from the power supply port (positive or negative) to the output port. Similarly to the CMRR measurement, Ap is proportional to Av in the linear operating region.

Figure 34 shows a configuration example of measuring the PSRR (positive PSRR) by using the gain-phase test port. Since |Av| = R2/R1 = 1, the measured gain of this circuit directly indicates the inverse of the OP amp's PSRR (= 1/Ap, which is a negative dB value). The source signal is applied to the positive power supply port with a DC bias voltage. The E5061B has a built-in DC bias source which enables you to internally superimpose the DC voltage bias onto the AC source signal.
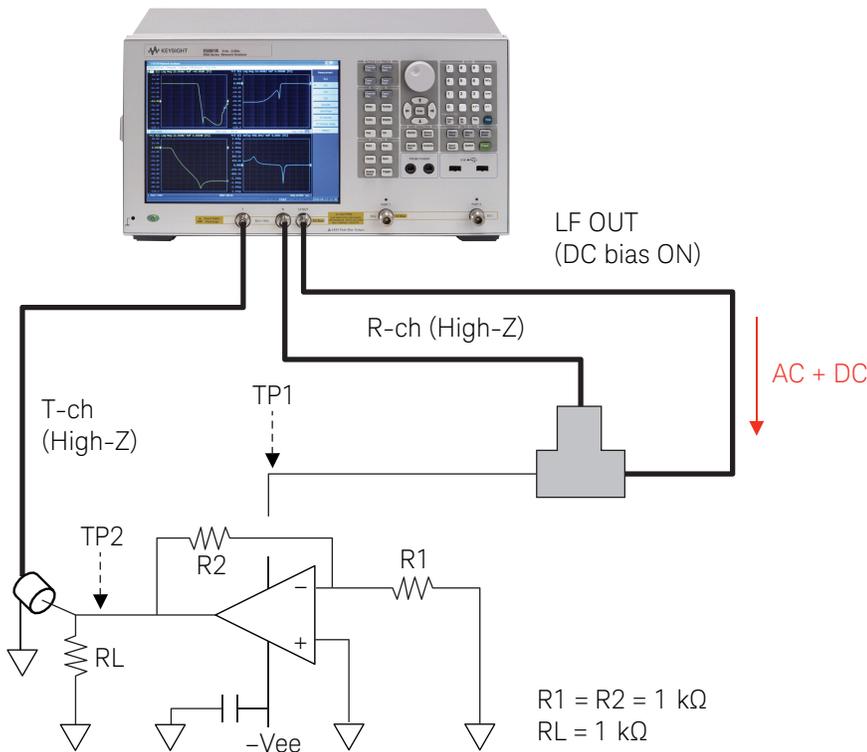

Figure 34. Configuration example of PSRR measurement with the gain-phase test port

OP Amp Measurement Example *(continued)*

PSRR *(continued)*

PSRR measurements higher than 30 MHz can be performed by using the S-parameter test port and the active probe. Similarly to the CMRR measurement with the S-parameter test port, it is recommended to attached magnetic cores to the test cable to eliminate the measurement error associated with the test cable ground loop. Figure 36 shows a PSRR measurement example with the gain-phase test port. The marker indicates that the PSRR at 1 kHz is –87 dB. E5061B-3L5 has a DC monitor function, and you can check the DC voltage level that is actually applied to the DUT.
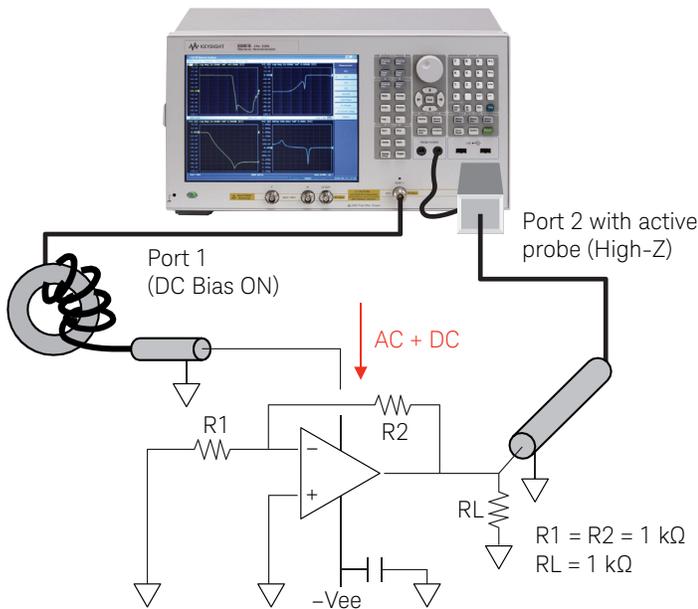

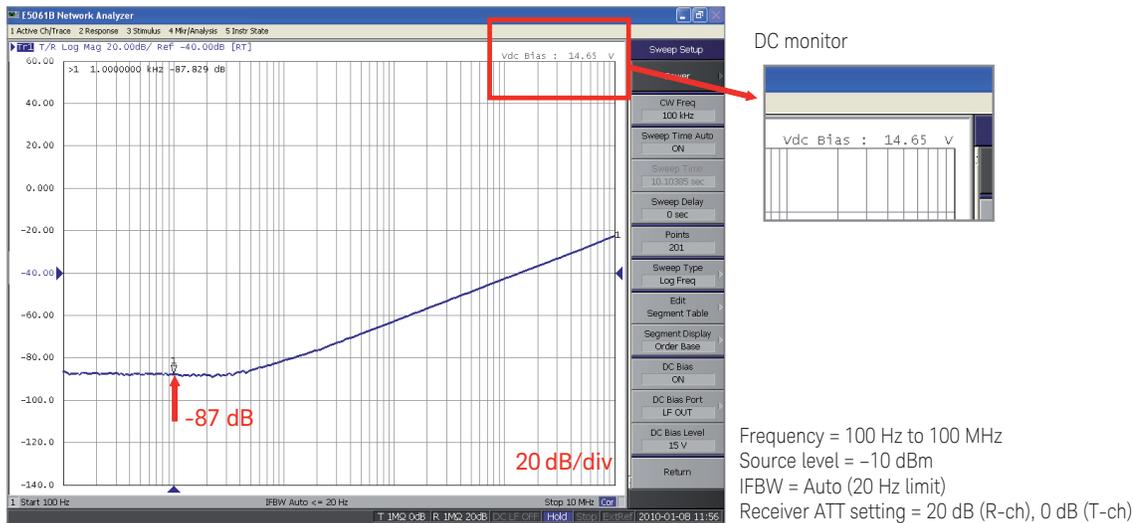Figure 35. Configuration example of PSRR measurement with the S-parameter test port

Frequency = 100 Hz to 100 MHz
Source level = –10 dBm
IFBW = Auto (20 Hz limit)
Receiver ATT setting = 20 dB (R-ch), 0 dB (T-ch)

Figure 36. PSRR measurement example with the gain-phase test port

OP Amp Measurement Example *(continued)*

Output impedance

This is not a 2-port transmission measurement, but a 1-port impedance measurement. In general, OP amps have closed-loop output impedances that range from several tens of mΩ at low frequencies up to 100 Ω at high frequencies. To fully cover this impedance range, the reflection measurement method is the proper solution. Figure 37 shows a configuration example of measuring the closed-loop output impedance of OP amps. The open/short/load 3-term calibration (1-port full calibration) must be performed.
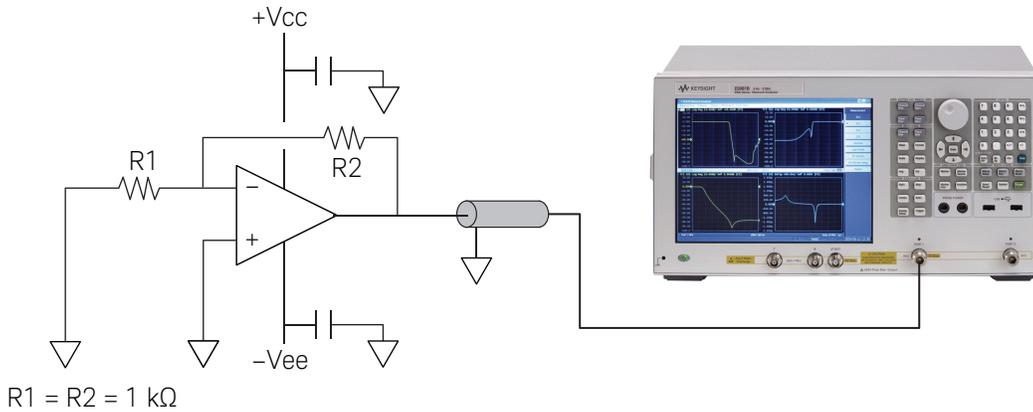

Figure 37. Configuration example of output impedance measurement

Figure 38 is a measurement example of the closed-loop output impedance. The measured traces show the impedance magnitude that are plotted by using the impedance conversion function. The trace shown on the left side indicates the output impedance in logarithmic scale [20 x log |Z| dB]. The trace shown in the right side indicates the output impedance in linear scale [Ω].


Figure 38. Output impedance measurement example

References

[1] Robert A. Witte, *Spectrum and Network Measurements*

[2] Willy M. Sansen, Michael Steyaert, Paul J. V. Vandeloo, *Measurement of Operational Amplifier Characteristics in the Frequency Domain*, IEEE Transaction on Instrumentation and Measurement, Vol. IM-34, No.1, March 1985

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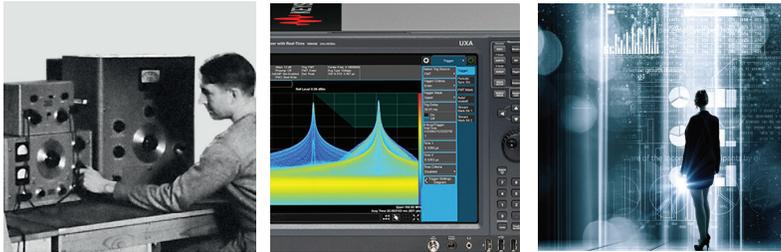
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