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Measuring Frequency Response with the LF Network Analyzer

Application Note

Introduction

This application note describes basics of frequency response measurements with a LF network analyzer, which is generally referred to as gain-phase measurements or LF network measurements. Here we discuss topics unique to LF network measurements such as high-impedance probing measurements, examples of typical low-frequency 2-port devices, and other measurement tips.

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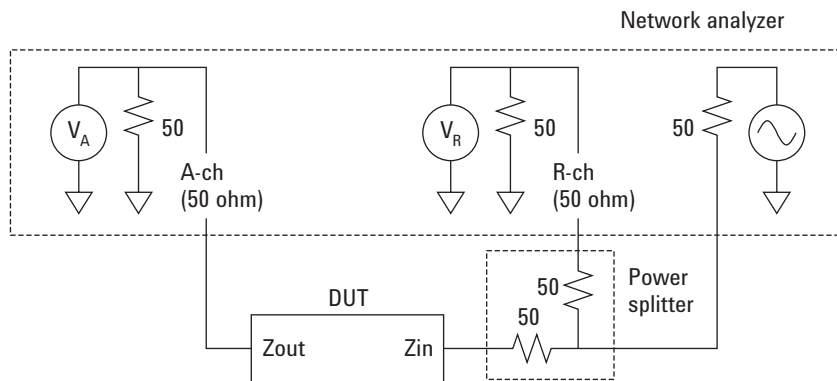## 50 ohm DUTs

First let's summarize how to connect DUTs in typical applications. Here the focus is on configurations for 2-port transmission measurements.

Figure 1 shows a configuration which is commonly used for measuring transmission characteristics of 50 ohm devices. The R-ch receiver V_R monitors the source output voltage applied to the 50 ohm system impedance (incident voltage to the 50 ohm transmission line), and

the A-ch receiver V_A monitors the transmitted voltage. The analyzer measures the voltage ratio V_A/V_R which indicates the transmission coefficient.

Figure 2 is a configuration that is not common for measuring 50 ohm devices, but has been used in some specific applications with old network analyzers or vector voltmeters. The main difference from the configuration of Figure 1 is that

the R-ch receiver V_R directly monitors the voltage that appears across the DUT's input impedance Zin with high-impedance probing, instead of monitoring the voltage across the 50 ohm system impedance, and the ratio V_A/V_R indicates the voltage transfer function Vout/Vin. Due to this, the measurement results will differ from the configuration of Figure 1 if Zin is not exactly 50 ohm in the entire measurement frequency range.

Calibration: Response thru cal. by connecting thru device in place of DUT

Figure 1. Configuration for measuring transmission coefficient of 50 ohm DUTs
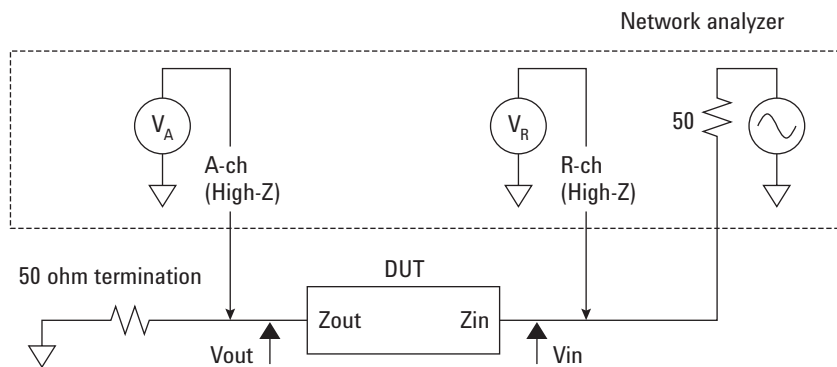
Calibration: Response thru cal. by connecting thru device in place of DUT

Figure 2. Configuration for measuring voltage transfer function

Basic Measurement Configurations
Non-50 ohm DUTs, Example 1

Low-frequency 2-port devices often have non-50 ohm impedances.

Figures 3 and 4 shows configuration examples for measuring a low-frequency amplifier whose output port is terminated with a non-50 ohm load ZL. The load impedance ZL depends on requirements of the targeted applications. 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 terminated with ZL. The voltage across ZL can be monitored with high-impedance probing at the A-ch receiver without affecting the DUT's load condition.

The difference between these two configurations is the input impedance of the R-ch receiver. The configuration of Figure 3 uses the 50 ohm input along with the power splitter, and the configuration of Figure 4 uses the high-impedance input. Generally both configurations will give the same measurement result. Note that the source power level applied to the DUT's input port in the configuration of Figure 3 is 6 dB lower than that of Figure 4 because of the insertion loss at the power splitter.

To compensate the frequency response errors of the probes and test cables, the response through calibration should be performed by contacting the A-ch probe to the point TP1 in both configurations.

If the amplifier's input impedance is very high, generally it is recommended to connect a 50 ohm feed through to the DUT's input so that the impedance seen from the analyzer's source will be about 50 ohm, the standing wave in the high frequency range is prevented. However, you can omit this 50 ohm feed through if the input impedance is not so extremely large (e.g. less than several kohm) and the measurement frequency range is below tens of MHz, because impedance matching is not as critical in this measurement application. Moreover, if the measured voltage is very small and the measurement error due to the ground loop problem is of concern, it would be preferred not to connect the feed through to avoid making an unwanted signal leakage path to the ground loop, as described later. In this case, the response through calibration by probing TP1 is necessary so that the ratio measurement is referenced to the point TP1 and the voltage transfer function across the DUT can be measured.
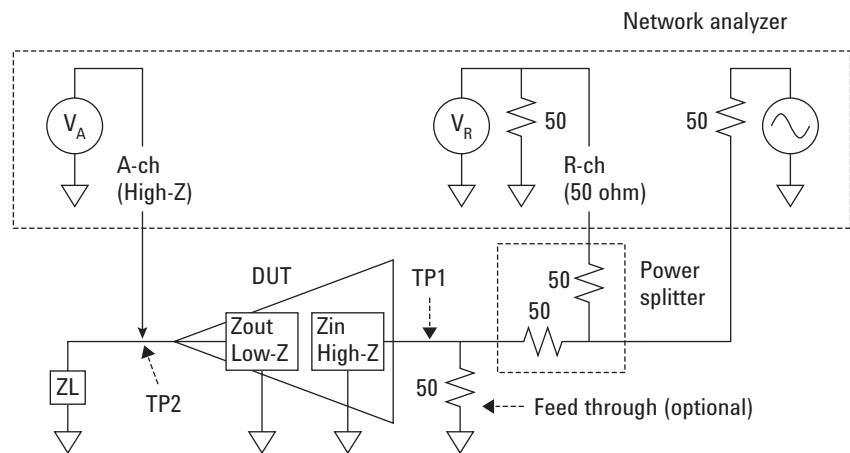
Calibration: Response thru cal. by contacting A-ch probe to TP1

Figure 3. Configuration for measuring amplifiers (1)
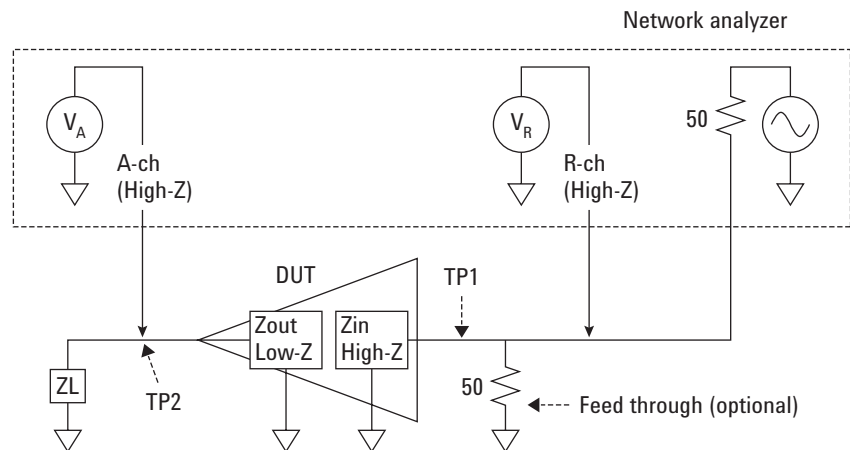
Calibration: Response thru cal. by contacting A-ch probe to TP1

Figure 4. Configuration for measuring amplifiers (2)

Basic Measurement Configurations
Non-50 ohm DUTs, Example 2

Figure 5 shows a configuration example for measuring 2-port devices whose input and output impedances are several hundreds of ohms to 1 or 2 kohm. Typical devices are low-frequency passive filters such as ceramic filters and LC filters. In this example, impedance matching is implemented by simply connecting a series resistor. In Figure 5, the ratio V_A/V_R indicates the transmission coefficient for the 1 kohm 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 as small as possible not to affect the filter's characteristics.

The equivalent measurement can be achieved by using the 50 ohm input instead of using high-impedance probing at the A-ch and connecting

another matching resistor as shown in Figure 6. This configuration is simpler and has an advantage that no probe capacitance is applied at the A-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.
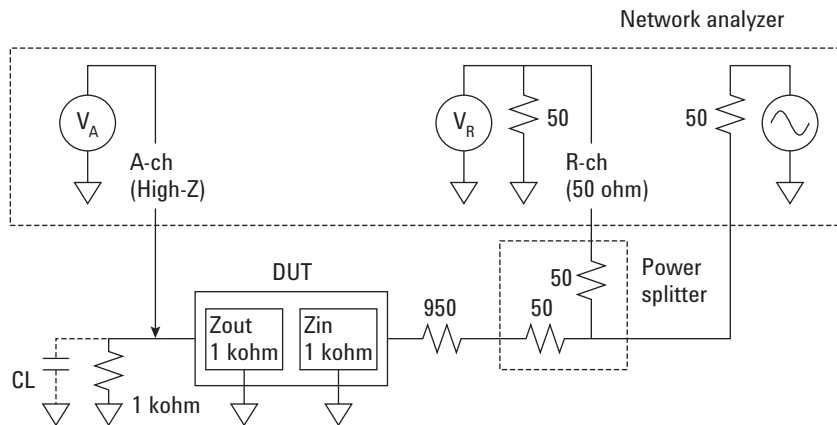
Calibration: Response thru cal. by connecting thru device in place of DUT

Figure 5. Configuration for measuring passive IF filters (1)
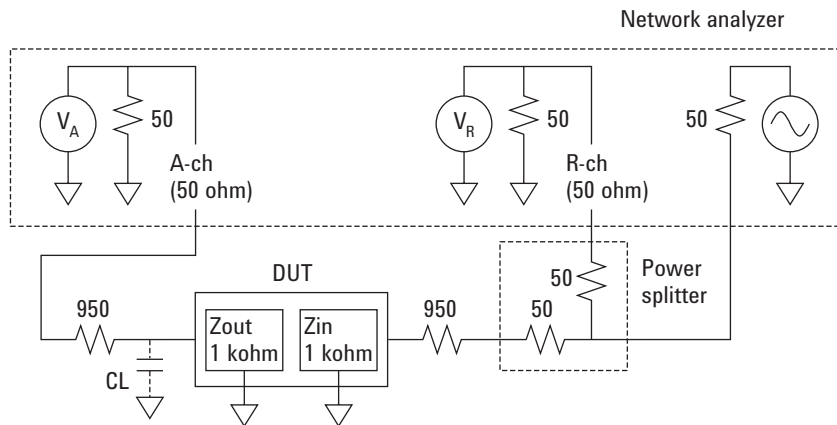
Calibration: Response thru cal. by connecting thru device in place of DUT

Figure 6. Configuration for measuring passive IF filters (2)

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Basic Measurement Configurations

In-circuit probing measurements

The high-impedance probing approaches shown in Figures 2 to 4 are applicable to in-circuit probing measurements, in which we measure the frequency response between two test points in the circuit under test. Figure 7 shows how to measure the frequency response of the block-2 with dual high-impedance probes.

By using the dual-step approach shown in Figure 8, the equivalent measurement can be achieved even when only one high-impedance probe is available. First we measure the response of the block-1 by contacting the A-ch probe to TP1 and save the measured data into the memory trace.

Next we measure the entire response of the block-1 plus block-2 by probing TP2 (measured data is stored into the data trace). Then we can obtain the frequency response of the block-2 by using the trace math function of the analyzer.

Another possible method to make the equivalent measurement with single probing is to perform the response through calibration by probing TP1 and then perform 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 very sensitive to the capacitance at TP1, the single probing methods may give a slightly different measurement result from that of the dual probing method, this is due to the input capacitance of the R-ch probe applied to TP1 when measuring TP2 with the A-ch probe in the dual probing method. To perform the single probing measurement under the completely same condition as the dual probing method for data correlation, just connect a capacitor whose value is equal to the probe input capacitance between TP1 and circuit GND when making the measurement of step-2.
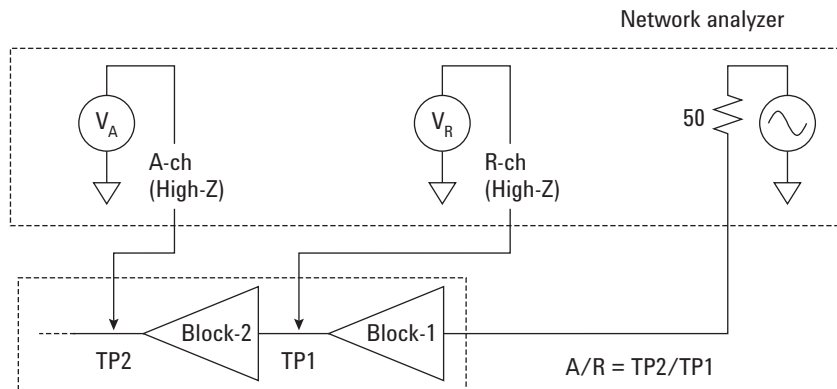

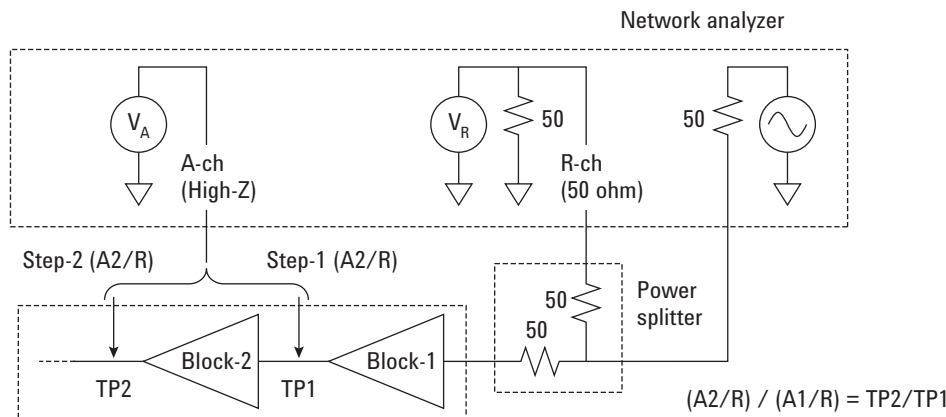Figure 7. In-circuit measurement with dual high-impedance probes

Figure 8. In-circuit measurement with single high-impedance probe

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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 in the 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 with 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 9, and this causes an unwanted large measurement response.

Figure 10 shows an example of measuring a 70 dB attenuator with the 4395A LF network analyzer under the conditions of source level = –10 dBm, start frequency = 1 kHz, and IFBW = 3 kHz. As you can see, an incorrect measurement response (the shape of receiver IF filter) 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.
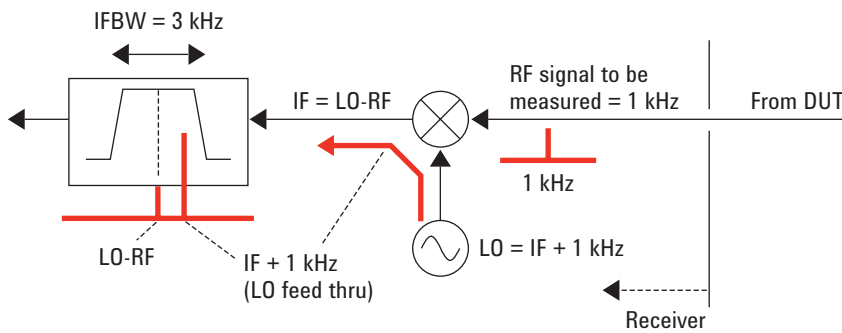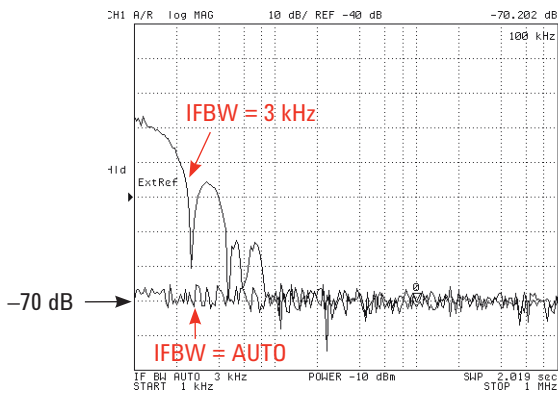

Figure 9. Measurement error caused by LO feed through

(Start = 1 kHz, IFBW = 3 kHz and AUTO)

Figure 10. Example of 70 dB attenuator measurement

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 (= Cin) is 50 pF, the input impedance (= $1/(2*pi*f*Cin)$) is 31.8 kohm at 100 kHz, which is still high impedance. If the frequency goes up to 10 MHz, the input impedance is 318 ohm, which is not high enough for many applications (generally, the probe input impedance should be at least 10 times greater than the impedance of the DUT). Also, the large input capacitance affects the measurements which are sensitive to capacitive loading, such as passive IF filters, resonant circuits, and some amplifier parameters which depend on the load capacitance condition (e.g. phase margin measurement). For those applications, it is necessary to use probing methods which provide small input capacitances.

If the network analyzer has a high-impedance input port, 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 11. If the measurement frequency range is lower than hundreds of kHz and if the 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 SNR. 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 hundreds of kHz. Also it is not suitable for the devices that are sensitive to capacitive loading.
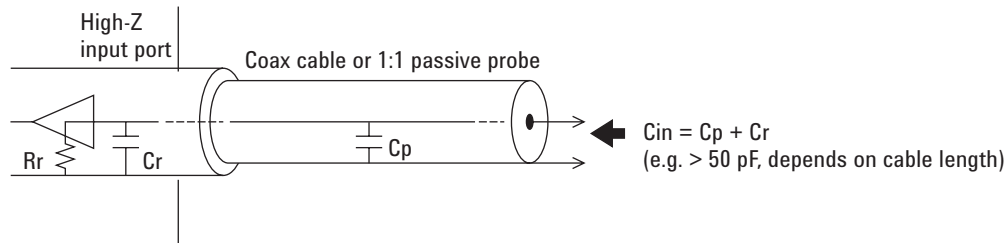

High-Z input port

Coax cable or 1:1 passive probe

Rr   Cr   Cp

Cin = Cp + Cr
(e.g. > 50 pF, depends on cable length)

Figure 11. Coaxial test cable or 1:1 passive probe

High-Impedance Probing Methods (continued)

The probe input capacitance can be reduced by using a 10:1 passive probe for oscilloscopes, which is designed for use with the high-impedance input port, as shown in Figure 12. 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 orthodox 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.

These problems can be solved by using an active probe. 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 13. For example, the input resistance/capacitance of the 41800A active probe (DC to 500 MHz) is 100 kohm/3 pF. Moreover, by adding the 10:1 adapter at the probe end, we can achieve 1 Mohm/1.5 pF, although the dynamic range is degraded by 20 dB in this case. If you need to measure up to very high frequency range over 10 MHz, or if the DUT is very sensitive to the load capacitance, it is recommended to use the active probe.
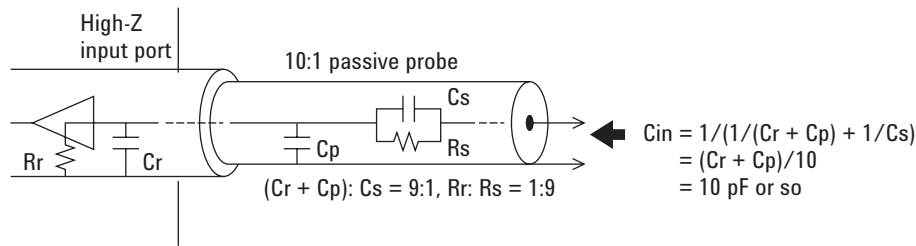

$$Cin = 1/(1/(Cr + Cp) + 1/Cs)$$
$$= (Cr + Cp)/10$$
$$= 10 \text{ pF or so}$$

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

Figure 12. 10:1 passive probe
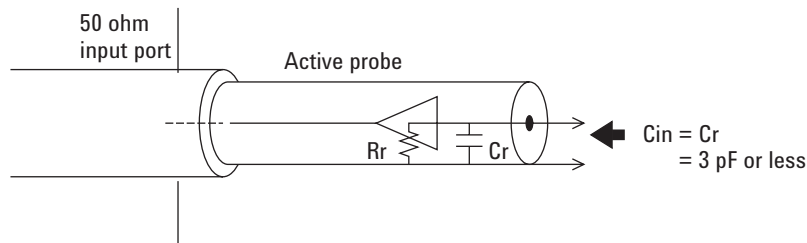

$$Cin = Cr$$
$$= 3 \text{ pF or less}$$

Figure 13. Active probe

Signal Separation for Ratio Measurement

To measure the transmission coefficient for 50 ohm devices such as passive filters in the system impedance Z0 = 50 ohm (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 ohm 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 by using an appropriate separation device. In general network analysis targeting linear devices, the most important requirement for the separation device is that it provides the 50 ohm source output impedance (source matching) when making the ratio measurement.

The most common separation device in the low frequency applications is a two-resistor type power splitter, which covers a very broad frequency range from DC to RF/microwave regions and provides an excellent source output impedance for the ratio measurement. An example is the 11667A/B (DC to 18 / 26.5 GHz, N-type/3.5 mm). The ratio measurement using the power splitter shown in Figure 14a is equivalent to making two measurements shown in Figure 14b 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 A-ch measurements will be precisely 50 ohm, which is generally an ideal source matching condition for 50 ohm 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 ohm system impedance because the splitter's physical output impedance seen from the DUT is not 50 ohm but 83.3 ohm.
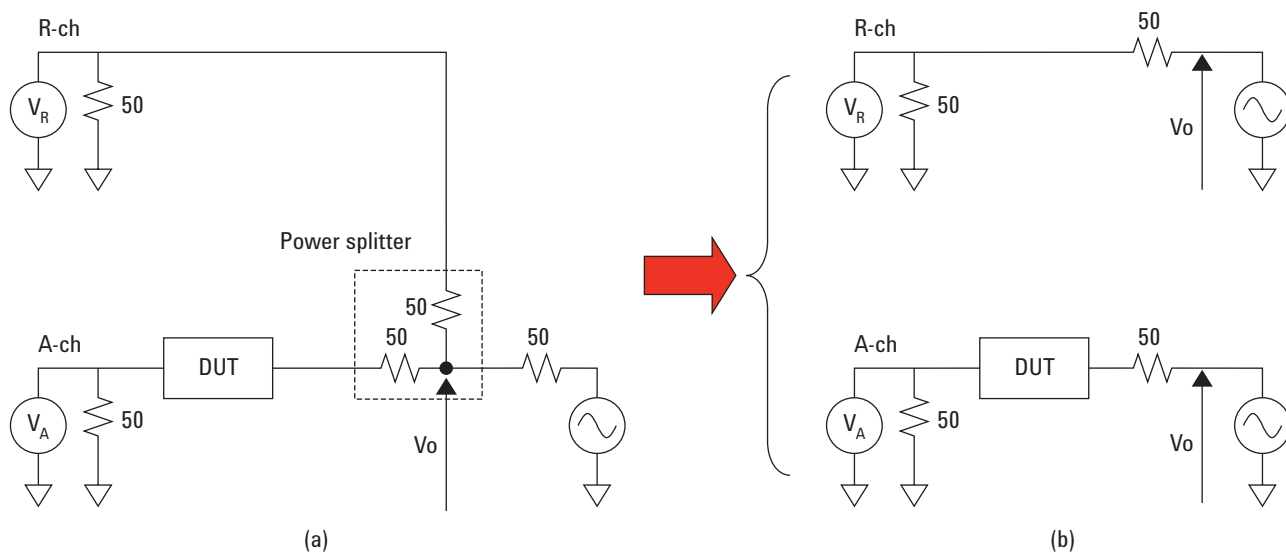

Figure 14. 50 ohm ratio measurement with power splitter

Signal Separation for Ratio Measurement (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 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 or so and the lower frequency coverage is 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 ohm. This enables you to perform the absolute voltage measurements in the 50 ohm environment, although this may not be so significant in the general low-frequency applications in contrast to RF applications.

If the analyzer has a high-impedance R-ch input port, the 50 ohm signal separation can be implemented by using an attenuator pad as shown

in Figure 16. The input impedance toward the DUT's input port is close to an ideal 50 ohm impedance due to the pad connected between the DUT and the branch point, the high-impedance R-ch receiver monitors the source output voltage applied to the 50 ohm system impedance. The attenuation of the pad should be more than 15 dB so that the R-ch measurement is sufficiently isolated from the DUT's actual input impedance. The measurement dynamic range is degraded by the attenuator.

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 ohm but 50/3 = 16.7 ohm 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 ohm.
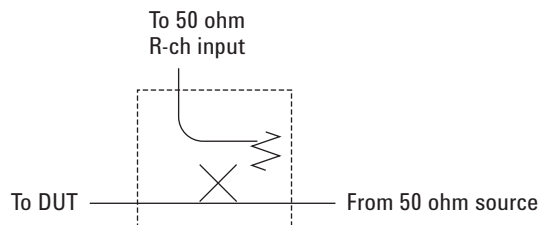

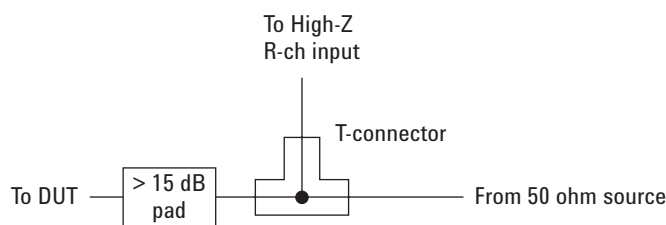Figure 15. Directional coupler/bridge
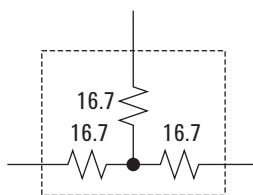

Figure 16. Method with high-impedance input

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

Measuring Small Signals at Low Frequencies

Measuring small signals in the 2-port transmission measurements of high attenuation devices or high gain devices with general network analyzers is likely to be affected by error factors related to the ground loop at low frequencies. Example applications are test parameters of OP-amps and other amplifier circuits, such as open-loop gain and CMRR. The most significant problem is the error caused by the shield resistance (braid resistance) of test cables, which is not negligible in the low frequency range below 100 kHz. Let's consider a simple model of a 2-port device that has an extremely high attenuation as shown in Figure 18. As the DUT's attenuation is very high, the voltage Vo is almost zero and the voltage V_A measured at the receiver should also be almost zero. However, since the source current flows into the ground loop as shown in the dotted lines, the voltage drop Vshield occurs across the shield resistance Rshield and the measured voltage V_A will be Vshield, which is higher than Vo that we want to measure. This degrades the measurement dynamic range.

In the actual applications where the error model around the DUT is more complicated, the source signal flowing into the ground loop may also cause other additional measurement errors. In addition, an external noise may enter into the ground loop and affect the measurement.

In the case of an analyzer with a floating receiver input, the ground loop is not formed between the source and receiver, which means these measurement errors will not occur.

With an analyzer whose receiver is not floating, there are several techniques to minimize these measurement errors. The most traditional 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 19. The magnetic cores increase the shield impedance and suppress 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. To suppress the ground loop current from the low frequency range, it is necessary to use a high-permeability core or to turn the cable many times around the core to increase the shield impedance as much as possible.
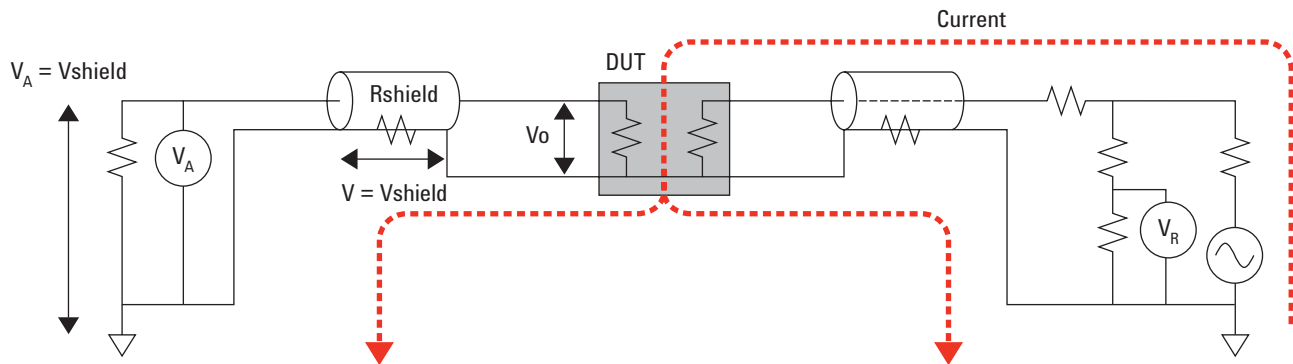

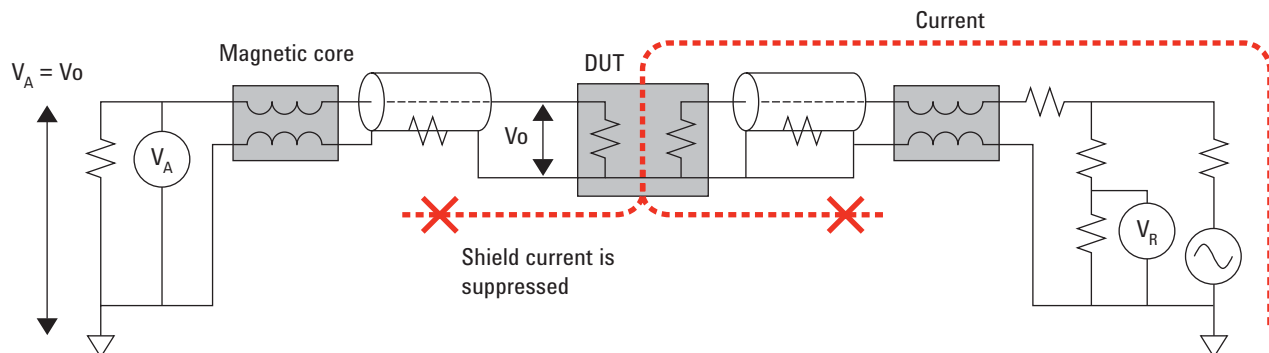*Figure 18. Measurement error due to cable shield resistance*

Figure 19. Suppressing shield current with magnetic cores

Measuring Small Signals at Low Frequencies (continued)

Another approach is to float the ground of the source or receiver to break the ground loop. We can implement this by using an isolation transformer or a differential probe. In the case of using the isolation transformer, theoretically it can be connected at the source side or receiver side. However in the applications targeted in this document such as amplifier measurements, it should be connected at the source side as shown in Figure 20. Connecting the transformer to the receiver side will affect the DUT's load condition, and also it seems that large residual responses are likely to occur if connecting off-the-shelf broadband

50 ohm isolation transformers to the receiver side.

Figure 21 shows an example of measuring a 100 dB attenuator with and without using the isolation transformer (North Hills Inc., 50 ohm Video Isolation Transformer, Model 0017CC, **www.northhills-sp.com**) at the source side. Without using the isolation transformer, the measured data in the low frequency range is higher than the correct value due to the effect of the cable shield resistance, and a strange dip appears around the center frequency area. These problems are solved by breaking the ground

loop with the isolation transformer as shown in the figure. A small residual positive peak appears around 200 kHz but it is small enough for most applications. When using the isolation transformer designed for the 50 ohm system impedance, the DUT's input impedance should not be that different from the 50 ohm range. If the DUT's input impedance is very high (e.g. 1 Mohm), a 50 ohm feed through should be connected between the transformer and the DUT.

Lastly, note that the DC power supply for the DUT should be floating as to not to create further ground loops.
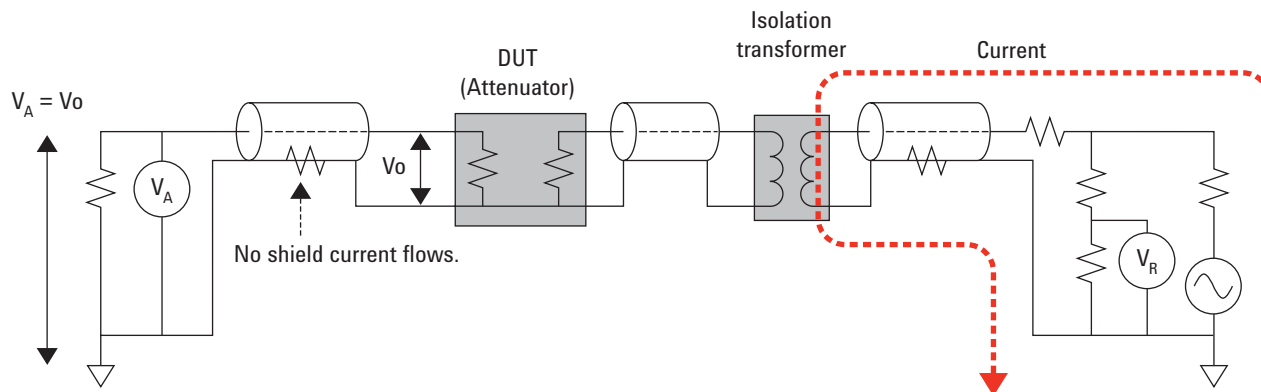
Figure 20. Solution with isolation transformer
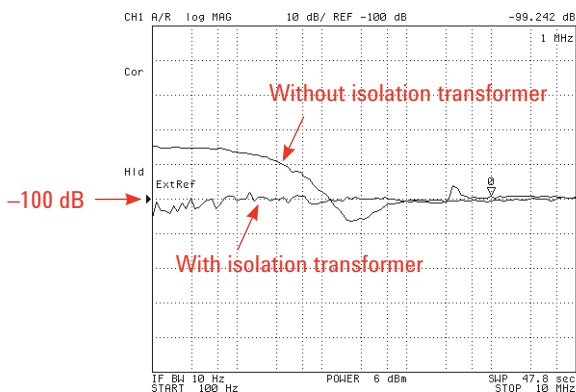
Figure 21. Effect of isolation transformer (DUT: 100 dB attenuator)

OP-Amp Measurement Example
Closed-loop gain

This section shows measurement examples of various frequency response characteristics of a high-speed operational amplifier.

Figure 22 shows a configuration example for measuring the closed-loop gain of a very simple inverting amplifier circuit with unity gain (Av = –1). The response through calibration should be performed by connecting the A-ch probe to the point TP1 so that the measured gain is referenced to the voltage at this point.

Figure 23 shows a measurement example of the closed-loop gain. In this measurement, the 41800A active probe is used as the high-impedance input to measure up to 100 MHz. The marker is put on the –1 dB cutoff frequency, which indicates the bandwidth of this amplifier circuit approximately 10 MHz.
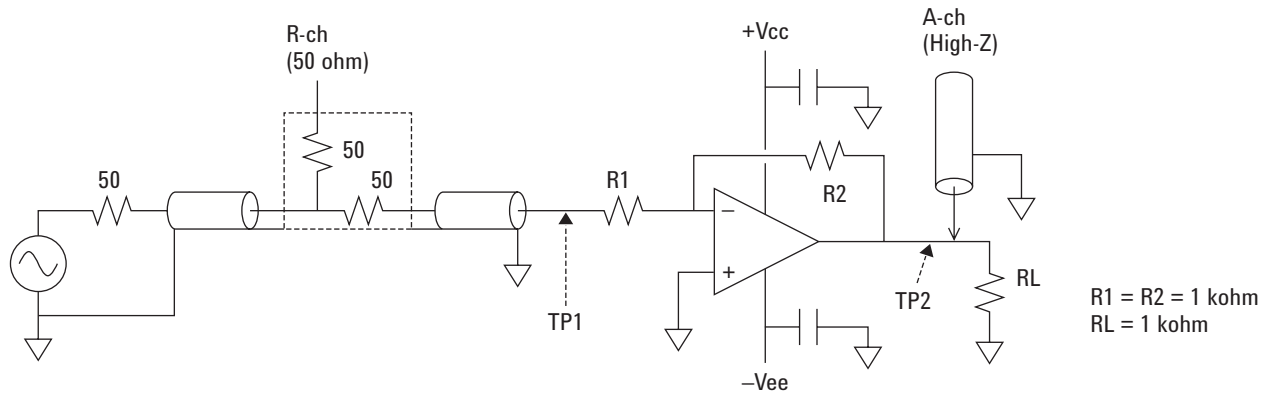
Figure 22. Configuration example of closed-loop gain measurement
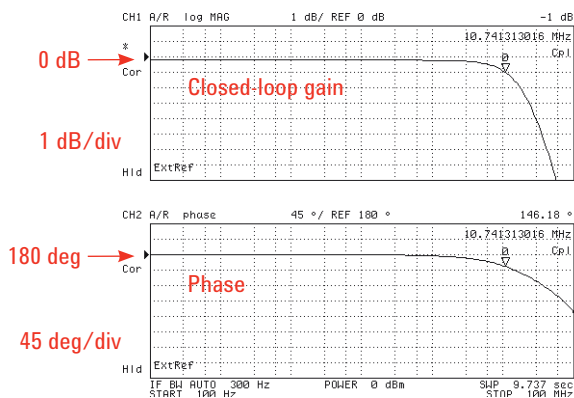
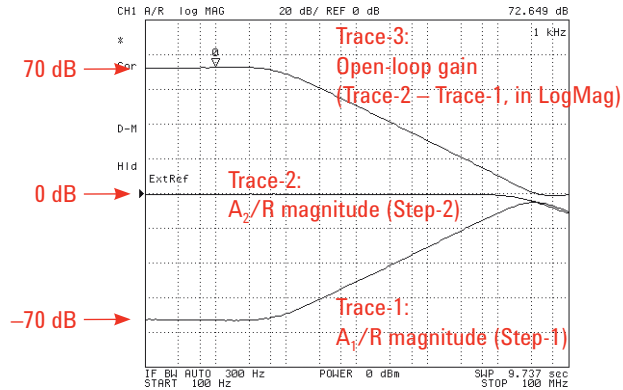*Figure 23. Closed-loop gain measurement example*

Network analyzer: 4395A

High-Z input: 41800A active probe (without divider)

Frequency = 100 Hz to 100 MHz

Source level = 0 dBm

NOP = 201

IFBW = AUTO (Upper limit = 300 Hz)

Receiver ATT setting = 10 dB

OP-Amp Measurement Example
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 shown in Figure 24. 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:

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

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

$$V_T/V_R = (-A - Zout/R2)/(1 - (Zout/R2))$$
$$= -A$$

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 we need to

avoid the receiver 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 narrowed if Av is high.

The open-loop gain measurement can be implemented by using either of the dual probing method or single probing method. Here we use the single probing method which is a little more complicated but a more reasonable solution especially in the case of using the active probe. Figure 25 is the configuration used in the measurement examples shown in Figures 26 through 28.
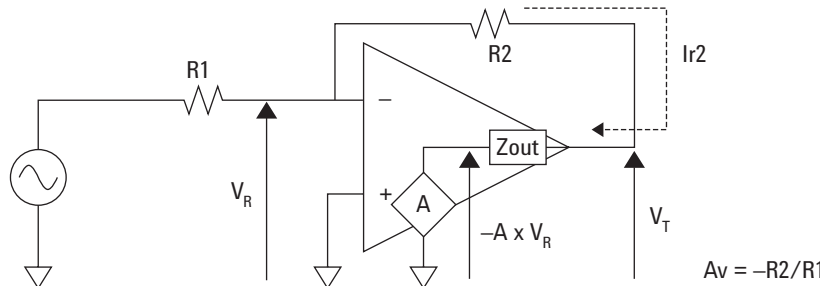
Figure 24. Configuration example of open-loop gain measurement
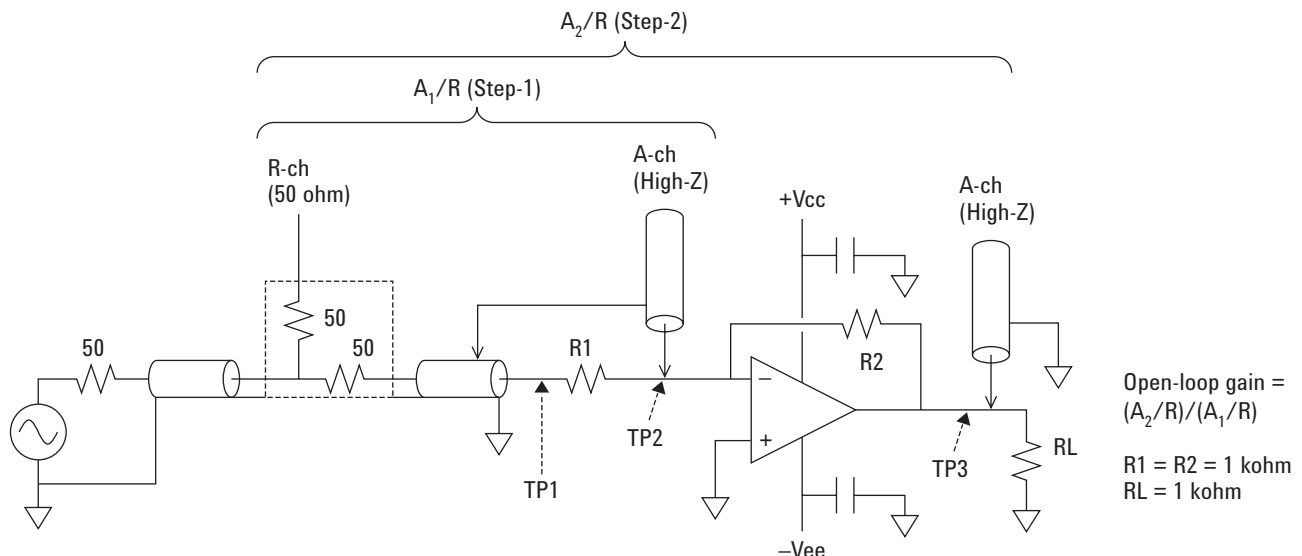
Figure 25. Configuration example of open-loop gain measurement

OP-Amp Measurement Example (continued)
Open-loop gain

Figure 26 shows a measurement example of the open-loop gain with the configuration of Figure 25. Trace-1 is the measured response by probing TP2, which indicates the ratio of the input voltage and the attenuated voltage V_R. Trace-2 is the measured response by probing TP3, which is the closed-loop gain Av and trace-3 is the open-loop gain calculated from these measurement results. The results are calculated by using the trace math function (Data/Memory in most analyzers including the ENA series, and Data-Memory in the case of the 4395A). The open-loop gain of this OP-amp is not so extremely high (about 70 dB), and it is accurately measured with the measurement circuit of $|Av| = 1$. To minimize the errors associated with the ground loop, a magnetic core is clamped to the cable of the 41800A active probe and the contacting point of the probe's ground lead is carefully chosen for the V_R measurement.

In this example, the ground lead is connected to the outer shield of the source-side cable.

Figure 27 shows the open-loop phase response measured with the same two-step measurement method. Figure 28 shows the loop gain and the phase margin derived from these measurements. By simply calculating the transfer function of the feedback path as $B = R1/(R1 + R2) = 1/2 = -6$ dB (assuming no phase shift), the loop gain $|A \times B|$ is derived by shifting down the open-loop gain trace by 6 dB with the trace math offset function. The marker is at the point of $|Av| = 1$ (0 dB). The phase margin of this amplifier circuit can be directly given by the marker on the phase trace (about 88 degrees), as we are looking at the round transfer function $-A \times B$ which includes the 180 degree inversion at the OP-amp input port.

Figure 26. Open-loop gain measurement example

Figure 27. Open-loop phase

Figure 28. Loop gain and phase margin

Network analyzer: 4395A

High-Z input: 41800A active probe (without divider)

Frequency = 100 Hz to 100 MHz

Source level = 0 dBm

NOP = 201

IFBW = AUTO (Upper limit = 300 Hz)
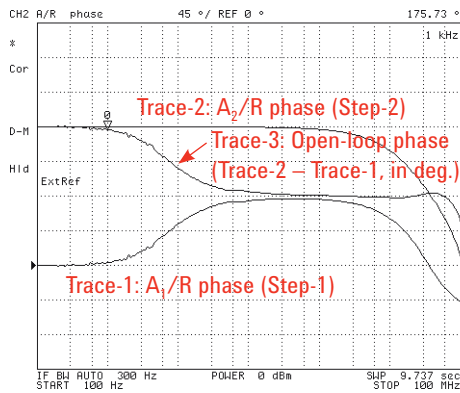
Receiver ATT setting = 10 dB

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OP-Amp Measurement Example
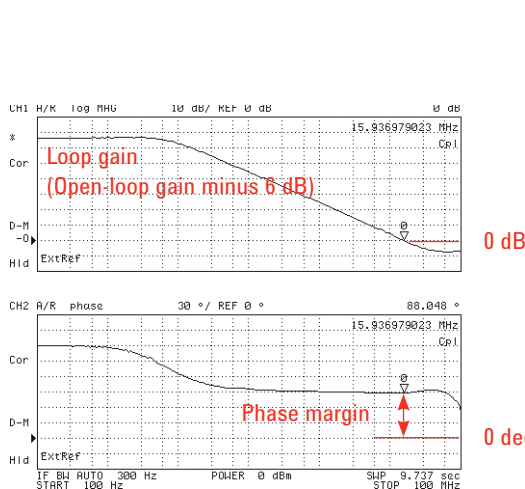## CMRR

Similarly to the open-loop gain, the CMRR (Common-Mode Rejection Ratio) of OP amps is generally difficult to measure, because we need to measure very small output voltages for common-mode inputs. The CMRR is defined as CMRR = Ad/Ac, where Ad is the differential-mode gain and Ac is the common-mode gain. Figure 29 shows a configuration example for measuring the CMRR. The differential gain Ad is measured by turning the switch SW1 to D-position. The common-mode gain Ac is measured by turning SW1 to C-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.

To avoid the errors associated with the ground loop, the 50 ohm isolation transformer is connected at the source side. The output of the transformer is terminated with the 50 ohm feed through. The response through calibration should be performed by probing TP1.

Figure 30 shows a CMRR measurement example. Trace-1 is the common-mode gain Ac, and trace-2 is the differential gain Ad (= 20 dB). The common-mode gain Ac of about −80 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 90 dB. In the lower frequency range, the CMRR is about 100 dB.
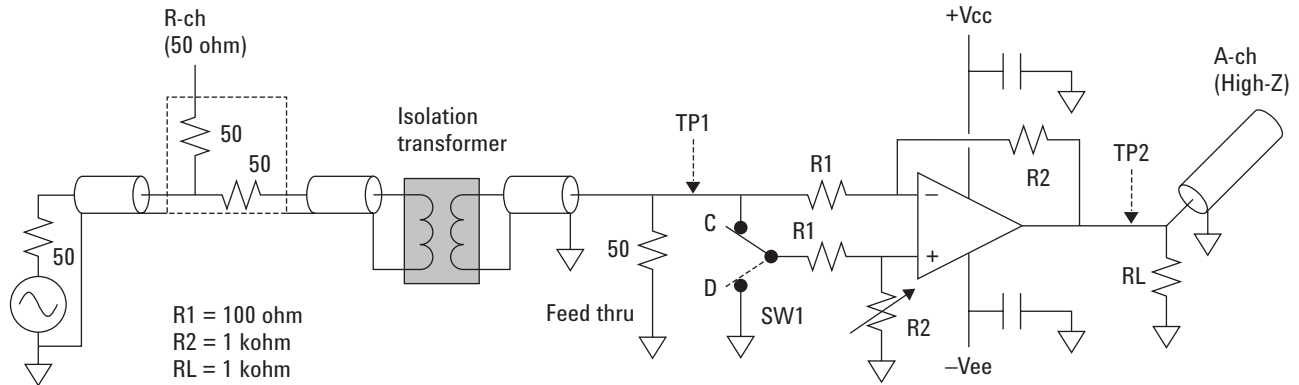
Figure 29. Configuration example of CMRR measurement
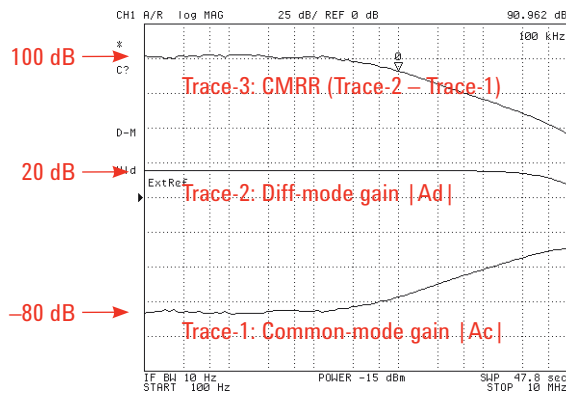
Figure 30. CMRR measurement example

Network analyzer: 4395A

High-Z input: 41800A active probe (without divider)

Isolation transformer: North Hills 0017CC (50 ohm, 10 Hz to 5 MHz)

Frequency = 100 Hz to 10 MHz

Source level = 0 dBm (for Ac measurement)
             −15 dBm (for Ad measurement)

NOP = 101

IFBW = 10 Hz

Receiver ATT setting = 10 dB

The balance of R1 and R2 is not fully optimized in this measurement example.

OP-Amp Measurement Example
PSRR

The PSRR (Power Supply Rejection Ratio) of OP-amps is another difficult parameter to measure as it requires small signal 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 31 shows a configuration example for measuring the PSRR (positive PSRR). 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 DC bias is provided with a very simple bias-tee circuit that consists of a capacitor and a resistor in this example. (If the analyzer has a built-in DC bias function, the external bias circuit is not necessary.) As we need to provide a DC bias voltage through the source port, the source isolation transformer is not used. Instead, we need to carefully measure the small output voltage by minimizing the effects of the ground loop.

Figure 32 shows a PSRR measurement example with this

configuration. The marker indicates that the PSRR at 100 kHz is about −60 dB. In the lower frequency range, the PSRR is less than −80 dB. To measure this very high attenuation, the analyzer's IFBW is set to 10 Hz, and the probe is connected to the DUT via a very short SMA coaxial cable by attaching the coaxial adapter to the probe to avoid unwanted coupling that may occur around the probe end.

To measure the OP-amp's PSRR higher than the example described (e.g. more than 100 dB), the measurement would be increasingly difficult and a circuit with a higher Av should be used.
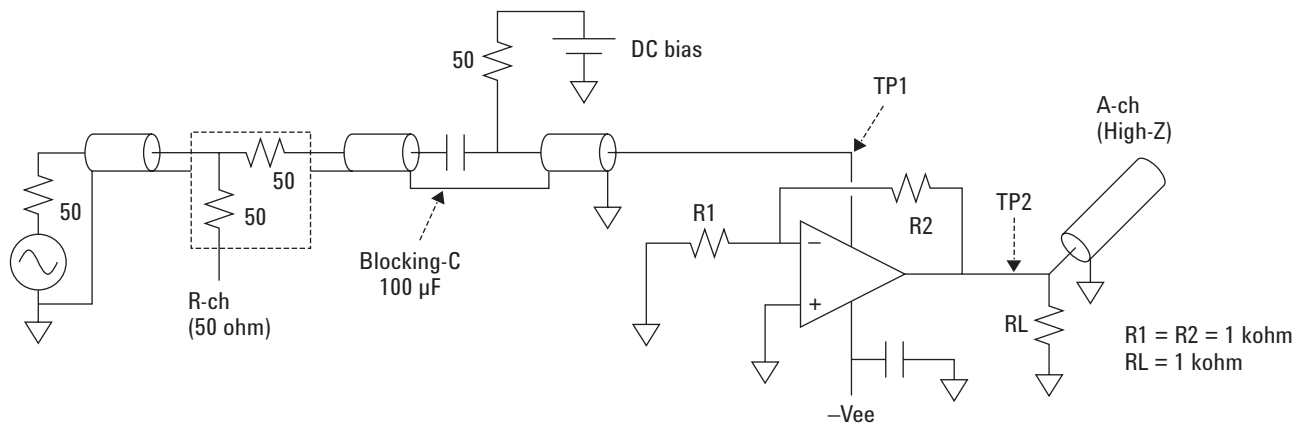
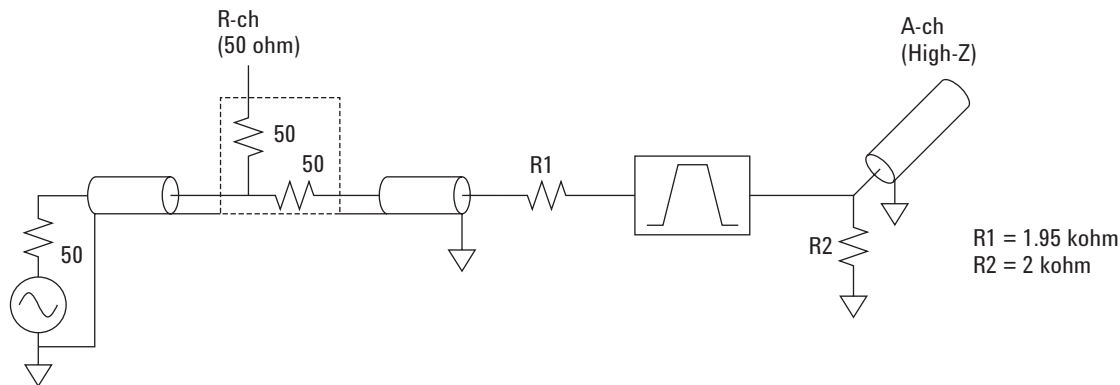*Figure 31. Configuration example of PSRR measurement*
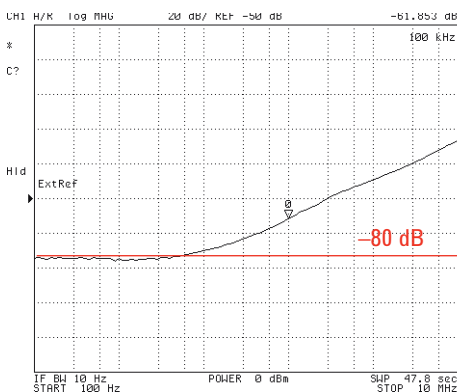
Network analyzer: 4395A

High-Z input: 41800A active probe (without divider)

Frequency = 100 Hz to 10 MHz

Source level = 0 dBm

NOP = 101

IFBW = 10 Hz

Receiver ATT setting = 10 dB

Figure 32. PSRR measurement example

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OP-Amp Measurement Example
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 milliohms at low frequencies and up to 100 ohms at high frequencies. To fully cover this impedance range, the reflection measurement method is the proper solution.

Figure 33 shows a configuration example for measuring the closed-loop output impedance of OP-amps. The open/short/load 3-term calibration (1-port full calibration) must be performed at TP1.

Figure 34 is a measurement example of the closed-loop output impedance. The measured trace shows the frequency response of the impedance in the logarithmic scale.
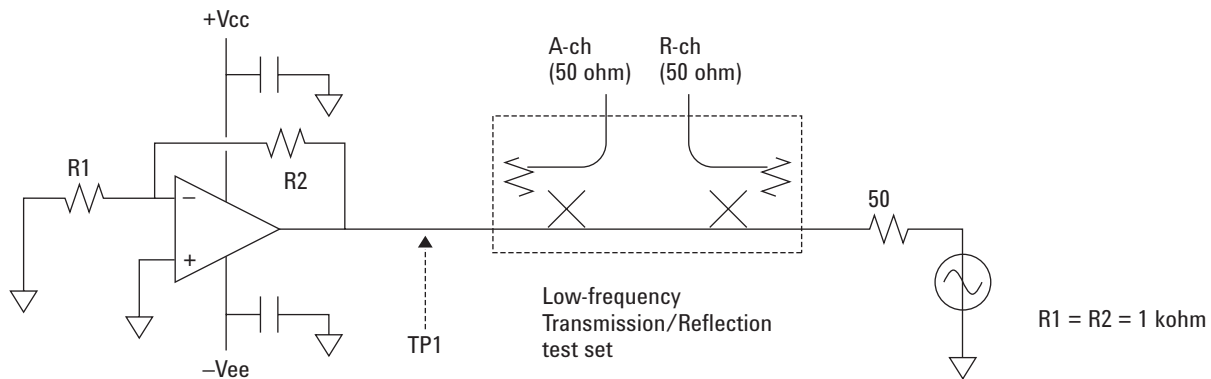
Figure 33. Configuration example of output impedance measurement

Figure 34. Output impedance measurement example
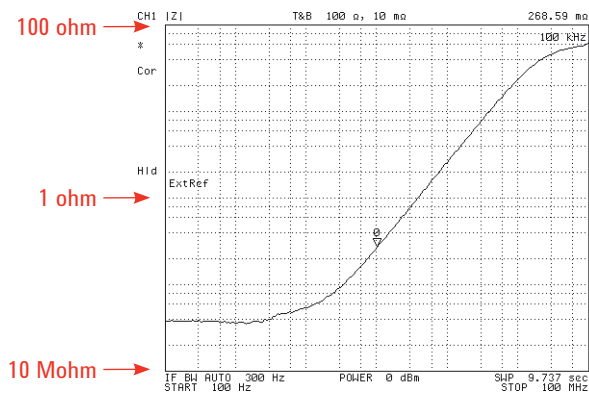
Network analyzer: 4395A with Option 010
(Used in the impedance analyzer mode)

T/R test set: 87512A

Frequency = 100 Hz to 100 MHz

Source level = 0 dBm

NOP = 201

IFBW = AUTO (Limit 300 Hz)

Receiver ATT setting = 10 dB

Ceramic IF Filter Measurement Example
Transmission characteristics

Figure 35 shows a configuration example for measuring a ceramic IF filter with 2 kohm input and output impedance by using the configuration shown in Figure 5. Since the DUT is sensitive to capacitive loading, it is necessary to use a high-impedance probing method that has small input capacitance, such as a 10:1 passive probe or an active probe. An active probe is a better choice as it has smaller input capacitance and

does not degrade the measurement dynamic range. The response through calibration should be performed by connecting a shorting device in place of the DUT.

Figure 36 shows measurement examples of a 455 kHz ceramic filter with 2 kohm input and output impedance. The high-impedance probe used in this example

is the 41800A active probe (Rin//Cin = 100 kohm//3 pF). The probe is connected to the fixture via a very short SMA coaxial cable not to increase the input capacitance. Figure 36 shows the magnitude and group delay response in the narrow span. Figure 37 shows the magnitude response in the wider span, which exhibits spurious responses of the DUT.

Figure 35. Configuration example of ceramic IF filter measurement
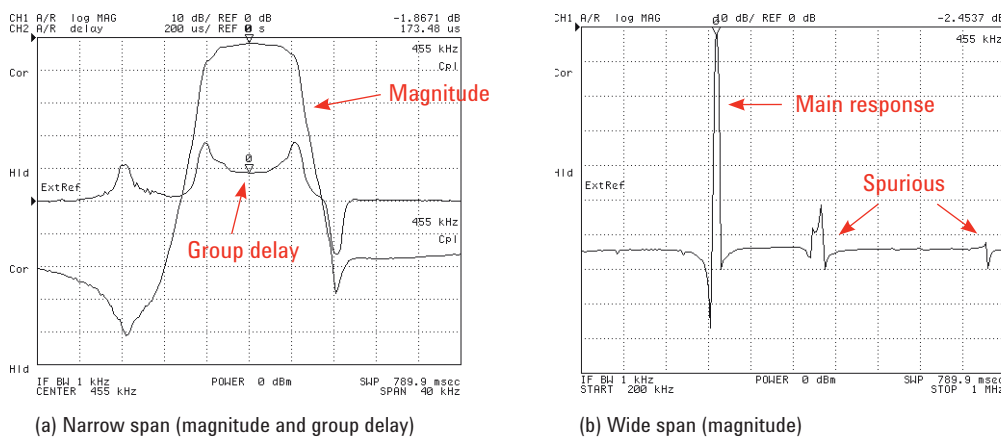
(a) Narrow span (magnitude and group delay)   (b) Wide span (magnitude)

Figure 36. Ceramic IF filter transmission measurement example

Network analyzer: 4395A

High-Z input: 41800A active probe (without divider)

Ceramic IF Filter Measurement Example
Effect of capacitive loading

Figure 37 shows pass band responses of the 455 kHz ceramic IF filter measured by connecting the active probe via a very short SMA coaxial cable and via a 60 cm BNC cable. The measurement result indicates that the measurement error of more than 0.5 dB occurs when using the 60 cm cable because the filter characteristic is affected by the large cable capacitance of about 60 pF. This measurement error cannot be eliminated with the response through calibration. So it is important to minimize the capacitance of high-impedance probing.

In the case of DUTs that are more sensitive to capacitive loading than this example indicates, even the short coaxial cable between the DUT and the probe would be undesired. In this case, the DUT should be directly probed with the active probe (if necessary, with an additional 10:1 divider to further reduce the probe capacitance).
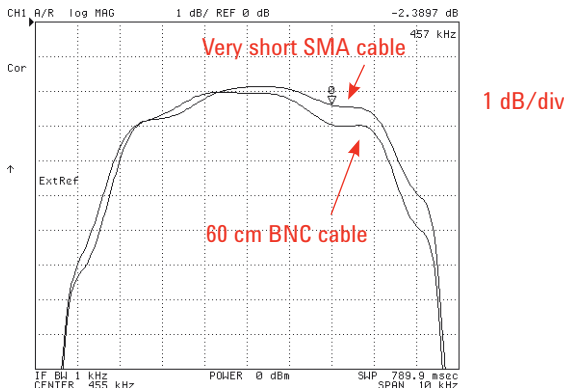
Figure 37. Effect of capacitive loading

References

[1]   Robert A. Witte, "Spectrum and Network Measurements", 1993

[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