

Agilent Technologies

Evaluating DC-DC Converters with LF Network Analyzer

Application Note



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► Introduction

Switch-mode DC-DC converters are widely used in electronic equipment in any field. One of the significant trends related to the DC-DC converters is that their load devices such as microprocessors, FPGAs, DSPs, and DDR memories are continuously evolving toward faster operating speeds and lower operating voltages. Also, various voltage levels are required by these high-speed digital ICs (3.3 V, 2.5V, 1.8 V, 1.5 V, 1.2 V, etc.). In accordance with this trend, the distributed power architecture is becoming common, where the DC-DC converters are mounted at the very close points to the load devices (POL converters) to improve the power integrity, especially in the IT devices such as servers and network infrastructure equipment. The DC-DC converters are key devices that determine the power integrity performance.

To make the DC-DC converters quickly respond to the load variations of the high-speed digital ICs, it is more important than ever to optimize the balance of the response speed and the stability of the feedback loop circuits. To minimize the voltage drops due to the large load current variations

(IR drop) and ensure the variations of the power plane voltages are within the allowable margins, it is necessary to confirm that the output impedance of the DC-DC converters is suppressed to a very small impedance of milliohm order. Evaluating these frequency domain characteristics is getting more important not only for the engineers of power supply manufacturers and power supply IC manufacturers, but also for the engineers of electronic equipment manufacturers who purchase the DC-DC converter modules to be integrated in their systems, or purchase the DC-DC converter ICs to build their own power supply circuits.

This application note describes the measurement methods for evaluating the frequency domain characteristics of the DC-DC converters by using the LF network analyzer. The topics discussed include the measurement methods for the feedback loop characteristics and the output impedance, which are the most important frequency domain characteristics. Also, we discuss measurement methods for evaluating the impedance characteristics of passive components used in the DC-DC converters.

► DC-DC Converter Basic Theory of Operation

First of all, we quickly summarize the basic theory of operation of the DC-DC converter. The example shown here is a simple non-isolated single-phase buck converter with a voltage-mode control.

The block diagrams and the timing charts shown in Figure 1 explain the basic operation of the buck DC-DC converter. The input DC voltage V_{in} is converted to the pulsed voltage with the switch (MOSFET), its on/off condition is controlled by the feedback loop circuit, and the pulsed voltage is converted to the output DC voltage V_{out} with the charging and discharging operations of the output LC filter. When the switch is turned on, the current I_{on} flows through the inductor L and the power is delivered to the output capacitor C_{out} and the load, then V_{out} is increased.

If V_{out} reaches a certain voltage level, the switch is turned off and the energy that was charged to L by the current I_{on} generates the current I_{off} and delivers the power to the load, together with the energy that was charged to C_{out} , then V_{out} is decreased. If V_{out} reaches a certain level, the switch is turned on and V_{out} is increased again. The output voltage level is determined by the pulse duty ratio. The longer the period T_{on} , the higher the output voltage is. The shorter the period T_{on} , the lower the output voltage is. With a current higher than a certain level continuously flows through the inductor L , the averaged output voltage is calculated as $V_{out} = T_{on} / (T_{on} + T_{off}) \times V_{in}$. By repeating this on/off operation while monitoring the output voltage and adjusting the pulse duty ratio, the regulated output DC voltage is obtained regardless of the load variations.

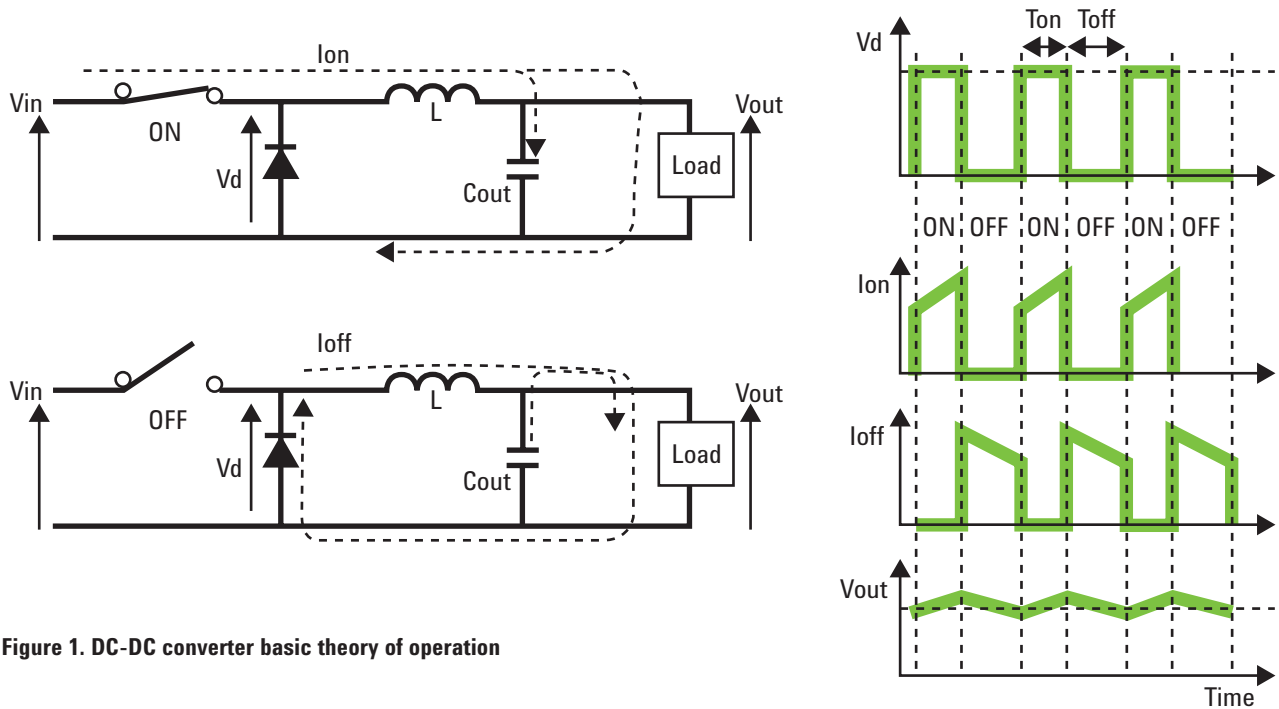


Figure 1. DC-DC converter basic theory of operation

Figure 2 shows an example of a detailed block diagram. The output voltage divided by R1 and R2 is fed back to the error amplifier, and the error amplifier compares the feedback voltage with the stable reference voltage Vref to provide the output voltage proportional to the difference between them. The Pulse Width Modulator (PWM) provides the pulse with the duty ratio determined by the error amplifier's output voltage, and the pulse turns the MOSFET on and off. When the feedback voltage is lower than Vref, the feedback system extends the period Ton to increase the output

voltage. When the feedback voltage is higher than Vref, the feedback system shortens the period Ton to decrease the output voltage. Thus the regulated DC output voltage is obtained.

C1, C2, C3, R3, and R4 are the components which adjust the gain and phase delay of the error amplifier to improve the feedback loop stability (feedback compensation), together with R1 and R2.

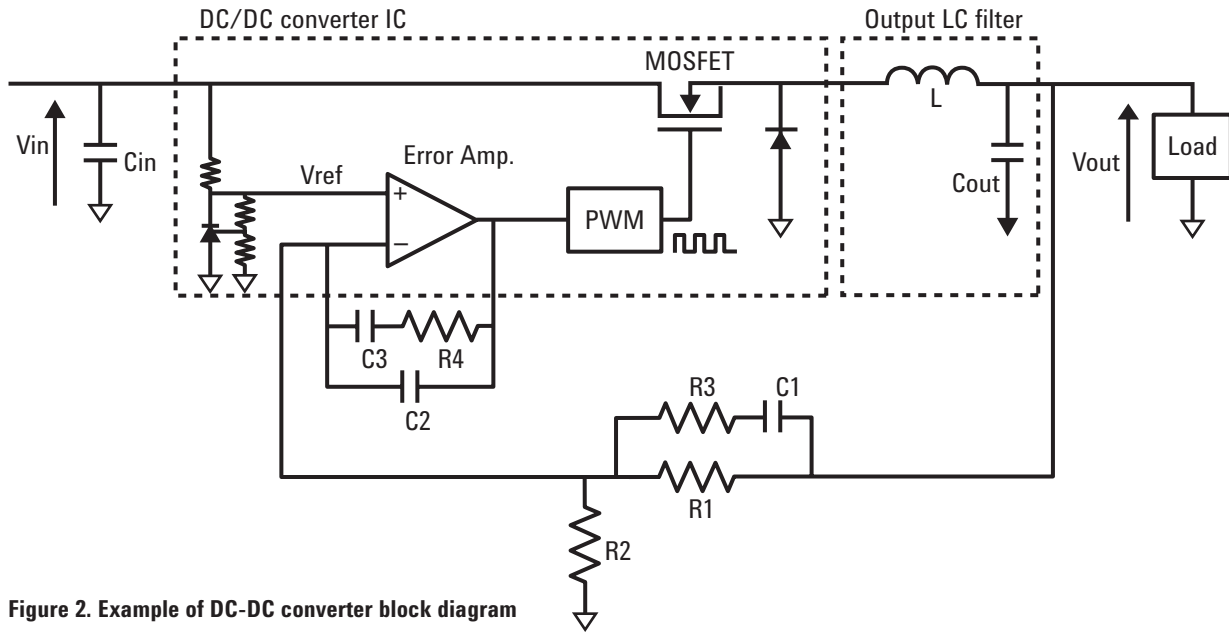


Figure 2. Example of DC-DC converter block diagram

► Evaluating Feedback Loop Characteristics

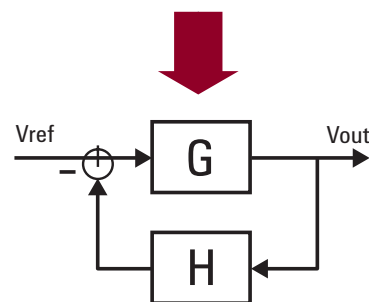
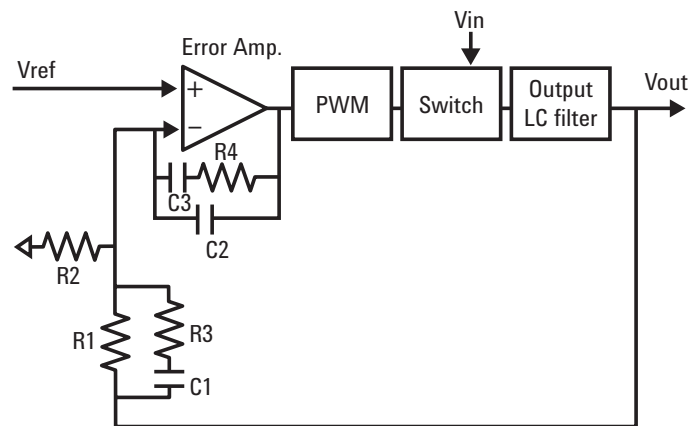
This section describes how to evaluate the feedback loop characteristics with the LF network analyzer.

Before starting the discussion on the measurement methods, we quickly summarize the basics of the feedback loop control.

Loop gain

As shown in Figure 3, the DC-DC converter can be regarded as a negative feedback control system which has the input signal V_{ref} and the output signal V_{out} . $|G|$ is called the open loop gain, $|V_{out}/V_{ref}| = |G/(1+GH)|$ is called closed loop gain, and $|GH|$ is called loop gain. Here it should be noted that the round transfer function is $GH \times (-1) = -GH$ because it includes the inversion at the error amplifier. The transfer function G corresponds to the total transfer function from the error amplifier to the output LC filter, and transfer function H corresponds to the resistive divider circuit that consists of $R1$ and $R2$. The resistors $R1$ and $R2$ also determines the gain and phase delay of the error amplifier, together with $R3$, $C1$, $C2$, $C3$, and $R4$.

This negative feedback control system regulates the varying output voltage V_{out} closer to V_{ref}/H . The larger the loop gain $|GH|$, the better the voltage regulation. As the frequency of the variation increases, the loop gain decreases and as the loop gain becomes lower than 1, the regulation will not work. The frequency where the loop gain $|GH|$ is equal to 1 ($= 0$ dB) is called the crossover frequency, which indicates the bandwidth of the loop (Figure 4). The higher crossover frequency enables the feedback loop to regulate the voltage variation of higher frequencies and the response speed to the load variations becomes faster.



$$\frac{V_{out}}{V_{ref}} = \frac{G}{1+GH}$$

Figure 3. Negative feedback loop control system

Phase margin and gain margin

The phase delay occurs at high frequencies. Now let's look at the phase delay of the round transfer function $-GH$. In the low frequency range near DC, only the 180 degree phase delay of the error amplifier occurs. As the frequency progresses higher, the phase delay of the error amplifier gets larger and additional delays occur at other locations in the loop. As shown in Figure 5, large phase delay occurs around the resonant frequency $f_c = 1/(2\pi\sqrt{L \cdot C})$ of the output LC filter. Especially in the case of the low-ESR MLCCs that are widely used for reducing the output ripple noise, the LC filter's phase response becomes close to the ideal LC filter's due to the extremely low ESR, and the phase delay around the resonant frequency becomes very large, close to 180 degrees. If the total phase delay of the feedback loop is close to 360 degrees, the feedback loop will appear as positive feedback, rather than the negative feedback. And, if the loop gain $|GH|$ is still more than 1, an unstable control loop is likely to cause oscillation due to the variation of the components used in the loop circuits and other environmental changes such as temperature. To avoid such a problem, the feedback compensation for stabilizing the loop is implemented, adjusting the gain and phase of the error amplifier around the resonant frequency of the LC filter by adding the feedback compensation components (R3, R4, C1, C2, and C3 in Figure 3).

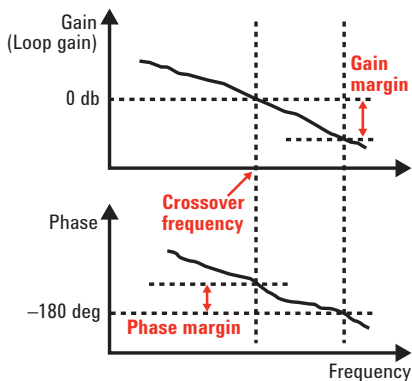


Figure 4. Gain-phase characteristics of GH (Bode plot)

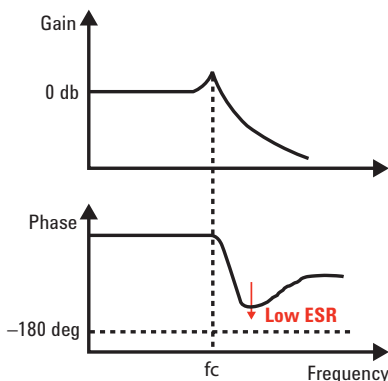


Figure 5. Gain-phase characteristics of output LC filter

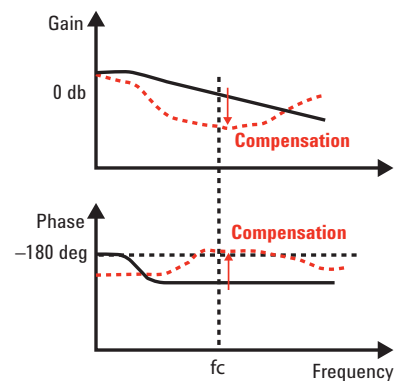


Figure 6. Gain-phase characteristics of error amplifier block and feedback compensation

As shown in Figure 4, the difference between the phase angle of $-GH$ and -360 degrees at the crossover frequency where the loop gain $|GH|=1$ (that is, the difference between the phase angle of GH and -180 degrees) is called the phase margin. The phase margin is an important parameter that shows the loop stability. The larger the phase margin, the more stable the feedback loop. The feedback loop must have an enough phase margin to ensure the stable operation under any load condition of the actual application.

However, if the crossover frequency becomes low as a result of implementing the feedback compensation focused on the phase margin, the response speed to the load variations will be slower. Therefore, it is necessary to design the feedback compensation circuits so that the stability and the response speed are optimized for the requirements of targeted applications. This optimization is becoming more important than ever especially in the DC-DC converters for the high-speed digital circuits. To optimize these parameters, validating the actual loop characteristics with the LF network analyzer is crucial, because the accuracy of the loop characteristics is unknown with simulation only.

Similarly, the difference between the gain of $-GH$ and 0 dB at the frequency where the phase is 0 degree is called the gain margin, which is also the key parameter for evaluating the loop stability.

Loop gain measurement method with LF network analyzer

The feedback loop circuit in its operating condition can be measured with the LF network analyzer without breaking the loop by injecting the analyzer's source signal into the feedback loop via an injection circuit. The analyzer measures the ratio of the AC voltages at both ends of the injection circuit with the receivers R and A that have high impedance inputs. The signal should be injected at the point where the input impedance Z_{in} when looking into the direction of the signal transmission is high and the output impedance Z_{out} when looking into the opposite direction is low. In the case of the DC-DC converter, the test signal is generally injected at the point just before the divider circuit on the feedback path by using the floating injection circuit that consists of a transformer and a resistor, as shown in Figure 7. By injecting the test signal at the point where $Z_{in} \gg R \gg Z_{out}$ is satisfied and using the resistor R that satisfies $Z_{in} \gg R \gg Z_{out}$, we can measure the round transfer function $-GH$ with the ratio measurement A/R without disturbing the original loop characteristics.

The injected signal level should not be very high to prevent the feedback loop circuit from getting into the nonlinear region. The probing should be done with the high input impedance as not to affect the operation of the feedback loop circuit. The input impedance of the probes must be sufficiently higher than the DUT's input impedance Z_{in} .

As for the measurement frequency range, it seems common that the measurement is started from the low frequencies like 10 Hz or 100 Hz. But in general, the important frequency range for evaluating the loop characteristics of the DC-DC converters is mainly several kHz to several hundreds of kHz where the LC filter's resonant frequency and the loop crossover frequency exist. Therefore, the measurement at the low frequency range does not have to be so strict.

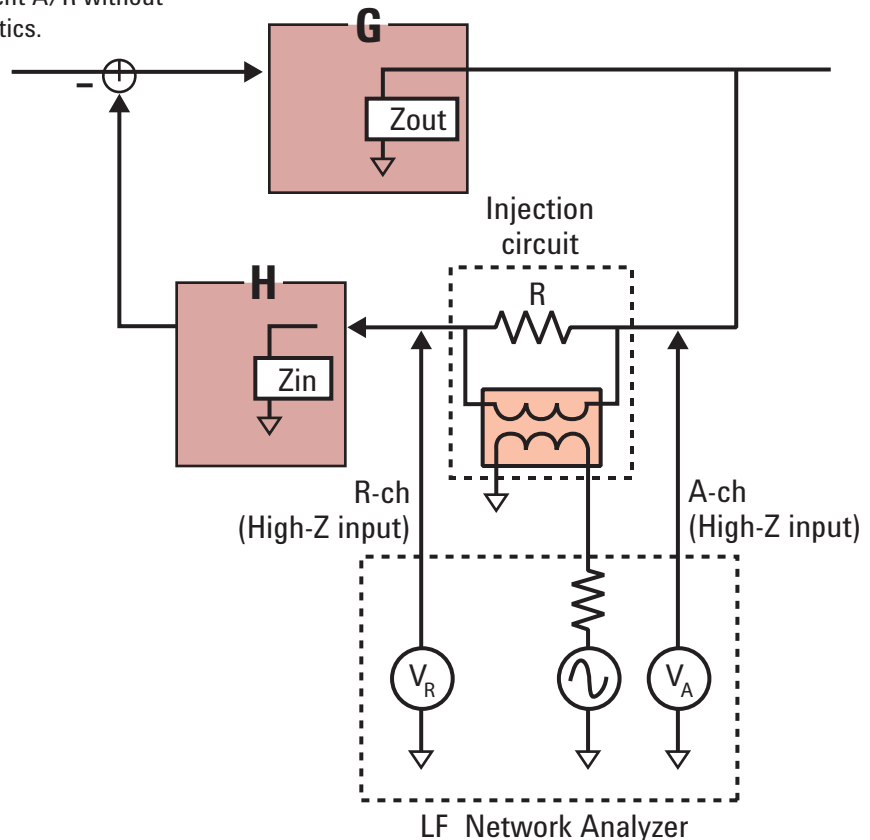
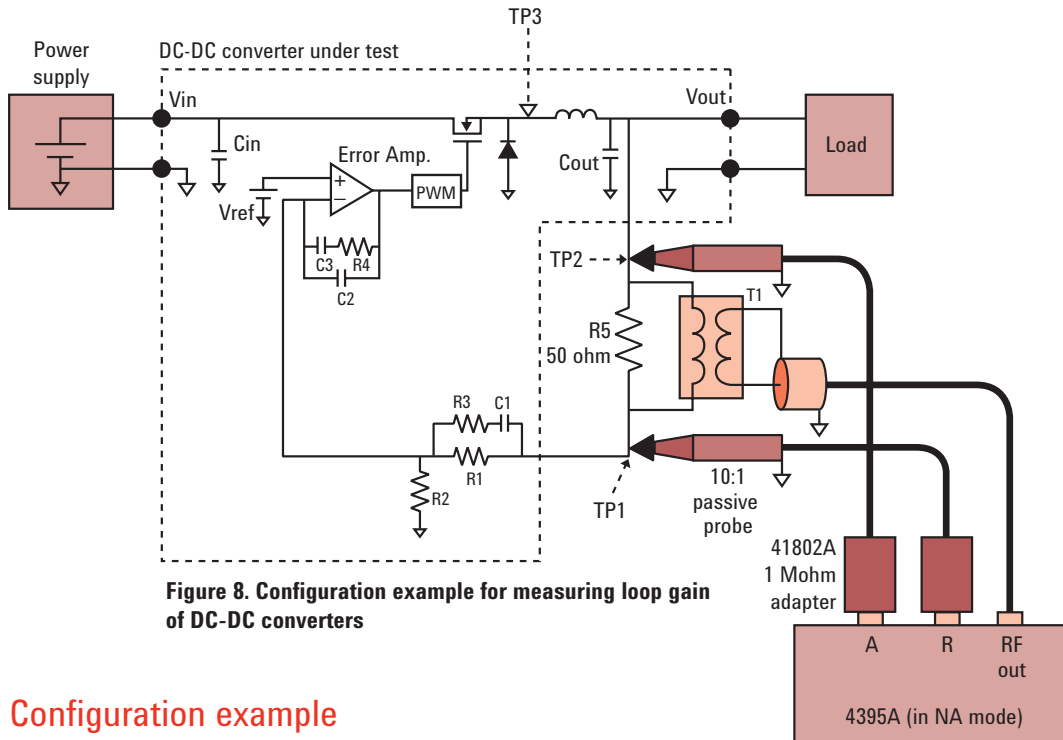


Figure 7. Loop gain measurement method for negative feedback control systems



Configuration example of loop gain measurement

Figure 8 shows a configuration example for measuring the loop gain. In this example the 4395A is used as the network analyzer. The 41802A high-impedance adapter (1 Mohm//12 pF) is used because the 4395A's receivers are 50 ohm inputs (this adapter is not necessary if your network analyzer has built-in high-impedance inputs.) The 10:1 high-impedance passive probes for oscilloscopes are used for connecting the high-impedance receivers to the DUT so that probing can be done with high impedance even at a frequency range of over 100 kHz.

If the 20 dB loss of the 10:1 passive probe is a concern (for example, if you want to precisely measure the loop gain at a very low frequency range below 100 Hz where the measurement SNR tends to get worse, or if the injection signal level must be set to a very low level due to the DUT's restrictions), you can improve the measurement SNR by connecting the analyzer's high impedance receivers to the DUT with test cables, such as BNC-alligator clip test cables. In this case, the length of the test cables should be as short as possible, otherwise the test cable capacitance is added to the input capacitance of the analyzer's high impedance receivers and this will reduce the probing input impedance at high frequencies. Also, it must be noted that this method is not applicable to the converters whose DC output voltages are higher than the maximum allowable DC input level of the high-impedance receivers being used (in the case of the 41802A, it is 50 Vdc).

The transformer T1 and the resistor R5 configure the signal injection circuit. The resistance value of R5 should be sufficiently smaller than Z_{in} , which is generally several kohms or several tens of kohms. Also, if the resistance of R5 is too small, the injected test signal will be excessively attenuated. In general, the resistance around 50 ohm is appropriate as R5. You can use a normal low-power resistor as R5 because a large current does not flow through the feedback path where R5 is placed.

As for the load of the converter, DC electronic loads or high power resistors are generally used.

To calibrate the measurement system, perform the response through calibration to eliminate the differences in the amplitude and the phase between two probes by contacting them to the point TP1. As described later, the power measurement values at the receiver R tend to be very small in the low frequency range where the loop gain is high. If this is of concern, perform the response through calibration in the condition where the signal of appropriate level appears at TP1. For example, by turning off the converter (turning off the input voltage V_{in}) to stop the feedback loop operation only when performing the calibration.

Measurement example of feedback loop characteristics

Figure 9 shows a loop gain measurement example of a 5 V-to-3.3 V buck converter in 3 A load condition by using the measurement configuration shown in Figure 8. The measurement frequency range is 100 Hz to 1 MHz, the source level is -10 dBm. The IFBW is set to AUTO mode, which automatically selects narrow IFBW at low frequencies and wider IFBW at higher frequencies. The upper graph shows the loop gain, and the lower graph shows the phase response. The marker is placed on the crossover frequency (about 30 kHz). In this measurement example, the analyzer measures the round transfer function $-GH$. The phase measurement value at the crossover frequency (about 95 degrees) indicates the phase margin. The converter measured in this example has enough phase margin and it seems possible to improve the response speed to the load variations by further tuning the feedback compensation circuits to make the crossover frequency higher.

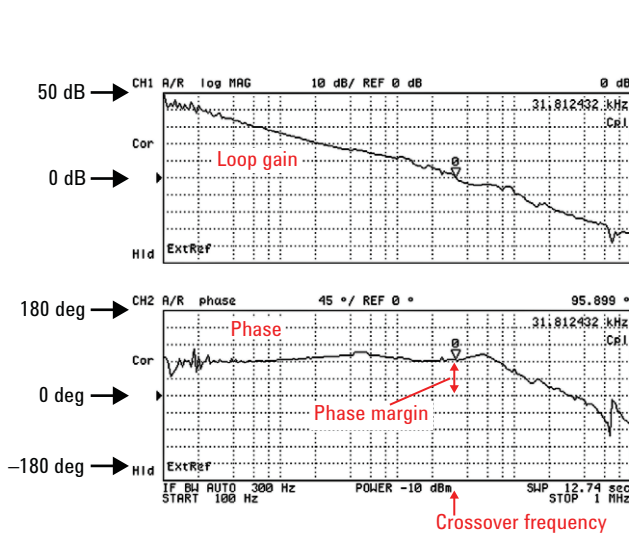


Figure 9. Loop gain measurement example of DC-DC converter

In addition, Figure 10 shows the measurement result of the transfer function of the LC filter block by changing the probing points of the receivers R and A to TP3 and TP2, and the measurement result of the transfer function from the error amplifier to the PWM output by changing the probing points of the receivers R and A to TP1 and TP2. (The load condition is different from the measurement shown in Figure 9.) The signal injection point is unchanged from that of the loop gain measurement. The combination of these two transfer functions will be the total round transfer function $-GH$. These measurement results indicate typical frequency responses of the LC filter and the error amplifier with the feedback compensation. In this way, you can check the phase delay of the LC filter and the effects of the feedback compensation circuits at the error amplifier in the converter's actual operating conditions by just changing the probing points in the same measurement configuration as the loop gain measurement.

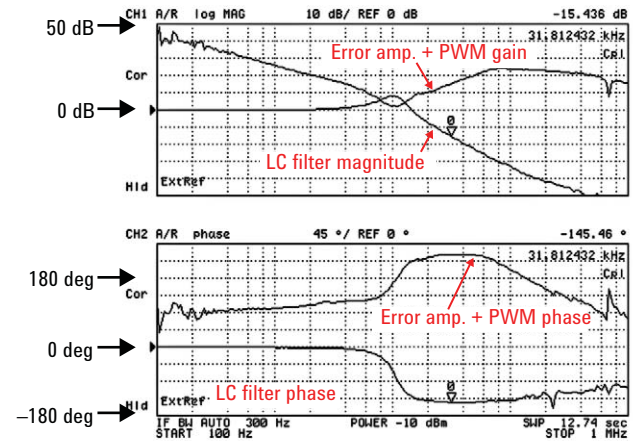


Figure 10. Transfer function of LC filter block and error amplifier-to-PWM block

Injection signal level

Now we assume that the constant floating AC voltage is injected at the resistor R5, which is in the secondary side of the injection transformer, regardless of the test frequency range. This floating AC voltage is divided into two AC voltages referenced to GND (or common) at the receivers R and A according to the loop gain at each frequency point. In the low frequency range where the loop gain is high, the small AC voltage appears at the receiver R, and the large AC voltage appears at the receiver A. As the frequency increases, the AC voltage at the receiver R increases, and the AC voltage at the receiver A decreases. At the crossover frequency where the loop gain is 0 dB, the same level of the AC voltage appears at both receivers R and A.

As for the injection signal level, in general, it is desired to inject a large AC signal in the low frequency range where the AC voltage that appears at the receiver R becomes low and the measurement SNR tends to worsen. However, doing so will make the feedback loop circuit get into the

nonlinear region because the AC signal level entering into the error amplifier and the PWM becomes too large. Therefore, the injection signal level should be set to an appropriate level that is not too large and not too small, so that the DUT is kept in the linear region for the entire test frequency range and also the measurement SNR in the low frequency range. To choose the appropriate signal level, first set the analyzer's source level to a sufficiently low level (e.g. -20 or -30 dBm) and perform the measurement. Repeat the measurements while increasing the source level gradually, then choose the source level that is slightly lower than the level where the strange behaviors such as discontinuous waveforms start to appear. Also, if necessary, check the signal level by monitoring the absolute power at the receiver A, in addition to the ratio measurement A/R. Confirm that the measured power level at the receiver A is linearly increased in the signal levels you are selecting. Or monitor the signal waveforms in the feedback loop circuits with an oscilloscope and confirm that the waveforms are not distorted.

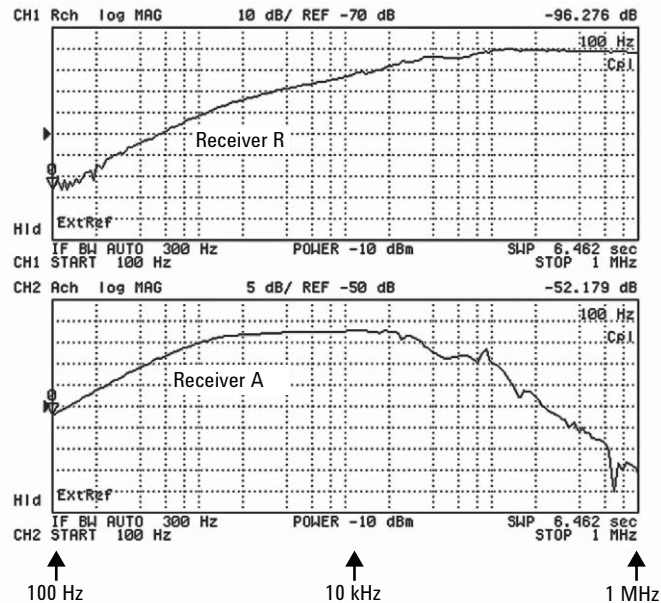


Figure 11. Absolute power measurements at receivers R and A in the loop gain measurement configuration

How to select injection transformer

Select the injection transformer with a flat transmission response in the entire test frequency range. The transformer should have an impedance that is not extremely smaller than the analyzer's source output impedance, 50 ohm. That is, the self inductance L should be large enough. Also, the transformer must work properly up to the high frequency region without making a self resonance. In the measurement example shown in Figure 9, the source signal was injected with the 1:1 pulse transformer whose self inductance is 3.4 mH (Agilent PN 5188-4425) in combination with a 50 ohm resistor as R5.

Figure 13 shows the transmission response of this pulse transformer, which is measured with the configuration shown in Figure 12. The measurement frequency range is 10 Hz to 10 MHz. As shown in the marker readings, the transformer gives a flat response in the high frequency range up to 1 MHz and more. On the other hand, in the low frequency range around 100 Hz, the 20 dB loss occurs because the transformer's impedance $|Z| = |j \cdot 2 \cdot \pi \cdot f \cdot L|$ is lower than the analyzer's source output impedance, 50 ohm, and the AC voltage applied to the primary side of the

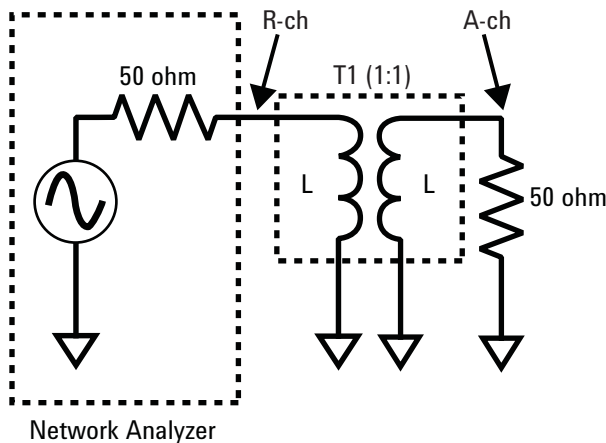


Figure 12. Evaluation method of the transformer's transmission characteristics

transformer becomes very small. This causes further disadvantage in terms of the measurement SNR because the 20 dB loss of the transformer is added on top of the originally small AC level due to the high loop gain in the low frequency range, which was discussed in the previous section. Figure 13 shows the absolute AC power measurement results at the receivers R and A by using the same configuration as the loop gain measurement. As you can see, the measured trace around 100 Hz is fluctuating because of these two factors. However, as mentioned before, the important frequency range for evaluating the feedback loop is generally around the crossover frequency, and the fluctuation at low frequencies is not a problem.

To further improve the loop gain measurement SNR in the low frequency range, use the transformer that has a flat response down to the low frequency range. If the resistor R5 is around 50 ohm, you can use 50 ohm isolation transformers. For example, the Video Isolation Transformer Model 0017CC (10 Hz to 5 MHz, 50 ohm to 50 ohm, BNC to BNC) supplied by North Hills Inc. (www.northhills-sp.com). Be careful not to apply any DC voltage between the high and low terminals of the isolation transformer.

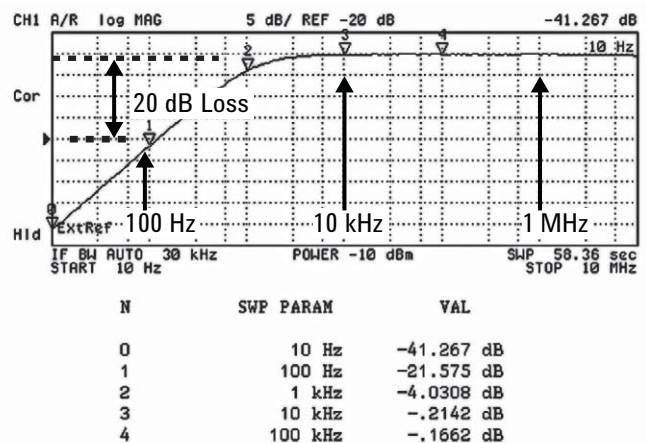


Figure 13. Transmission characteristic of pulse transformer (PN 9100-0855)

► Evaluating Output Impedance

As described in the beginning of this document, it is getting very important to evaluate the very small impedance of the power planes (Power Delivery Network: PDN) especially in the latest high-speed digital circuits that operate at low voltages. Here if we assume Z_{plane} is the power plane impedance that is seen from the load devices and ΔI is the current variation caused by the operation of the load devices, the voltage drop $\Delta V = \Delta I \times Z_{plane}$ will occur in the power plane (IR drop). In the low-voltage systems, ΔI can be several amperes or several tens of amperes, and the voltage drop ΔV will not be negligible. This may cause signal integrity problems and/or EMI problems. To prevent this, it is necessary to suppress the power plane impedance Z_{plane} to a small value in the broad frequency range from DC to GHz range, especially in the low frequency range, an extremely small impedance of milliohm order is often required. The DC-DC converter provides this extremely small impedance in the low frequency range. Regulating the converter's output voltage with the feedback loop control regardless of any load variation equivalently means that a very small output impedance is achieved. The relationship between the output impedance and the loop gain is given as $Z_{closed} = Z_{open} / (1 + GH)$, where Z_{open} is the open-loop output impedance, Z_{closed} is the closed-loop output impedance, and GH is the loop gain. The closed-loop output impedance will be very small in the low frequency region where the loop gain is high.

To evaluate the output impedance of the DC-DC converter, we directly measure the closed-loop output impedance Z_{close} by probing the output terminal of the DC-DC converter that is working in the closed-loop condition by using the LF network analyzer. To measure the very small impedance while blocking the DC output coming from the DUT, we need a measurement approach which is different from general impedance measurements for passive components. The following section describes how to make the output impedance measurements of the DC-DC converters.

The measurement concepts described here are basically applicable to general switch-mode power supplies, and to the PDN impedance measurements in the frequency range up to several tens of MHz. (The calibration methods and probing methods for high-frequency PDN impedance measurements are not covered in this application note.)

Current-voltage detection method

Assuming that the receiver R is floating, we consider the measurement configuration shown in Figure 14-a. The DUT in this figure corresponds to the DC-DC converter and the load connected to it. The analyzer actually measures the parallel impedance of the DC-DC converter and the load, but generally the output impedance of the DC-DC converter is dominant because it is much smaller than the load impedance. The blocking capacitor prevents the DUT's DC output from flowing into the analyzer. Its impedance $|Z| = |1/(j \cdot 2 \cdot \pi \cdot f \cdot C)|$ should be sufficiently small. The receiver A detects the AC voltage V_{dut} across the DUT, and the receiver R detects the AC voltage across the 1 ohm resistor, which is equivalent to the AC current I_{dut} that flows through the DUT. The ratio measurement A/R gives the DUT's impedance because of the equation $A/R =$

$V_A/V_R = V_{dut}/(1 \times I_{dut})$. This measurement configuration provides good measurement sensitivity for small impedance measurements. Since the receivers of general-purpose network analyzers are not floating, it is necessary to use a high-power differential probe to make the floating input.

The equivalent configuration shown in Figure 14-b, where the source is configured to floating instead of the receiver by using the transformer, which is similar to the loop gain measurement configuration. This configuration also detects I_{dut} and V_{dut} with the receivers R and A, and the ratio measurement A/R directly indicates the DUT impedance with good sensitivity for the small impedance. This is the output impedance method with the current-voltage detection by using the floating signal source. This method is traditionally often used for evaluating the output impedance of the switch-mode power supplies.

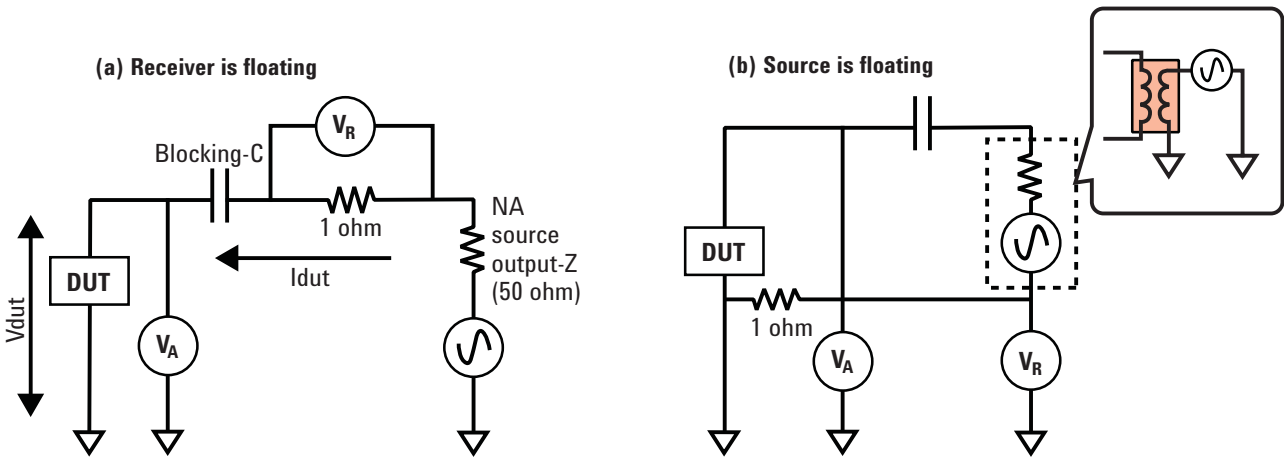


Figure 14. Block diagram of output impedance measurement with current-voltage detection

Configuration example of current-voltage detection method

Figure 15 shows a configuration example of the current-voltage detection method. For transformer T1, you can use the same pulse transformer that was used for the loop gain measurement. (Since the transformer is terminated with the small impedance, isolation transformers designed for the 50 ohm system impedance are not recommended.) The large-capacitance electrolytic capacitor of 1 mF is used as the blocking capacitor to sufficiently inject the source signal from the low frequency range. The connections between the high-impedance receivers and the DUT are made with test cables such as BNC-to-alligator clip cables instead of the 10:1 high-impedance passive probes to improve the measurement SNR. Since the DUT's impedance is small, the degradation of the probe input impedance at high frequencies is not a problem. But the test cable length should not be very long because the cable capacitance may affect the loop characteristics. The leads connected to the GND should be as short as possible because their lead inductances may affect the small impedance measurement at high frequencies.

The 1 ohm resistor for sensing the AC current should be as precise as possible. By measuring the impedance of the resistor before measuring the DC-DC converter, you can do compensation using simple math. For example, if the impedance of the resistor is 0.98 ohm, you can obtain the DUT impedance Z_{dut} by multiplying the measured impedance Z_{meas} with 0.98, because Z_{meas} is calculated as $Z_{meas} = V_A/V_R = V_{dut}/(I_{dut} \times 0.98) = Z_{dut}/0.98$. If you use the lead resistor, make the lead length as short as possible to reduce the measurement error at high frequencies. You do not have to use a high-power resistor because the DC current does not flow through this resistor due to the blocking capacitor. To calibrate the measurement system, perform the response thru calibration by connecting the test cable of the receiver A to the same point as the receiver R.

The advantage of this measurement method is that the probing measurement can be easily made because the complicated calibration is not needed. However, as the measurement system is not strictly calibrated, this method is not suitable if you want to precisely measure very small impedance.

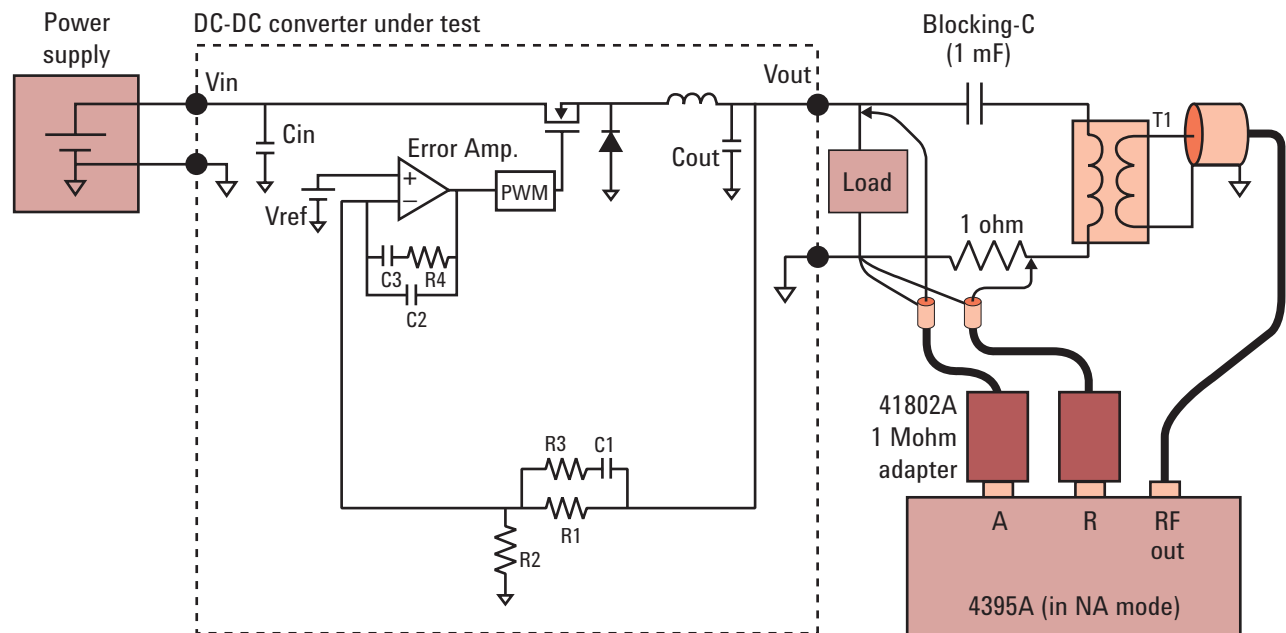


Figure 15. Configuration example of current-voltage detection method for output impedance measurement

Shunt-thru method

Another measurement approach that gives more accurate small impedance measurements, next we consider the shunt-thru method which is commonly used for small impedance measurements in the high frequency range. Figure 16 shows the measurement block diagram. The transmission coefficient S_{21} is measured by connecting the DUT in the shunt connection between the signal line and GND, and the impedance is derived from S_{21} . The relation between the DUT's impedance Z_{dut} and S_{21} is given as $Z_{dut} = 25 \times S_{21} / (1 - S_{21})$ in the shunt connection.

The DC output coming from the DUT is blocked by the blocking capacitors (If your network analyzer can receive the DC voltage, the blocking capacitors are not necessary.) Use the blocking capacitor whose impedance $|Z| = 1 / (j \cdot 2 \cdot \pi \cdot f \cdot C)$ is much smaller than the analyzer's system impedance, 50 ohm, to obtain a good measurement SNR in the low frequency range.

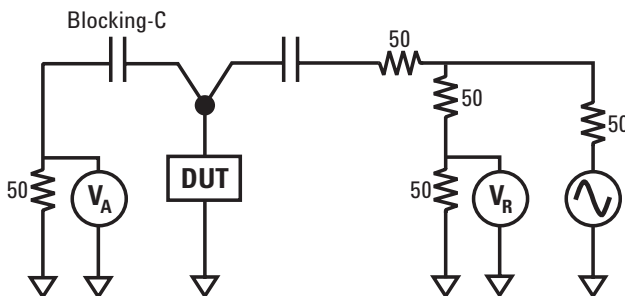


Figure 16. Shunt-thru measurement method

Measurement error factor of shunt-thru method

While the complicated 2-port full calibration is required in the high frequency shunt-thru measurements, you can use simpler calibration methods in the low frequency measurements below tens of MHz. However, it is difficult to measure the small impedance below 10 milliohm in the low frequency range, or below 10 kHz only with the response thru calibration, because it cannot eliminate the measurement error caused by the outer conductor resistance of the coaxial test cables (cable braid impedance) that occurs in the low frequency range near DC. The cable braid impedance is generally several tens of milliohms, although it actually depends on the thickness of the outer conductor and the cable length. The block diagram shown in Figure 17 roughly explains the measurement error caused by the cable braid impedance. Ideally, the measured voltage V_A should be close to 0 V when the DUT's impedance Z_{dut} is very small and the loss of the blocking capacitor is eliminated with the response thru calibration. But actually the measured voltage V_A cannot be lower than the voltage V_{cbl} due to the cable braid impedance R_{cbl} , and the measured impedance cannot be lower than R_{cbl} even when Z_{dut} is close to 0 ohm.

The hardware approaches for solving this problem are to attach ferrite cores to the test cables, or to connect an isolation transformer to isolate the grounds between the port-1 and port-2 as shown in Figure 18 [1]. Figure 19 compares the S_{21} measurement traces of a SHORT device with and without the isolation transformer after the thru calibration is performed (displayed in the linear magnitude format). As you can see, the impedance in the low frequency range is close to the ideal value (zero ohm) when using the isolation transformer. The isolation transformer used in this example is the same pulse transformer as that used in the loop gain measurement. It must be noted that the blocking capacitor is necessary to prevent the isolation transformer from being saturated by the DC current.

Next, we consider the solution by performing the zero compensation with a SHORT device as shown in Figure 20, which is a quite different software approach to solving the problem without using any additional hardware devices, such as isolation transformers or ferrite cores. The easiest way is to perform the isolation calibration by connecting the SHORT device in addition to the response thru calibration. The isolation calibration will mathematically compensate the SHORT measurement data closer to zero in the low frequency range. But the thru & isolation calibration cannot fully eliminate the error factors of the measurement system especially in the high frequency range. For more accurate measurements, the OPEN/SHORT/LOAD calibration is recommended, rather than the thru & isolation calibration. The OPEN/SHORT/LOAD calibration adjusts the measurement system with three standard devices and enables more accurate measurements up to the higher frequency range, not just improving the small impedance measurement range at the low frequency range due to the SHORT calibration.

The OPEN/SHORT/LOAD calibration can be executed by making the thru connection as the open, connecting a good shunting plate and a 50 ohm precision resistor as the short insert, and load conditions. If you use the network analyzer that allows you to perform the OPEN/SHORT/LOAD calibration in the impedance domain (not in the S-parameter domain), the impedance values can be directly read out by performing the OPEN/SHORT/LOAD calibration in the impedance domain, without regard to the conversion equation from S21 to the shunt impedance. One thing that must be noted is the measurement may be unstable if you perform the OPEN/SHORT/LOAD calibration, if the cable braid impedance R_{cbl} is extremely large, and the residual impedance of the SHORT measurement is too high. It would be possible to combine a hardware approach using the isolation transformer and the OPEN/SHORT/LOAD calibration, but in general, you can obtain practically enough measurement quality down to the milliohm range only with the OPEN/SHORT/LOAD calibration if you use general high-performance coaxial test cables whose diameter is not extremely small.

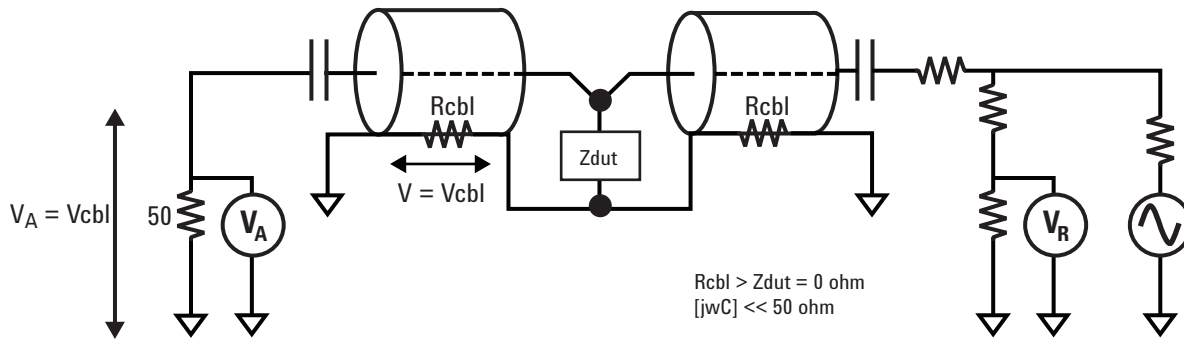


Figure 17. Error caused by cable braid impedance

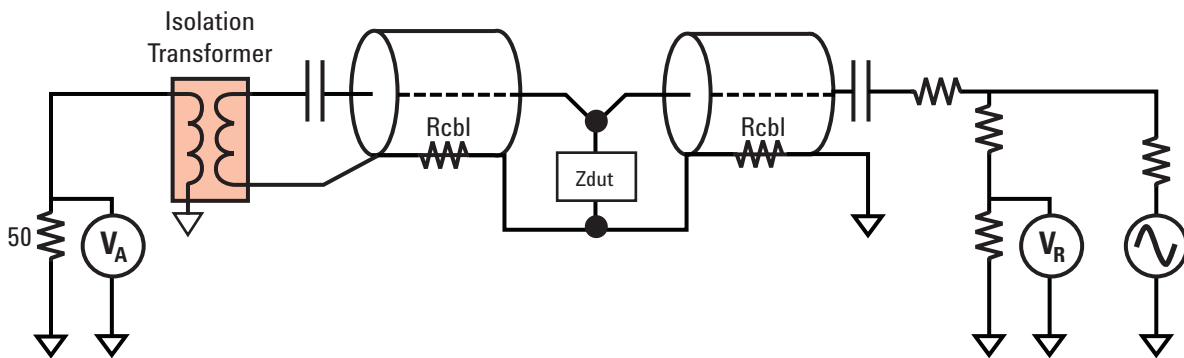


Figure 18. Solution with isolation transformer

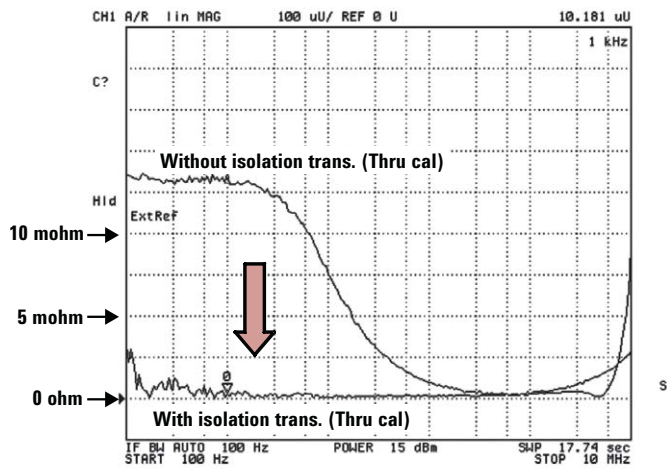


Figure 19. Effect of isolation transformer (SHORT device measurement)

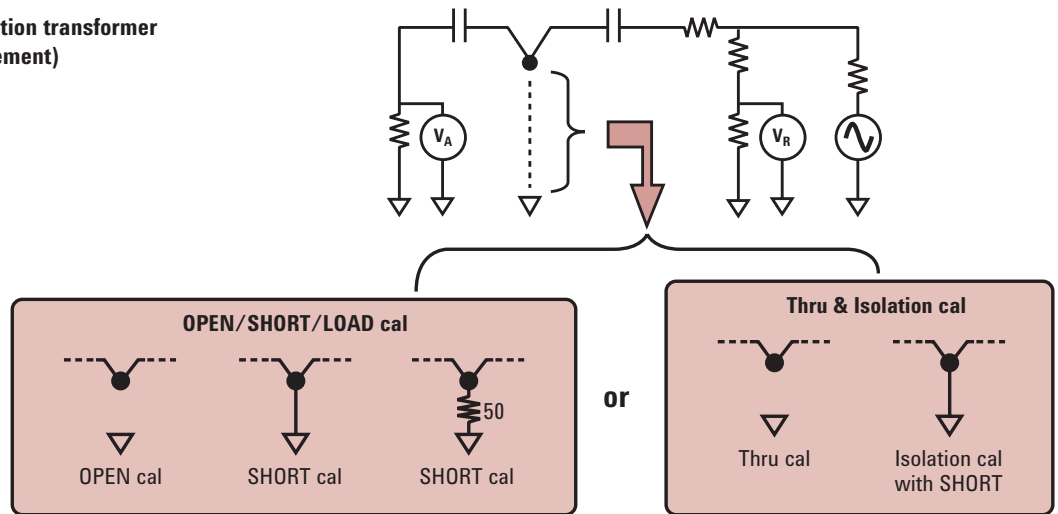


Figure 20. OPEN/SHORT/LOAD calibration, thru & isolation calibration

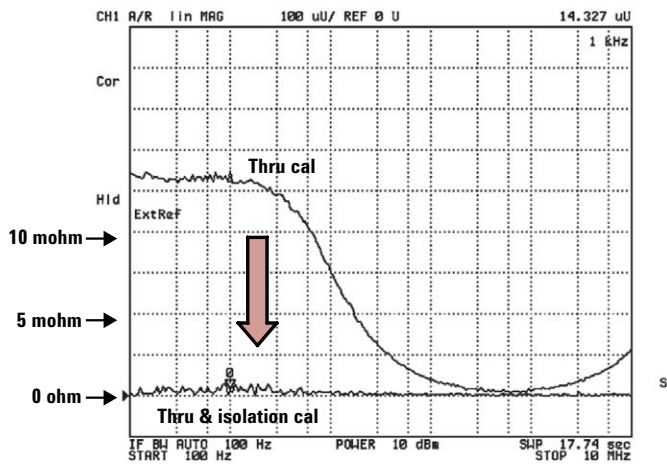


Figure 21. Calibration with SHORT device
Example of thru & isolation cal. OPEN/SHORT/LOAD cal gives the same effect.

Configuration example of shunt-thru method - 1

Figure 22 shows a configuration example of the shunt-thru method. The network analyzer used in this example is the 4395A with option 010 impedance measurement firmware, and the measurement is performed in its impedance measurement mode. Similarly to the current-voltage detection method, the DC output coming from the DUT is blocked with the 1 mF electrolytic capacitors.

Figure 23 shows an example of probing. Two home-made probes are connected to the end of the test cables and the probes are contacted to the output terminal of the DUT to measure its output impedance. The home-made probe can be made for example by utilizing a SMD receptacle (cutting off its three GND pins and using the remaining GND pin and the center pin for probing) or a SMA semi-rigid cable (cutting it short, stripping the center conductor,

and soldering a short pin to the outer conductor). To calibrate the measurement system, perform the OPEN/SHORT/LOAD calibration at the end of the probes as shown in Figure 23. To measure the very small impedance of milliohm order, the probes should be precisely soldered to the DUT and the calibration devices so that the contact resistance can be minimized. When soldering the devices to the probes, disconnect the probes from the analyzer. If you use SMA receptacles as the probes, it is convenient to prepare the adapters of the open, short and load conditions beforehand and perform the OPEN/SHORT/LOAD calibration by using these calibration adapters. As for the load calibration device, you can use a 50 ohm radial-leaded resistor Agilent PN 0699-2014 (lead distance 5 mm) or a 50 ohm SMD resistor Agilent PN 0699-2829 (3.2 mm x 1.6 mm). For the SHORT device, solder a good shorting plate such as a copper plate to the probes to minimize the SHORT residual inductance.

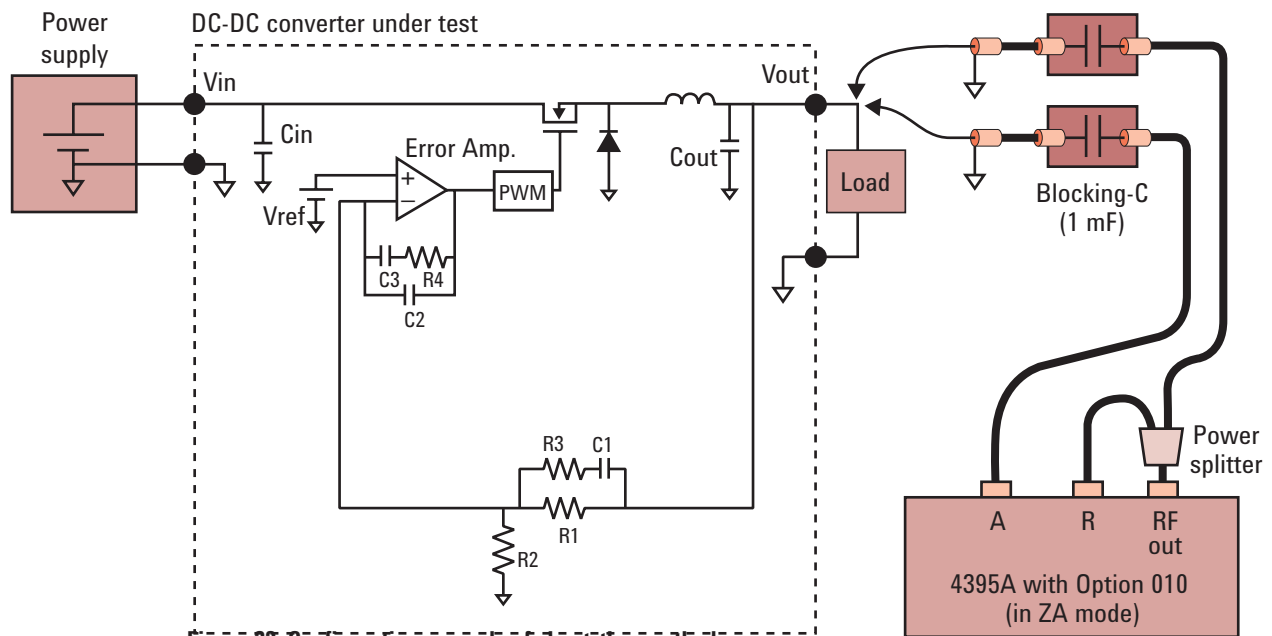


Figure 22. Configuration example of shunt-thru method

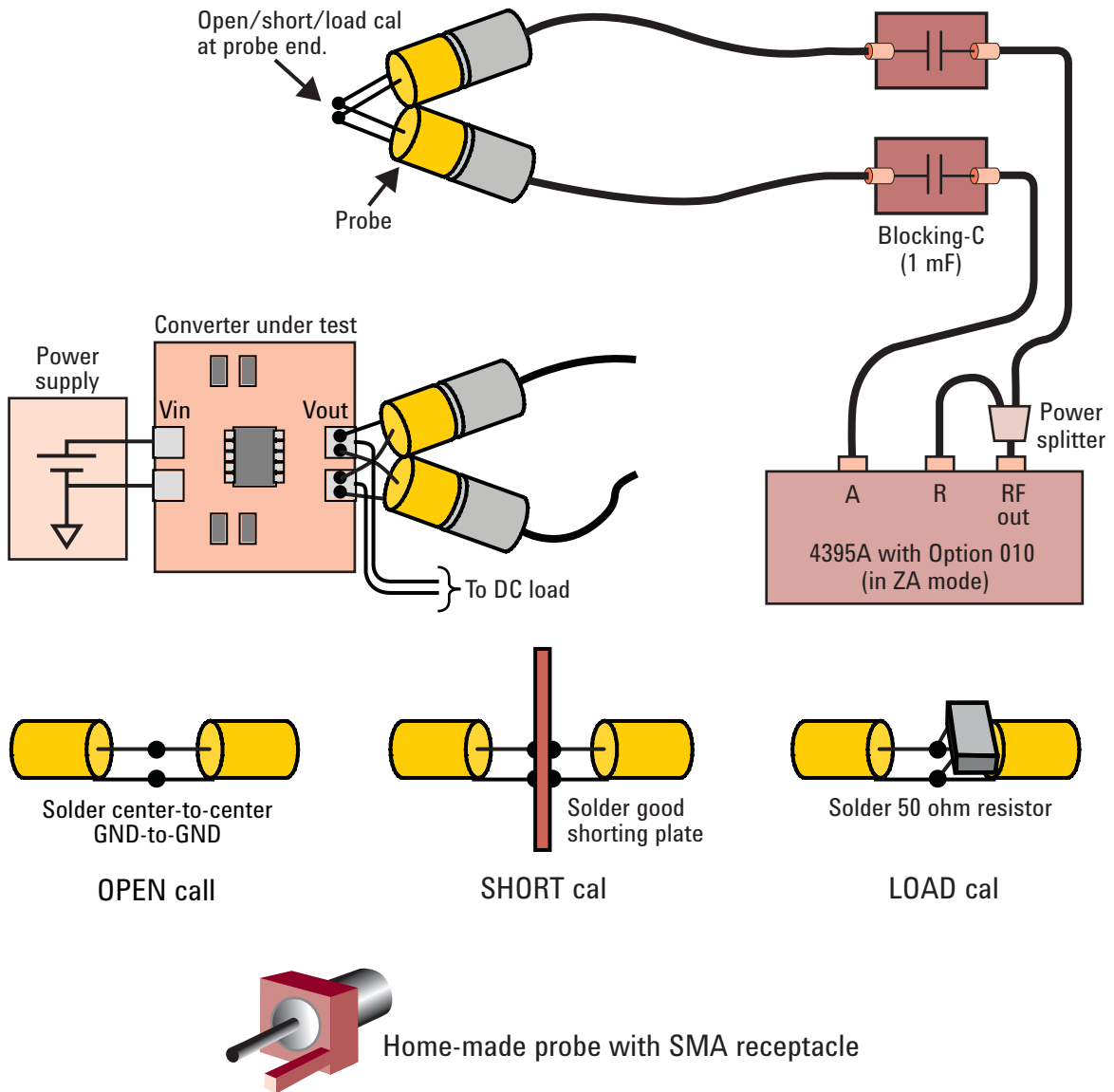


Figure 23. Example of probing method for shunt-thru method (1)

Configuration example of shunt-thru method - 2

By modifying this 2-port probing shunt-thru configuration, we can make the simpler 1-port probing configuration shown in Figure 24. The two test cables are converted to the 1-port configuration by using the SMA T-connector, and the probe is connected to the T-connector. Although this method is applicable just up to several tens of MHz (not applicable to higher frequency PDN measurements), it provides the same good sensitivity for small impedance measurements as the 2-port probing configuration with the simpler 1-port probing. Also, this configuration allows you to make the OPEN/SHORT/LOAD calibration at the end of the T-connector by using the 3.5 mm coaxial calibration kit and then make the SHORT compensation to remove the effect of the probe's residual impedance, which reduces the burden of the calibration using home-made calibration kits. Although the 2-port probing is generally better than the 1-port probing in terms of minimizing the effects of the contact resistance and inductance [1] [2], the 1-port probing provides practically enough performance if the

measurement frequency is low (below 10 MHz) like the application discussed here and if the probe is precisely soldered to the DUT to minimize the contact resistance. The probe should be directly connected to the T-connector without using any further cable extension to make the length between the T-connector and the probe-end as short as possible so not to degrade the small impedance measurement sensitivity. Also, the blocking capacitors should be placed after the signal path is divided into the two port configuration as shown in the Figure 24. Placing the blocking capacitor between the probe and the T-connector will degrade the measurement sensitivity for the small impedance.

In the measurement configurations shown here, the very large blocking capacitor is used. In general, this will not have an affect on the DUT's loop characteristics (the output LC filter characteristics) because the dumping resistor (the analyzer's port resistance, 50 ohm) which is much larger than the ESR of the DUT's output capacitor exists in series with the blocking capacitor.

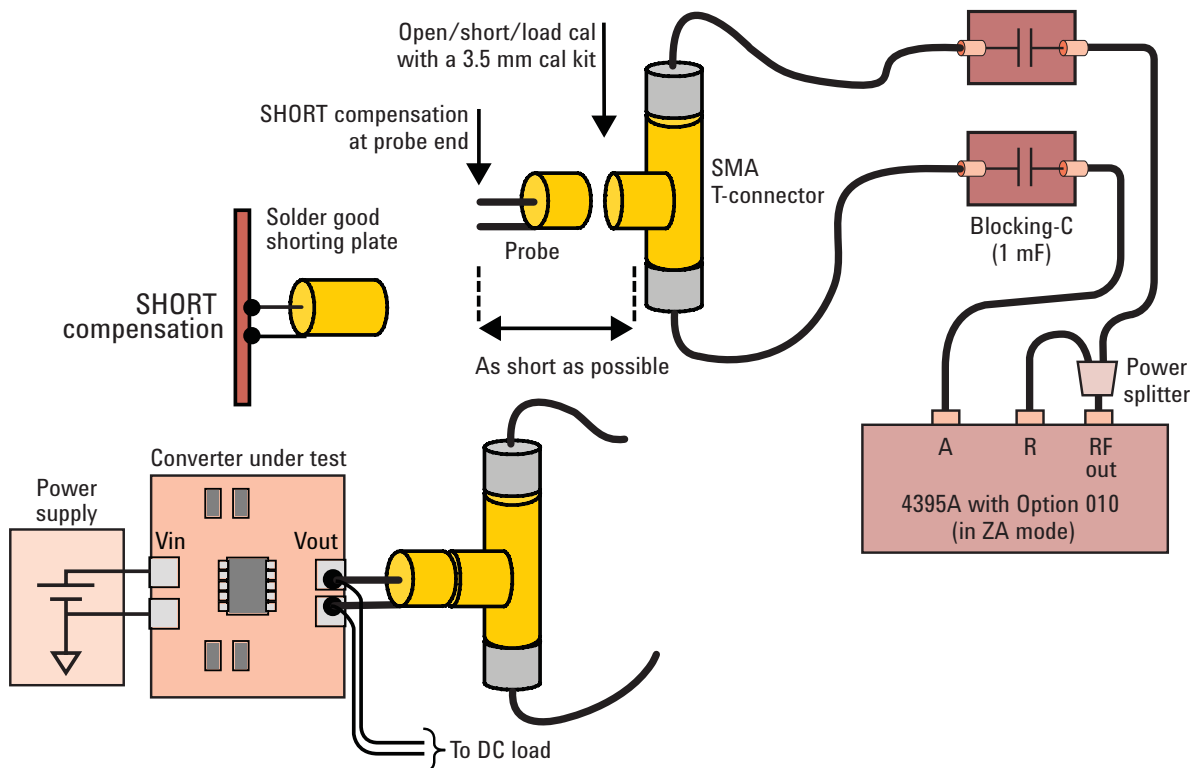


Figure 24. Example of probing method for shunt-thru method (2)

Measurement example of output impedance

Figure 25 shows an output impedance measurement example of the 5 V-3.3 V buck converter, which is the same one measured in the loop gain measurement example, by using the shunt-thru measurement configuration shown in Figure 24. The measurement frequency range is from 100 Hz to 10 MHz. The IFBW is set to 10 Hz to measure the small impedance with a good measurement SNR in the entire frequency range. The source signal level is set to a relatively small level 0 dBm, during the OPEN/SHORT/LOAD calibration not to overload the receiver of the 4395A being used in the impedance analyzer mode when the thru connection is made for the OPEN calibration, and then increased to a higher level, 10 dBm, to improve the measurement SNR. Since the DUT's impedance is much smaller than the analyzer's source output impedance, 50 ohm, excessive signal level is not applied to the DUT even if the source level is set to high.

In this measurement example, the output impedance characteristics of the converter's 1 A load, 3 A load, and off-state conditions (the input voltage to the converter is set to zero and DC electronic load is turned off) are plotted. The off-state output impedance indicates a

self-resonant impedance response of the output capacitor. If the converter is turned on, the impedance below the converter's crossover frequency (marker point, 30 kHz) is suppressed to the very small impedance due to the feedback loop control as shown in the traces of the 1 A and 3 A load conditions. The impedance around the crossover frequency is 29 mohm, and the impedance values have become very small, below 10 mohm in the lower frequency range where the loop gain gets higher. Now looking at the worst impedance value in the loop bandwidth, 29 mohm, if there is a load current variation of 1 A, the voltage variation would be roughly estimated as $1 \text{ A} \times 29 \text{ mohm} = 29 \text{ mV}$.

Comparing the measurement results of the 1 A and 3 A load conditions, the output impedance is mostly unchanged, although a small variation occurs only in the very low frequency range. Like this example, it is important to evaluate the output impedance under various load conditions and verifying the converter provides stable operations with small output impedance variations. Also, another important evaluation point is that the output impedance trace should not have a large positive peak that may cause transient noise.

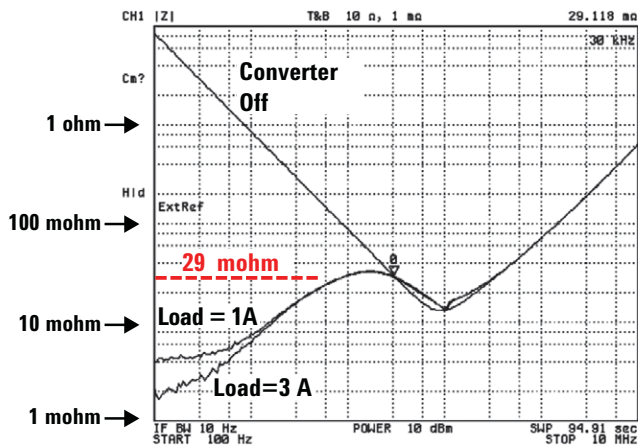


Figure 25. Output impedance measurement example with shunt-thru method

► Evaluating Impedance of Passive Components

To improve the quality and efficiency of the feedback loop circuit design of the DC-DC converters, it is important to evaluate the impedance characteristics of passive components used in the output LC filters and the feedback compensation circuits. Although the dedicated impedance analyzers such as the 4294A are the best solutions in terms of measurement accuracy, the 1-port reflection measurement with the LF network analyzers can provide practically enough measurement accuracy for this component evaluation purpose with the impedance range of 10 mohm to 10 kohm from the low frequency range. As for the test fixtures, you can use the existing 7 mm type component test fixtures because the measurement configuration is not 2-port but 1-port.

Evaluating large capacitance MLCCs

An example of passive component evaluation, here we look at the measurement for large-capacitance MLCCs (Multi-layer Ceramic Capacitors). The MLCCs with very low ESR and ESL are becoming widely used as the output capacitors of the DC-DC converters to reduce the output ripple noise. Since the large-capacitance MLCCs with high dielectric constants have an AC voltage level dependency and a DC voltage level dependency, the MLCCs give different capacitance from their nominal values in their actual operating conditions of the DC-DC converters. The nominal values of the MLCCs are generally measured at 120 Hz by applying the AC voltage level of 1 Vrms and no DC bias voltage. However, when the MLCCs are used as the output capacitors of the DC-DC converters, DC voltages (the converter's output voltages) and much smaller AC voltages (the output ripples) are applied to them. To accurately design the feedback loop circuits, it is important to evaluate the MLCC capacitance under the conditions similar to the converter's actual operating conditions by applying the DC bias voltages and the small AC voltages.

Figure 26 shows a configuration example for this measurement. The 4395A network analyzer with the option 010 impedance measurement firmware and the option 001 DC voltage source is configured to the reflection impedance

measurement. The DC bias is applied via the home-made DC bias circuit that consists of a blocking capacitor and a 1 kohm resistor. (The home-made DC bias circuit is not necessary if you can apply the DC bias from the analyzer.) The measurement system is calibrated with the OPEN/SHORT/LOAD calibration at the 7 mm connector plane and the OPEN/SHORT compensation at the 16092A 7 mm type test fixture.

To measure the MLCCs that have very low ESR and ESL as accurately as possible with the LF network analyzers and test fixtures of 2-terminal contact, it is necessary to perform the SHORT compensation accurately by using the SHORT device that has a small residual inductance. In the case of the 16092A test fixture, use the dedicated SHORT device provided with the 16092A as shown in Figure 27-a. Also, the contact to the DUT should be a point-contact or a line-contact to minimize the contact variations. Furthermore, the DUT should be held at the same position as the SHORT device to minimize the SHORT compensation errors. In the example shown in Figure 27-b, the DUT is contacted with the line-contact at the edge of the fixture electrodes, and the residual inductance of the fixture can be correctly compensated because the length L2 is the same as the length L1 of the SHORT compensation. On the other hand, in the example shown in Figure 27-c, the contact variations are likely to occur with the plane-contact, and the SHORT compensation subtracts the fixture residual inductance too much because the length L3 is shorter than the length L1, which makes the ESL measurement errors in the frequency range higher than the self-resonant frequency.

By taking care of above mentioned points, the practical measurement accuracy can be considerably improved. However, if you need to evaluate the very small ESR and ESL very strictly, it is recommended to use the impedance analyzer of the auto-balancing bridge method, such as 4294A, and the 16044A Kelvin contact test fixture.

For more details about the fixture compensation and other tips on component impedance measurements, refer to the Impedance Measurement Handbook [3].

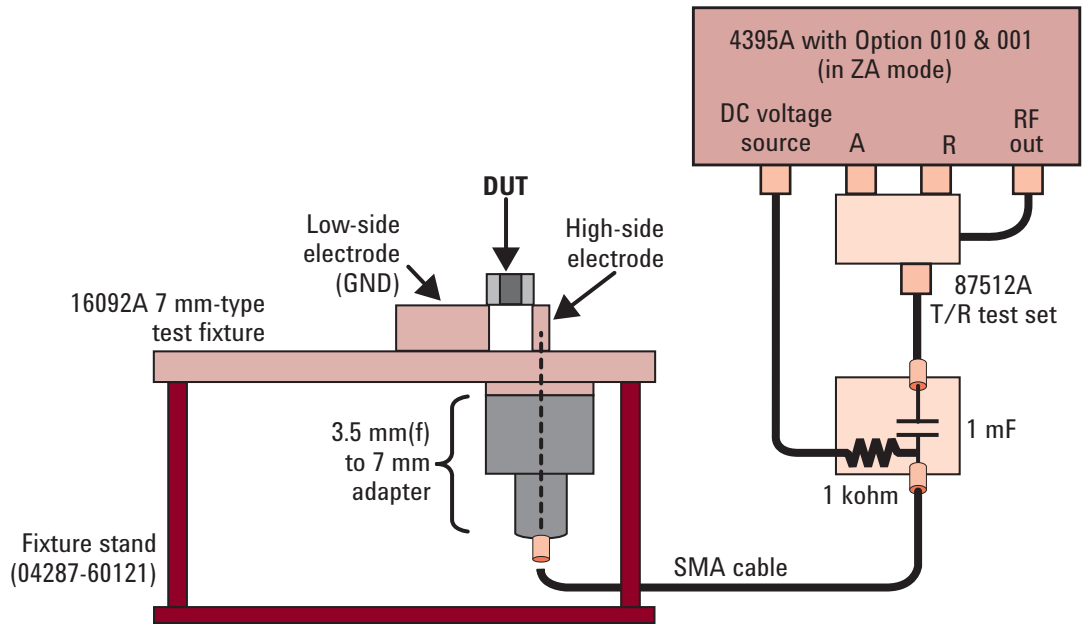


Figure 26. Configuration example for measuring passive component impedance

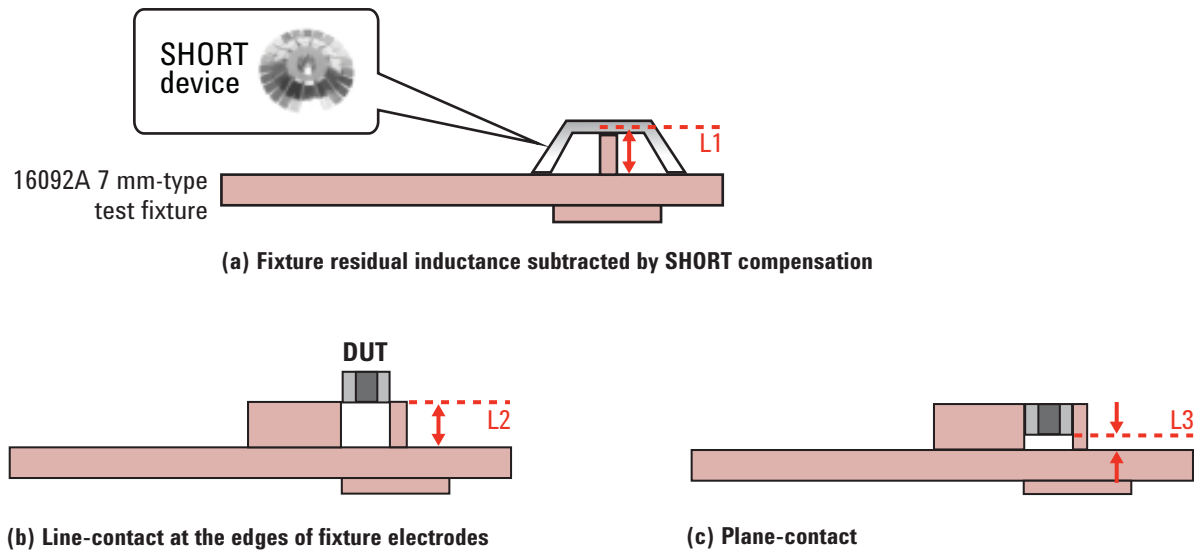
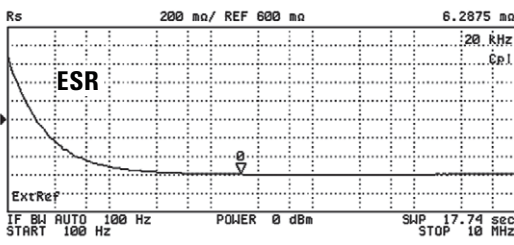
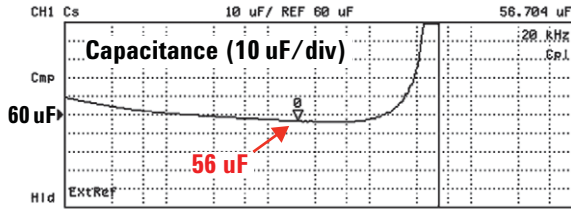
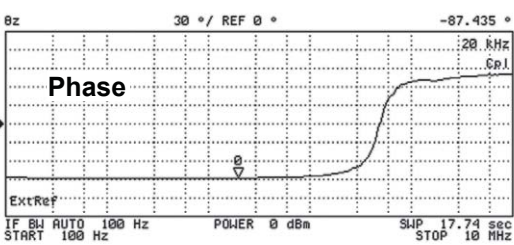
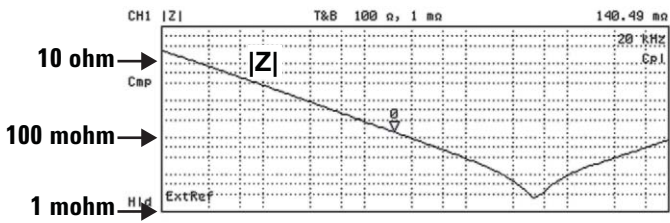


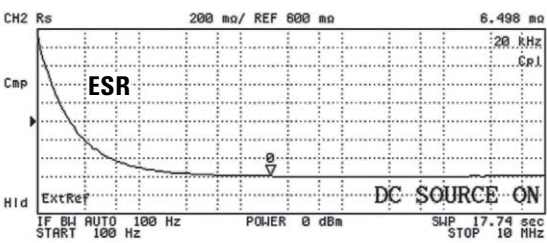
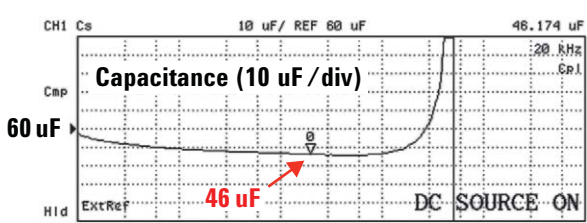
Figure 27. Tips for measuring low-ESR and low-ESL capacitors



(a) C-ESR characteristics (bias = 0 Vdc)



(b) |Z|-phase characteristics (bias = 0 Vdc)



(c) C-ESR characteristics (bias = 3.3 Vdc)

Measurement example

A MLCC of the nominal capacitance 100 uF was measured with the configuration shown in Figure 26. Figure 28-a and 28-b show the measured frequency responses of the C-ESR and |Z|-phase without applying the DC bias voltage. The measurement frequency range is 100 Hz to 10 MHz. As shown in Figure 28-a, the capacitance at 100 Hz is about 70 uF, which is smaller than the nominal value, 100 uF. This is because the AC voltage level applied across the DUT is smaller than 1 Vrms, which is the level that gives the nominal capacitance value. The AC voltage applied to the DUT at 100 Hz is about 50 mVrms because the source output level at the test port of the reflection bridge is -10 dBm (= 71 mVrms when terminated with 50 ohm) and the DUT's impedance $|Z| = 1/(j \cdot 2 \cdot \pi \cdot f \cdot C)$ is about 27 ohm at 100 Hz. The capacitance at the marker point 20 kHz, is 56 uF ($|Z| = 0.14$ ohm). The AC voltage applied across the DUT at 20 kHz is below 1 mVrms. Figure 28-c shows the measurement result with 3.3 Vdc DC bias voltage assuming the capacitor is used in the 3.3 Vdc converter. As you can see, the capacitance values become smaller, 60 uF at 100 Hz and 46 uF at 20 kHz.

References

- [1] *Frequency Domain Characterization of Power Distribution Networks*, Istvan Novak, Jason R. Miller (2007, Artech House)
- [2] *Ultra-Low Impedance Measurements Using 2-port Measurements*, Agilent Application Note, Literature Number 5989-5935EN
- [3] "Impedance Measurement Handbook", Agilent Literature Number 5950-3000
- [4] *Switching Power Supply Evaluation*, Agilent 4395A Product Note, Literature Number 5968-7274E

Figure 28. Measurement example of large-capacitance MLCC

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