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Measuring MilliOhms and PicoHenrys in Power-Distribution Networks

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Abstract:

The real challenge in today's power-distribution systems is to provide a sufficiently low impedance over a wide frequency range. For many systems, the impedance should be a few milliohms or lower at low frequencies and a few times ten picohenry's impedance at high frequencies. Earlier [2] the two-port self-impedance measurement setup was shown to extend the capabilities of one-port vector network analyzer measurements by about an order of magnitude, enabling the measurement of impedances as low as a few times ten milliohms.

This paper applies the two-port self-impedance measurement concept and extends the setup arrangement in different ways at low and at high frequencies, extending the impedance measurement to the sub-milliohm range. It was found that at low frequencies, the limitation of single-ended two-port self-impedance measurement stems from the residual attenuation formed by the ground loop of the connecting cables' shields. To reduce this error, an isolating transformer and/or a floating-input differential isolation amplifier is used. As a result, measurements with an HP4395B VNA show a residual error of a fraction of a milliohm. At high frequencies, the limiting factor turns out to be the finite surface-transfer impedance of the connecting cables and the resonance peaks in the braid current due to the large mismatch from the low DUT impedance. Placing a series of absorbing ferrite clamps around the coax cables reduces this limitation. Measurement results with an HP4396B VNA show residual inductance readings in the order of a few pHs.

The measurement concepts are illustrated on Voltage Regulator Modules (VRM) over the frequency range of 100Hz-1MHz (with HP4395B) and various power distribution networks over the frequency range of 100kHz - 1GHz (HP4396B).

The measurement results were compared against simulated impedances of the same structures.

Slide#1

Measuring Milliohms and PicoHenrys in Power Distribution Networks

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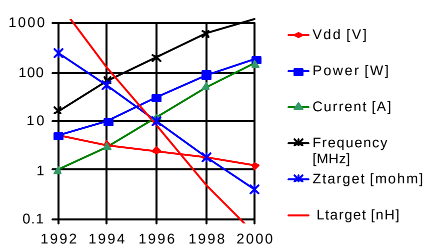
Slide#2

Outline

- Introduction
- Two-port VNA impedance measurements
- Low-frequency limitations
- Enhancement with transformer or amplifier
- High-frequency limitations
- Enhancement with ferrite-covered cable
- Measured power-distribution networks
- Resources
- References

Slide#3

Requirements in Power Distribution



With the constant decrease of supply voltages and signal transition times, rise of supply currents and clock speeds, the power-distribution networks need to provide lower impedances over wider bandwidth [1]. As shown on the chart in Slide#3, the typical target impedance has dropped by a factor of five in every two years. When combined with the increasing bandwidth of systems, the high-frequency equivalent inductance requirement drops even more drastically, by about ten times in every two years. While techniques to verify the signal integrity of high-speed signals have been developed and are widely available, the measurement of low impedances at high frequencies create new challenges in the verification process.

It was shown [2] that the four-wire DC resistance measurement concept can be applied at high frequencies by using Vector-Network Analyzers (VNA), and two-port impedance measurements can greatly reduce the effect of series connection discontinuities, the biggest error contributor of one-port measurements.

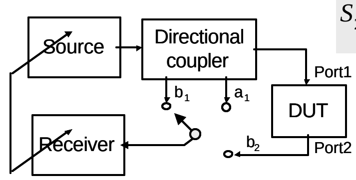
Slide#4

What is a VNA

- Tuned sinewave generator
- Directional couplers
- Tracking receiver(s)

$$S_{11} = \frac{b_1}{a_1}\Big|_{a_2=0}$$

$$S_{21} = \frac{b_2}{a_1}\Big|_{a_2=0}$$



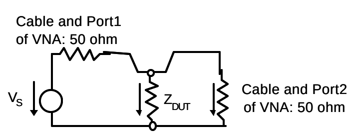
The vector-network analyzer is an instrument for measuring scattering parameters [3]. A synthesized tuned source with calibrated source impedance and calibrated source voltage generates the incident wave at the excited port. The signal is connected to Port1 of the device under test through a directional coupler which separates the a_1 and b_1 incident and reflected waves, respectively. Port2 of the device is terminated in the reference impedance, and the wave appearing at that port is b_2. A selective receiver is used to measure the incident and reflected waves at the ports. If the network to be measured has more than two ports, we can excite one by one all of the ports.

Because the frequency of the test signal is exactly known, the sensitivity and accuracy of the measurement is readily improved by locking the receive frequency to the transmit frequency, and/or by using averaging.

Slide#5

Two-Port Self-Impedance Measurement

- S_{21} instead of S_{11} is measured
- S_{21} uncertainty is less
- Z_p is in series to 50 ohms instead of Z_{DUT}

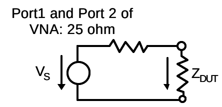


To achieve reasonable accuracy for low DUT impedance values, the two-port self-impedance measurement makes use of both ports of the vector-network analyzer, and its operating principle is similar to the four-wire measurements of very low resistances. Rather than sending the test signal and sensing the receive signal on the same port of instrument, in the two-port measurement Port1 is used only to launch a current through the unknown impedance, while Port2 is used to measure the voltage drop across the unknown impedance. This way we take the S_{21} transfer parameter reading from the VNA instead of S_{11}. We need to remember though that in this particular connection arrangement the S_{21} transfer parameter refers to one single node on the DUT, since both Port1 and Port2 of the VNA are connected to the same point on the DUT. With low impedances, the uncertainty of S_{21} measurements is much less than that of the corresponding S_{11} readings.

Slide#6

Two-Port Self-Impedance Reading

First-order calculation:
Assume that
- $L_p \sim 0$
- $Z_{DUT} << Z_0$



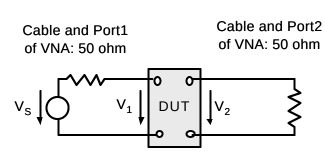
$Z_{DUT} = Z_{11} = S_{21} * 25$ [ohm]

In a first-order approximation, we may neglect the discontinuity, which is - as opposed to one-port measurements - in series to 50 ohms rather than the very low unknown impedance. With these assumptions, the equivalent circuit of the setup is a voltage divider between the two times 50 ohms port impedances of the VNA in parallel and the unknown impedance of device under test. If we calibrate and set the S_{21} reading of VNA without the DUT to zero dB, the unknown impedance is simply $Z_{DUT} = S_{21}*25$. Here both Z_{DUT} and S_{21} may be complex values, and S_{21} is the dimensionless ratio of the output and input waves. If the magnitude of S_{21} is obtained on a dB scale from the VNA, it can be converted into its dimensionless equivalent by using:

$$S_{21} = 10^{\frac{S_{21}[dB]}{20}}$$

Slide#7

Transfer Impedance Measurement



In the transfer-impedance measurement setup, Port1 of the VNA is used to launch a current at a selected input point of the DUT, while Port2 is used to measure the voltage at another point of the DUT. This time the S_{21} transfer parameter reading from the VNA is a true transfer parameter since the VNA ports are connected to different points on the DUT. With two selected connection points on the DUT, the transfer-impedance measurement can be taken either in one or the other direction. However, with just RLC elements in the bypass-capacitor bank and with commonly used printed-circuit-board planes, the power-distribution network is reciprocal. The reciprocity means that transfer parameters are the same regardless of the direction we look at them, therefore we know that for all i and j: $Z_{ij}=Z_{ji}$.

Slide#8

Transfer Impedance Reading

First-order calculation:

Assume that

- $L_p \sim 0$
- $Z_{11} << Z_0$
- $Z_{22} << Z_0$
- $Z_{21} << Z_0$

$Z_{21} = Z_{12} = S_{21} * 25$ [ohm]

As a first-order approximation, the transfer-impedance value from the measured parameters can be calculated as follows. With no DUT, the through calibration is done and the S_{21} reading is set to 0dB (ratio=1). There is a 2:1 voltage divider formed by the two 50-ohm impedances of the VNA, therefore this corresponds to $V_s=2$, and $V_1=V_2=1$ voltages. We further assume that all DUT impedances are much smaller than the 50-ohm VNA impedance, so that the i_1 input current is simply the shunt current of the source: $i_1=2/50$. The measured $v_2 = Z_{21}*i_1$ output voltage is equal to the S_{21} reading, therefore $Z_{21}=25*S_{21}$. Note that this is the same expression that we had for the two-port self-impedance reading.

Slide#9

S21 Uncertainty

- $|S_{21}|$ uncertainty of HP8720D:

  <1dB in the $|S_{21}|$ > -60dB range
  <3dB in the $|S_{21}|$ > -70dB range

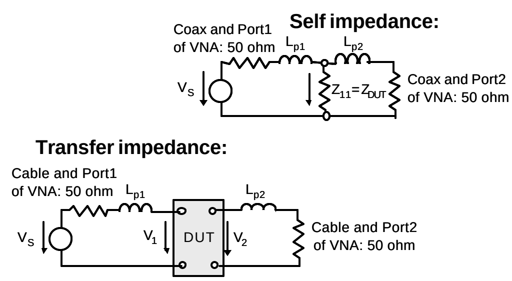
- Impedance uncertainty:

  1dB (10%) for Z_{DUT} > 25milliohms
  3dB (40%) for Z_{DUT} > 8milliohms

The uncertainty of S_{21} reading is a function of frequency and S_{21} magnitude. The uncertainty gets bigger at lower frequencies and lower readings. For the Hewlett Packard 8720D vector-network analyzer, the uncertainty is less than 1dB and 3dB at any frequencies between 50MHz and 20GHz as long as the S_{21} reading is above -60dB and -70dB, respectively. This corresponds to a 1dB and 3dB impedance-reading uncertainty of 25 milliohms or higher and 8 milliohms or higher, respectively.

Slide#10

Equivalent Circuit of Probes Connection



To analyze the error introduced by the discontinuity between the semirigid probe and the DUT, as well as to provide a more accurate transformation from the measured S_{21} readings to the self and transfer impedances, the equivalent circuits shown on the slides were created [6]. The equivalent circuits assume that the interconnection discontinuity is lumped and inductive within our entire frequency range of interest. In general, the discontinuities on the two connections may be different L_{p1} and L_{p2} inductances are assumed on side 1 and side 2, respectively.

Slide#11

S_{21} Conversion to Self Impedance

$$Z_{ii} = S_{21}\frac{Z_1}{2}\frac{1}{1-S_{21}\frac{Z_1+Z_2}{2Z_2}} \approx S_{21}*25*\frac{1+j\omega\tau_p}{1-S_{21}}$$

Where $Z_1 = 50+j\omega L_{p1}$

$Z_2 = 50+j\omega L_{p2}$

$\tau_p = L_p/50$

By solving the Z-parameter circuit equations, the Z_{ii} self-impedance parameter can be expressed in terms of the measured S_{21} transfer ratio. For the self-impedance expression the result depends only on the self impedance of DUT at the measured point, and the transfer-impedance parameters have no effect on this reading. The expression can be simplified if we assume that the two discontinuities are equal: L_{p1}=L_{p2}. This yields the expression on the right, which contains the 25*S_{21} main term, which was used in the approximate calculations, multiplied by an error term. The nominator of the error term describes the frequency dependent error due to the inductive discontinuity. Its corner frequency is determined by the L_p/50 time constant. Assuming a 0.4nH inductance, the corner frequency is around 20GHz. The denominator of the error term depends on the S_{21} reading. This error is smaller when we measure lower values of impedances. When the measured impedance is 0.2 ohms or less, the denominator yields a less then 1% error.

Slide#12

S_{21} Conversion to Transfer Impedance

$$Z_{ji} = S_{21}\frac{Z_1}{2}\frac{\left(1+\frac{Z_{11}}{Z_1}\right)\left(1+\frac{Z_{22}}{Z_2}\right)}{1+\frac{S_{21}Z_{21}}{2}\frac{Z_1}{Z_2}} \approx$$

$$S_{21}*25*\frac{1+j\omega\tau_p}{1+50*\left(\frac{S_{21}}{2}\right)^2}*\left(1+\frac{Z_{11}}{Z_1}\right)\left(1+\frac{Z_{22}}{Z_2}\right)$$

Where $Z_1 = 50+j\omega L_{p1}$

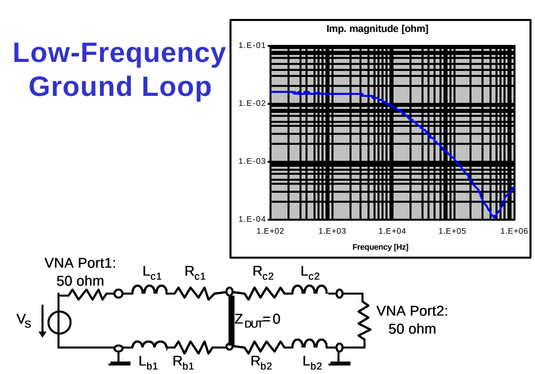
$Z_2 = 50+j\omega L_{p2}$

$\tau_p = L_p/50$

The Z-parameter equations can also be solved to obtain the Z_{ji} transfer-impedance parameter. If we assume that the two discontinuities are equal: L_{p1}=L_{p2}, the expression can be simplified to have the same 25*S_{21} main term, which was used in the approximate calculations, multiplied by an error term. The error term now is more complex, because the transfer parameter depends on the self impedances at both connection points as well. The nominator contains the frequency dependent error due to the inductive discontinuity, the same term we have in the self-impedance expression. The denominator now varies with the square of the S_{21} reading, this means that this error term diminishes even faster as we measure lower values of impedances. The other two multiplicative error terms depend on the self impedances at the two connection nodes, and their error values linearly go down as the self impedances of the DUT gets lower.

Slide#13

Low-Frequency Ground Loop

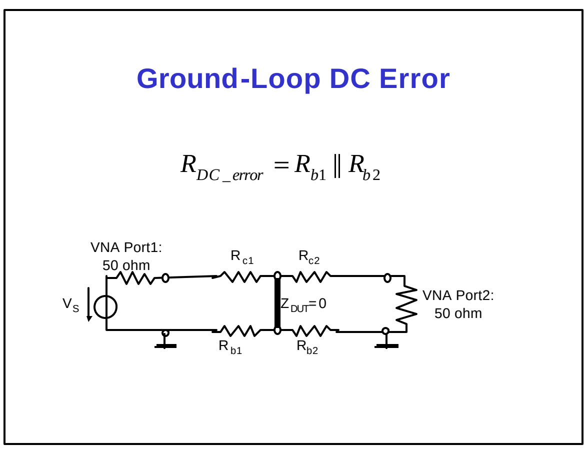


Beyond of the one-port connecting-discontinuity error (that we eliminate by using the two-port measurement setup) there are further limiting factors in measuring low impedances. If we connect two regular coax cables to Port1 and Port2 of the VNA, with cable hot wires shorted to the shield, and the two shorted cable ends connected, there is still a residual reading way above the noise floor of VNA. This is explained by the schematic of Figure 13.

At low frequencies, a ground loop is formed by the internal ground connections of Port1 and Port2 of the VNA and the cable shield impedances. The equivalent schematic of the slide shows the extreme case when ZDUT=0 is assumed.

The top measured chart shows the residual reading with two pieces of 24" RG178U/B coax cables, the probes being soldered to a solid copper sheet. The low-frequency residual reading is 14.2 milliohms.

Slide#14

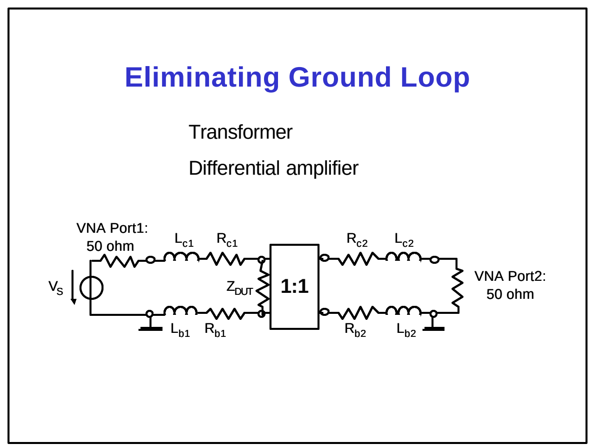


At very low frequencies, where wL<<R for the center wire and braid of the coax cables, the equivalent circuit can be simplified by dropping the inductive terms. Assuming that the resistance of coax center wires are much less that 50 ohms (Rc1<<50 and Rc2<<50), and shorted probe tips (ZDUT=0), what we really measure in this setup is the parallel equivalent of the two braid resistances.

On the previous slide, the residual reading with shorted probe tips after 2 pieces of 24" RG178B/U coax cables was 14.2 milliohms. The Belden RG178B/U 30AWG coax datasheet gives the nominal shield DC resistance as 47.9 ohms/meter, which corresponds to 1.22milliohm/inch. The nominal DC resistance of two 24-inch long braids in parallel is then 14.6 milliohms, which is very close to the measured 14.2 milliohms.

The equivalent circuit of the slide suggests that one way of reducing this error is to shorten one of the two (or both) cables, such reducing Rb. For generic power-distribution-network measurements, however, we usually need the flexibility to reach internal points of larger printed-circuit-boards or systems, therefore we need other means of reducing the ground-loop error.

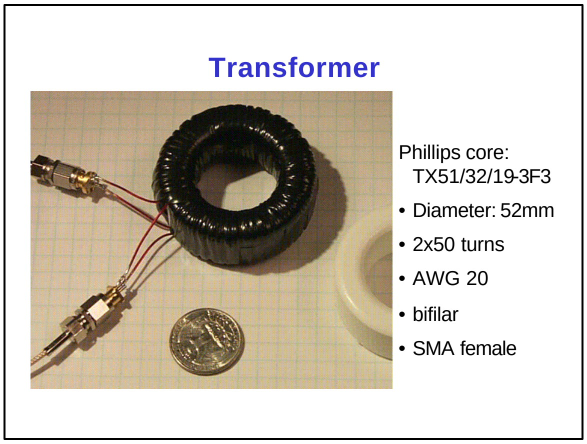
Slide#15



The ground loop formed by the two coax cable braids can be eliminated either by using an isolation transformer, or an isolation amplifier with sufficiently large common-mode rejection.

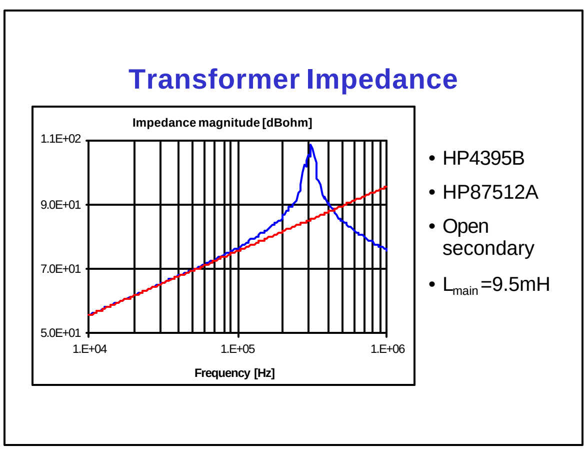
Slide#15 depicts the equivalent circuit with an isolation transformer connected into the receive (Port2) path. The transformer has a 1:1 turns ratio, but as long as it does not restrict the response bandwidth, the turns ratio can be any arbitrary value as well, and it can also be located in the transmit (Port1) path. The transformer isolation makes it possible to use long connecting cables with no worry about the ground loop formed by the braids.

Slide#16



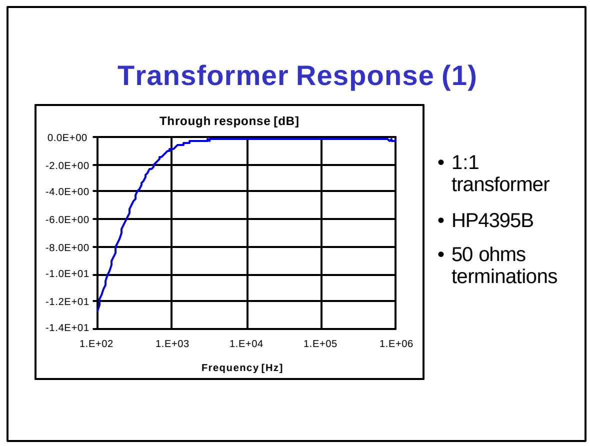
For the measurement results shown later, an isolation transformer of 1:1 turns ratio was used. The ferrite toroid was Philips TX51/32/19-3F3. The 52 mm diameter toroid was chosen to enable easy hand winding and to handle the maximum (+15dBm) output power of the VNA without any worry of running into nonlinearities. The 1:1 isolation was achieved with two times 50 turns of AWG#20 wire. The wires of the primary and secondary were wound together in a bifilar fashion. Two SMA female connectors were soldered to the winding ends, making it easy to use the isolation transformer with SMA cables. The slide shows the completed transformer with the connecting SMA coax cables, together with a spare ferrite toroid ring.

Slide#17



First, a winding of ten turns of AWG#20 wire was created on the toroid ring, and its inductance was measured. From the 0.37mH measured inductance (at 100kHz), the AL inductance constant was found to be 3.75mH/turns2. The slide shows the impedance of the 50+50-turn transformer, measured across the primary windings, with its secondary left open. The measurement was taken with a Hewlett-Packard 4395B VNA. The vertical scale shows the impedance magnitude in dBohms. The thin straight line corresponds to 9.5mH of main inductance, which agrees well with the calculated inductance of L = AL*n2 = 3.75m*502 = 9.375mH. Note the parallel resonance at 316kHz, however the impedance value is so high that below 1MHz we can expect very little influence on the measurement reading.

Slide#18

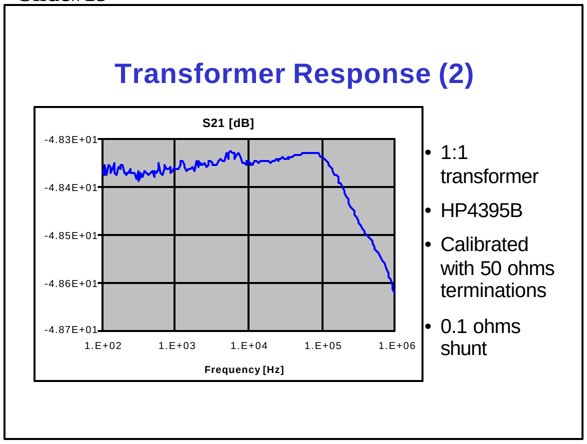


Between 50-ohm terminations, the isolation transformer has a flat response within -3dB from 500Hz to at least 10MHz, with a -13dB response at 100Hz. To exclude a parasitic ripple of about 2 dB at around 3MHz, all later measurements limited the transformer setup to 1MHz.

For calibrated low-frequency measurements, the through-response calibration included the transformer as well.

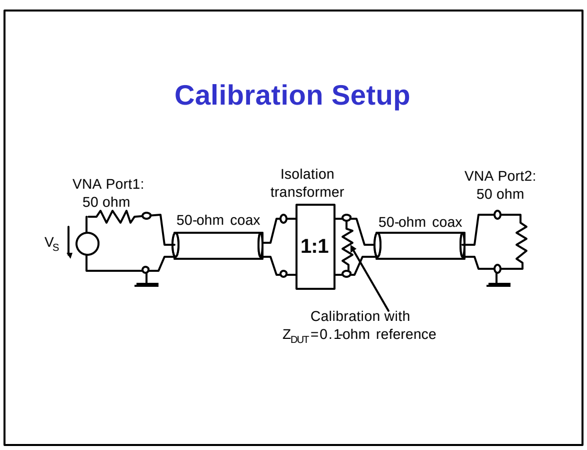
Note that transformer isolation can be used mostly to measure only passive impedances, because the transformer's low DC resistance may draw excessive DC current from an active circuit.

Slide#19



Slide#19 shows the frequency response of the same transformer when the DUT is a 0.1-ohm shunt calibration resistor. For this response measurement, the calibration was done in a 50-ohm through-connect fashion, without the isolation transformer. The 0dB reading corresponds to 25 ohms, the 0.1 ohms DUT nominally has a 20*Log(0.1/25)=–48dB reading. Note that due to the low DUT impedance, the transformer response is flat within 0.1dB even without including the transformer in the calibration loop.

Slide#20

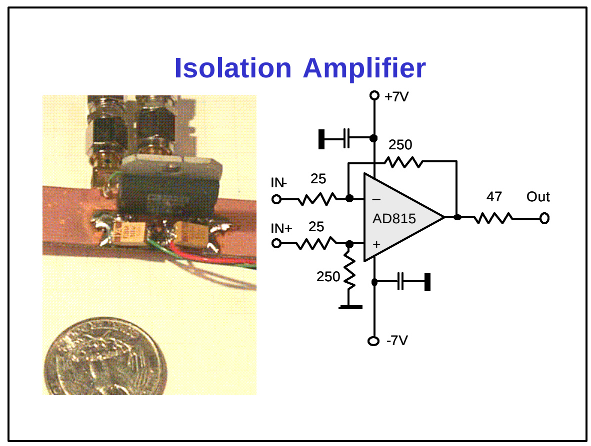


The calibration setup for frequencies below 1MHz is shown in Slide#20. The cables were 24-inch RG178B/U SMA-SMA coax cables, the isolation transformer was the 1:1 toroid ring shown on Slide#16.

The 0.1-ohm calibration reference was made of two SMA female posts and five 0.5-ohm 805-size SMD resistors. The ground and center posts of one of the SMA female connectors were cut away, and the five 805-size resistors were soldered directly between the connector flange and center conductor. The second SMA female connector was then soldered on top of the resistors, creating a female-female SMA 0.1-ohm calibration standard.

Instead of the obvious 50-ohm through-connect calibration, the 0.1-ohm calibration standard was chosen to maximize the avilable dynamic range of the VNA. This way the 0dB full scale reading of the VNA corresponds to 0.1 ohms.

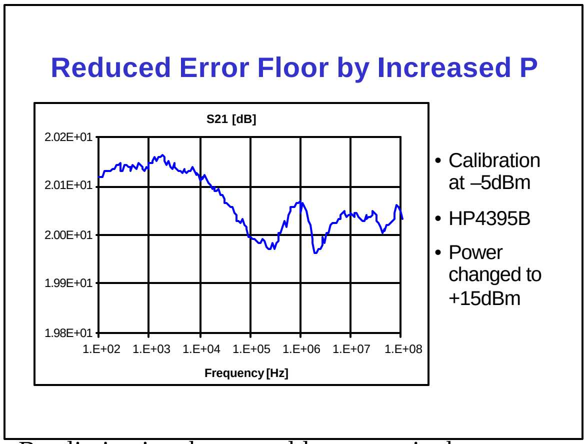
Slide#21



Another possible way of opening up the coax-braid ground loop is to use an isolation amplifier with differential input and/or output with sufficiently high common-mode rejection. Slide#21 depicts the equivalent schematic of an isolation amplifier built around one half of the Analog Devices AD815 dual operational amplifier. The supply voltage was set to +-7V in order to make sure that the output voltage cannot exceed the +-7V maximum input voltage range of the HP4395 VNA.

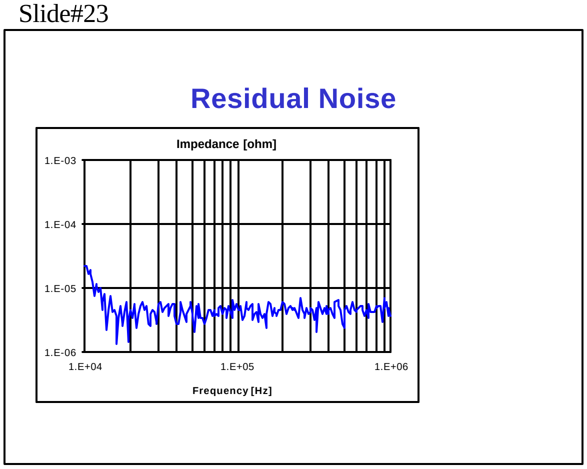
The isolation amplifier had precision surface-mount resistors to create the 50-ohm differential input and 10-times nominal gain. Assuming high loop gain, the output impedance was set by a series 47-ohm resistor. The supply rails were bypassed by three capacitors on each supply voltage: 100uF tantalum, 10uF and 0.1uF ceramic capacitors. To facilitate easy connections, SMA female connectors were soldered to the input and output resistors.

Slide#22



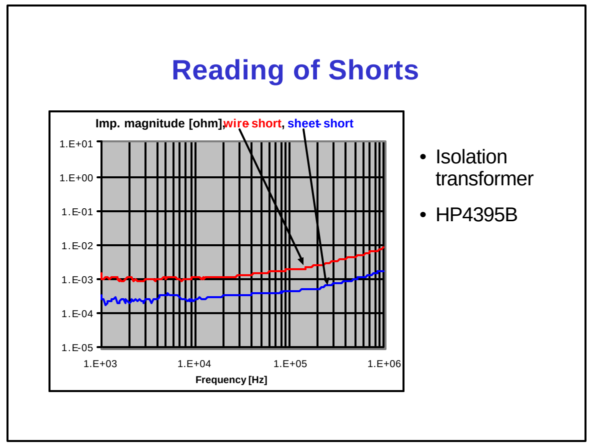
By eliminating the ground-loop error in the two-port impedance-measurement setup, the last remaining limiting factor is the noise floor of the VNA. Knowing that we want to measure low impedance values, one can extend the useful signal range within the given dynamic range of the VNA by calibrating the setup with a given (lower) power, and by increasing the power when the actual measurements are conducted. This way, however, the absolute error and flatness difference of the power-setting circuit is outside of the calibration loop. To illustrate the expected error due to this, Slide#22 shows the reading of a HP4395B VNA with the power having switched to +15dBm after the calibration was done with –5dBm. On the particular instrument, this error was less than +-0.05dB in the 100Hz-100MHz frequency range.

Slide#23



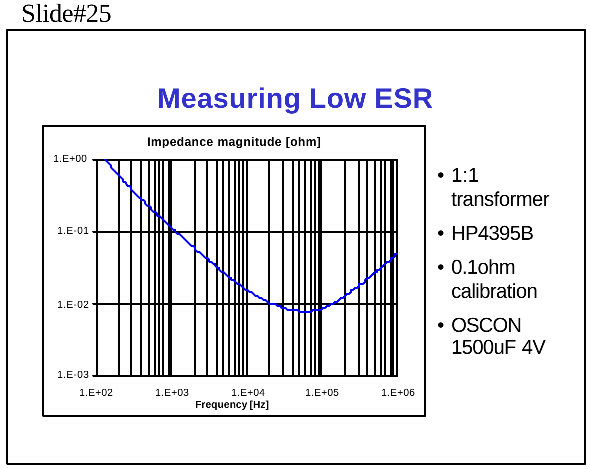
The graph on Slide#23 illustrates the background noise of the two-port impedance measurement setup with a HP4395B VNA, and 1:1 isolation toroid ring transformer, with the power switched to +15dBm after the calibration was done with -5dBm power. The graph is scaled in impedance. The measurement noise floor (with open probes) is below 10 micoohms almost in the entire 10kHz-1MHz frequency range. The noise floor below 10kHz raises to a few hundred microohms.

Slide#24



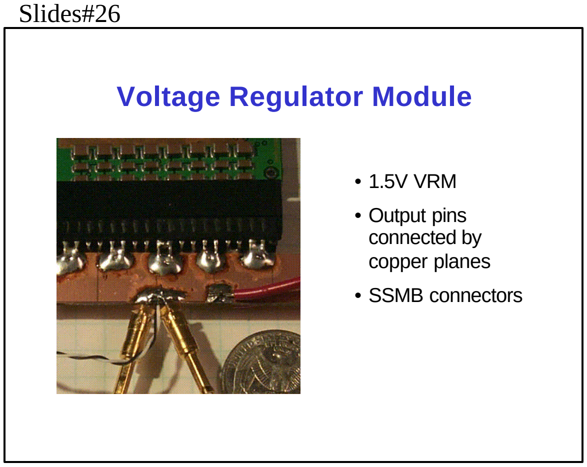
Slide#23 shows the noise floor with the probes open, not connected to DUT. On Slide#24 the reading with different shorts are shown. After calibration with the isolation transformer and 0.1-ohm reference, the probe tips were shorted by a AWG#24 wire (wire short) and by soldering the probe tips to a solid copper plane (sheet short).

Slide#25



The two-port impedance-measurement setup can be conveniently used to characterize bypass capacitors with high capacitance and low ESR values, when the series resonance frequency is low. Single capacitor pieces would not otherwise need the two-port impedance-measurement setup with coaxial connection cables, as they can easily be soldered down to testpads with direct connection to the measuring instruments. However, when the capacitor's ESR value is measured with a VNA in a shunt location along a printed-circuit-board trace, at low frequencies the ground-loop error described on slides 13 and 14 would limit the measurement of ESR to values not lower than about 10-15 milliohms. With the isolation transformer, the 7.8milliohm of the OSCON 1500uF 4V capacitor can be accurately measured at the 60kHz series resonance frequency.

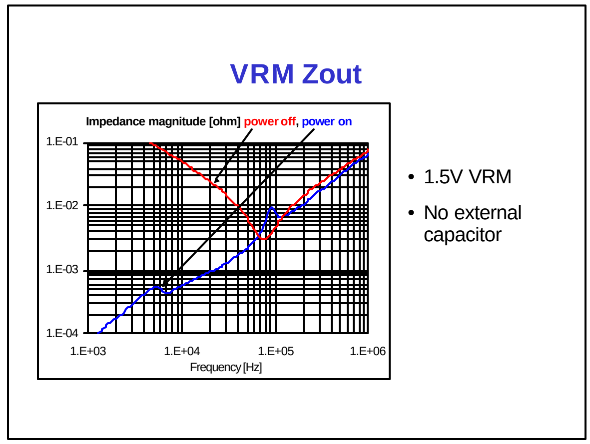
Slides#26



Slide#26 shows the probe connections to a 1.5V VRM made by Lucent [4]. The VRM uses standard card-edge connectors with multiple pins for the output ground and power. To measure the VRM in a stand-alone configuration, all of the corresponding ground and power output pins of the socket were soldered to a thin double-sided printed-circuit-board strip. The uncut, solid copper sheets of the PCB on the top and bottom side carried ground and power, respectively. The two-port impedance-measurement setup was connected through 24-inch coax cables and miniature coaxial connectors. The coax connectors were soldered to the copper sheets within a 50-mil distance.

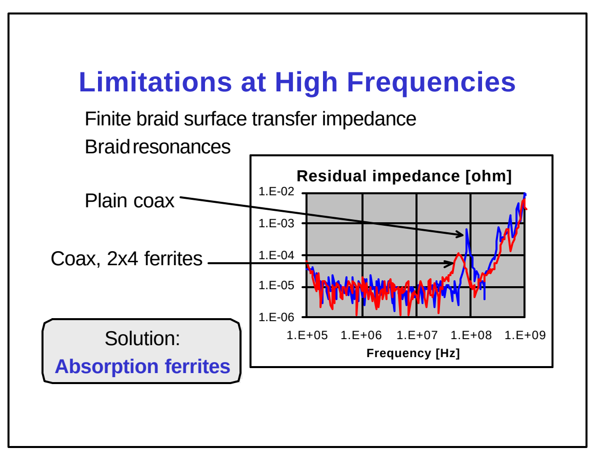
To ensure that we measure the low output impedance enforced by the active regulator loop, the remote-sense wires were soldered to the planes at the same points where the coax probes were connected.

Slide#27



The output impedance of the VRM is dependent on whether it is powered or not. With no input power applied, the reading follows the curve of the impedance of output filter capacitors. With the input power applied, the control loop reduces the low-frequency impedance. As shown in Slide#27, the active loop maintains a less than one milliohm output impedance up to about 20kHz. The measurement on the slide was taken with a 1A DC load current, however it was found that the output impedance showed no noticeable change with various DC loads.

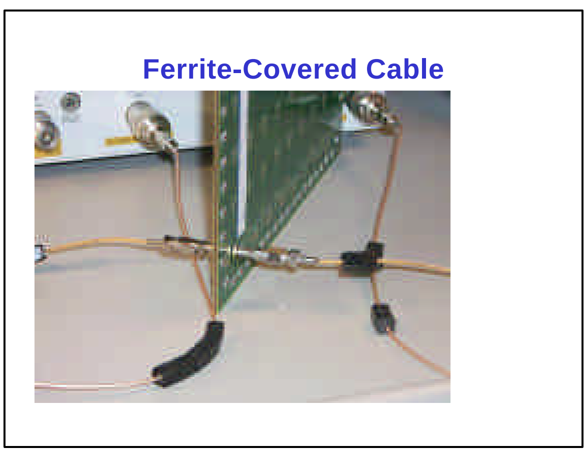
Slide#28



When measuring very low or very high impedances, there are almost full reflections at the ends of connecting cables. The finite surface-transfer impedance creates resonance peaks in the residual reading. The resonance frequency is directly related to the delay of cable. With the usual 12-36-inch cable length, the lowest resonance frequency is in the tens and hundreds of MHz range.

The slide shows the S21 reading converted into impedance values in a setup where two 24” long RG178 cables run loosely in parallel at a distance of one inch. There is no galvanic connection between the probes. The curve with a large peak at 90MHz is with plain coax cables, the trace with a peak at 60MHz is with four pairs of ferrites on each cable. Note that the peak magnitude is reduced by about 13 dB. Further reduction of the peak can be achieved by covering the entire length of coax with absorbing ferrites.

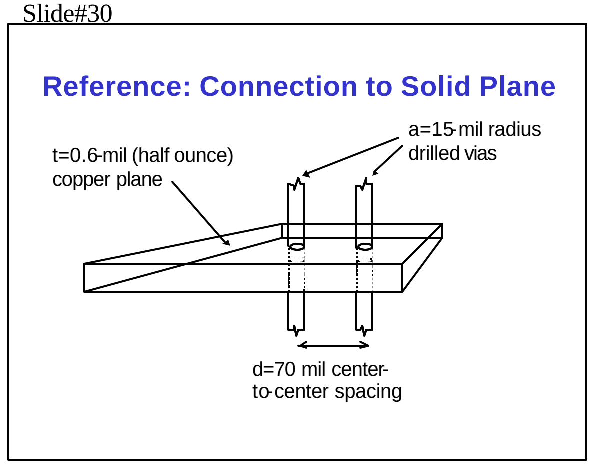
Slide#29



The photo on the slide shows the ferrite-covered RG178B/U cables connecting the HP 4396B VNA to a power-distribution test board. Fair-Rite Round-Cable Ferrite Suppression Cores with material 43 (part# 2643166751) were mounted in pairs around the RG178B/U coax cable.

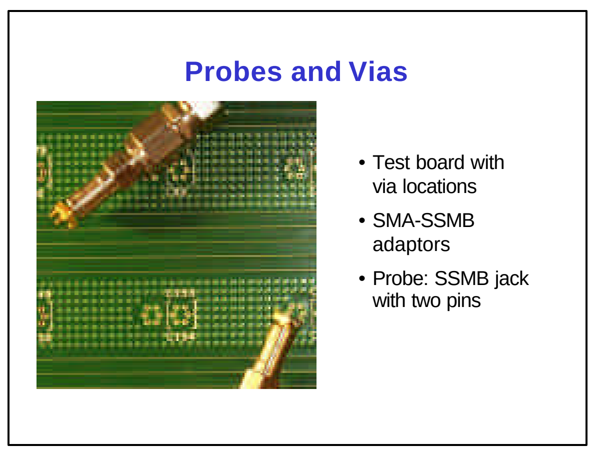
The ferrite suppression parts were taped together to form loosely attached beads.

Slide#30



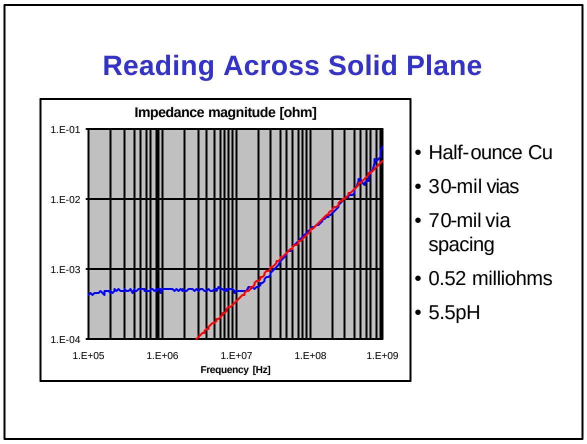
The test board shown in Slide#29 had multilayer structure with solid copper planes on specific layers. At regular intervals, plated through holes connect the planes to the surface. The plated through holes have 30-mil drilled size, without plating. One pair of vias with 70-mil center-to-center separation was selected to perform a high-frequency reference reading of the impedance of the half-ounce copper-plane.

Slide#31



The photo illustrates the location of plated through holes on the test board, together with the miniature coaxial probe connectors. The straight PCB-mount SSMB jacks have four ground posts and one center pin. Three of the four ground posts were cut away, forming a two-pin probe tip. To make the measurements, the probe pins were inserted into the plated through holes.

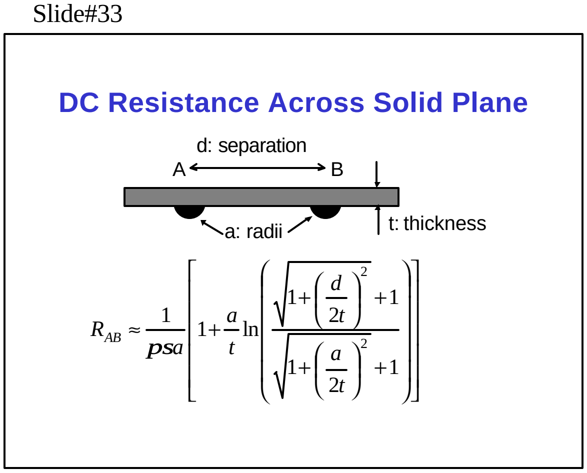
Slide#32



The impedance of the half-ounce copper plane between vias with 70-mil separation is shown on the slide. To avoid coupling through the via-loops, and to connect both probes to the samne points on the plane, the two probes were inserted into the same plated through holes from the opposite sides. This way the current-carrying portions of the via loops were isolated by the copper plane itself. The skin depth in copper at 1GHz is 2 mm, which, together with the 15.2mm thickness of the half-ounce copper plane provides an exp(15.2/2)= 2000 times isolation between the opposite sides of the plane.

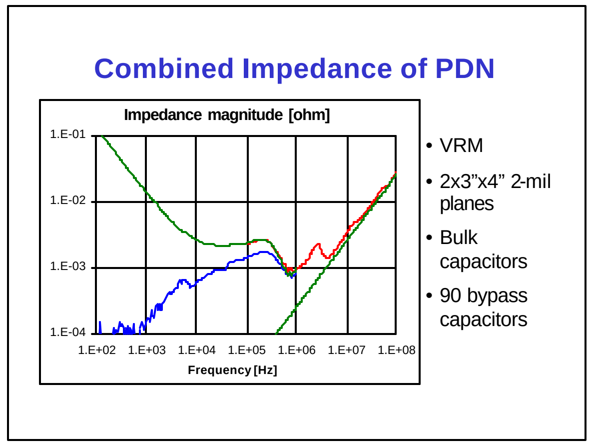
The low-frequency flat reading is 0.52 milliohms. The high-frequency impedance looks like an inductance: the straight matching line is the impedance of a 5.5pH inductance.

Slide#33

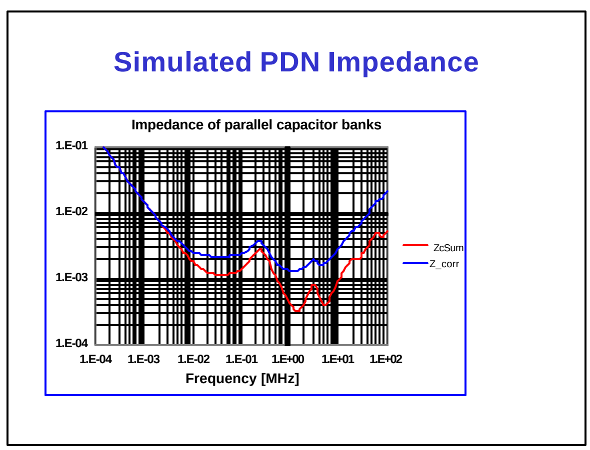


The low-frequency reading can be double-checked against the calculated resistance. Loyka published a simple closed-form expression to calculate the resistance of a finite-thickness plane with given contact radius and separation [5]. By substituting the input numbers shown on Slide#30, the calculated DC resistance is 0.547 milliohm, which is in good agreement with the measured 0.52 milliohm value.

Slide#34



Slide#35



Slide#34 and 35 show the measured and simulated impedance profile of a power-distribution network, respectively. The measured impedance profile is a combined picture taken with different instruments and setups, and contains four different traces. At frequencies between 100Hz and 1MHz, an HP4395 VNA was used with isolation transformer and 0.1-ohm calibration standard.

Below 1MHz, two readings were taken: one with the VRM not powered, and one with the VRM powered. At frequencies between 100kHz and 100MHz, the HP4396 VNA was used with ferrite-covered probe cables and 50-ohm through calibration. The fourth trace is a straight line corresponding to an asymphtotical inductance of 40pH. Note that there is an overlap between the two different measurement setups' results in the frequency range of 100kHz to 1MHz, and the traces show good continuation.

Slide#35 is the simulated impedance profile of the same network without VRM. There are two traces on the graph: one shows only the bypass capacitors with no plane loss assumed (Zcsum), and another one with the plane resistance included (Z_corr). The Z_corr trace shows sufficient agreement with the measured impedance profile.

Slide#36

Recommended Resources

Hewlett Packard Vector Network Analyzers:
- HP 4395 VNA 10Hz-500MHz
- HP 4396 VNA 100kHz-1.8GHz
- HP87512A DC-2GHz Transmission/Reflection Test Set
- HP 8720D VNA 50MHz-20GHz

Circuit simulator software:
- Avant! HSPICE

Slide#37

Conclusions

- Two-port measurements reduce effect of discontinuities
- Limitation at low frequencies: cable-braid ground loop
- Ground-loop is eliminated by transformer or amplifier
- Limitation at high frequencies: braid leakage and resonance
- Reliable reading and good correlation to simulations is achieved in the sub-milliohm range

References:

[1] Larry Smith, Istvan Novak, "Power Distribution Challenges in the New Millennium," Short course at the 1999 Topical Meeting on Electrical Performance of Electronic Packaging, October 25-27, 1999, San Diego, CA

[2] I. Novak, "Probes and Setup for Measuring Power Plane Impedances with Vector Network Analyzer," Proceedings of the 1999 DesignCon High-Performance System Design Conference, Feb. 1-4, 1999, Santa Clara, CA, pp. 201-215.

[3] Understanding the Fundamental Principles of Vector Network Analysis, Hewlett Packard Application Note 1287-1.

[4] Titania Power Modules. Lucent Technologies datasheets, October 1999.

[5] Sergey L. Loyka, "A Simple Formula for the Ground resistance Calculation," IEEE Transactions on EMC, Vol.41, No.2, May 1999, pp. 152-154.

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Current activities

Istvan is signal-integrity senior staff engineer at SUN Microsystems, Inc. Besides of signal-integrity verification of high-speed serial and parallel buses, he is engaged in the design and characterization of power-distribution networks, bypassing and decoupling of printed-circuit boards as well as in the development of future technologies at workgroup servers. He creates and uses SPICE models, and develops measurement techniques for the power-distribution networks.

Istvan's background

Istvan has more than twenty years of experience with high-speed digital, RF, and analog circuit and system design as well as in teaching related subjects at university regular courses and industry short courses. He is Fellow of IEEE for his contributions to the signal-integrity and RF measurement and simulation methodologies. Istvan received his masters degree from the Technical University of Budapest, Budapest, Hungary, and his PhD from the Hungarian Academy of Sciences.