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How To Make Sense Of Sense Resistors

by Dennis Feucht, Innovatia Laboratories, Cayo, Belize

Some components are relatively unimportant in their effect on overall circuit behavior. Sense resistors are not among them. They might only be resistors, but they can be important resistors with subtleties that if missed can stall a project. This article is a mini-tutorial on fundamental concepts for power-circuit design involving sense resistors.

This article discusses a variety of problems that must be understood to obtain accurate measurements with sense resistors including temperature effects, properties of resistor materials, make-versus-buy options, and Kelvin sensing. The explanations on those topics set the stage for the more-detailed discussions that follow on how to measure resistor parasitics.

Two methods for measuring the parasitic series inductance of sensor resistors on the bench—frequency sweep and pulse response methods—are explained to enable designers to properly evaluate and characterize sense resistors for use in power supply circuits. Test circuits and set-ups are described and illustrated. These methods are highly accessible in that they employ an oscilloscope, sine and square-wave generators, and basic probing accessories, rather than turning to the more-expensive network analyzers.

Temperature Effects

Changes in temperature cause changes in resistor values as expressed quantitatively by the *temperature coefficient*,

$$TCR = \frac{dR}{dt} \approx \frac{\Delta R}{\Delta T}\text{, in } \Omega/\text{K}$$

or in fractional form,

$$TCR\% = \frac{dR/R}{dT} \text{ in \%/K}$$

Ideally, *TC* and *TC%* are zero, and one of the goals of resistor manufacturers is to make it so, and for circuit designers, to either find low-TC resistors or ways of compensating for temperature effects.

Temperature Coefficient

Sense resistors in power circuits are usually power resistors; they dissipate significant power and are typically rated at 0.5 W to 5 W. While dissipating their full-scale power, they must hold their resistance value, for it contributes proportionally to the current measurement. *TCR* is consequently of high design importance. Using a clipped length of copper wire for a sense resistor can demonstrate this. In a motor drive, the torque will fall off as the motor operates because of the increasing resistance of the copper sense resistor, as it heats.

However, sometimes a copper sense resistor—as either a discrete wire resistor or a circuit-board trace—can be used to advantage. The TC of copper is nearly constant over a wide temperature range, and its positive TC of about +0.4%/°C can compensate for the characteristically negative TC (−2 mV/°C) of a semiconductor junction elsewhere in the current-control loop. Astute engineers do not forget the design adage: "If you can't fix it, feature it."

Actual Sense Resistors

One simple means of choosing sense resistors is to make your own out of a low-TC conductive material. This can be an in-house task because it requires no special equipment or skills and reduces resistor cost. Manganin wire is an alloy of copper, manganese, and nickel, and has a low *TCR%* of within 15 μ/K = 15 ppm/°C from 0 to 80°C.

Manganin wire of size 18 AWG has a resistivity of 0.361 Ω/m. Smaller diameters are available and the wire can be bought by the roll. Cut the wire to length for the desired resistance, tin the ends, and solder into the board. A manganin resistor is shown below (left) in Fig. 1. By keeping the half-loop area small, inductance is minimized. The one shown on the left in Fig. 1 has a resistance of 25 mΩ.

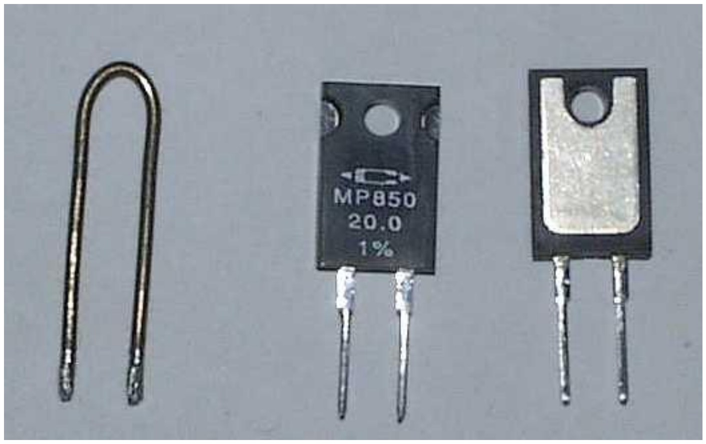

Fig. 1. Manganin resistor of 25 mΩ on left and 1% commercial resistor in power package, center and right.

If you do not make your own sense resistors, those shown in Fig. 1 (right, front and rear) are commercially available power resistors made of a low-TC metal foil on an anodized aluminum substrate, and in TO-220 or TO-247 packages. Several companies, such as Caddock Electronics, now supply these accurate low-TC power resistors at an attractive price for power-circuit design.

Another commonly-used low-TC material is nickel-chromium, or *nichrome*. It's resistivity of 133 μΩ-cm requires less wire length than mangagin, with a resistivity of 43 μΩ-cm, which can reduce inductance for very low-value resistors. Manganin, however, is superior to nichrome in TC and long-term stability of resistance value. If future circuit-board fabrication technology presents a wider range of substrate materials than copper, thin-film power resistors can be integrated onto the board during layout. With clever circuit design, even copper traces can be TC-compensated with BJT junctions.

Kelvin Sensing Resistors

Series parasitic resistance is an undesirable design factor for low-value resistors. It is necessary to sense across the actual resistance of specified value and not some of the component lead resistance in addition. To solve this problem, four-wire, or *Kelvin*, sensing is made possible with four-wire resistors, such as those shown in Fig. 2. The leads are attached internally to the desired resistance and are brought out of the package in pairs.

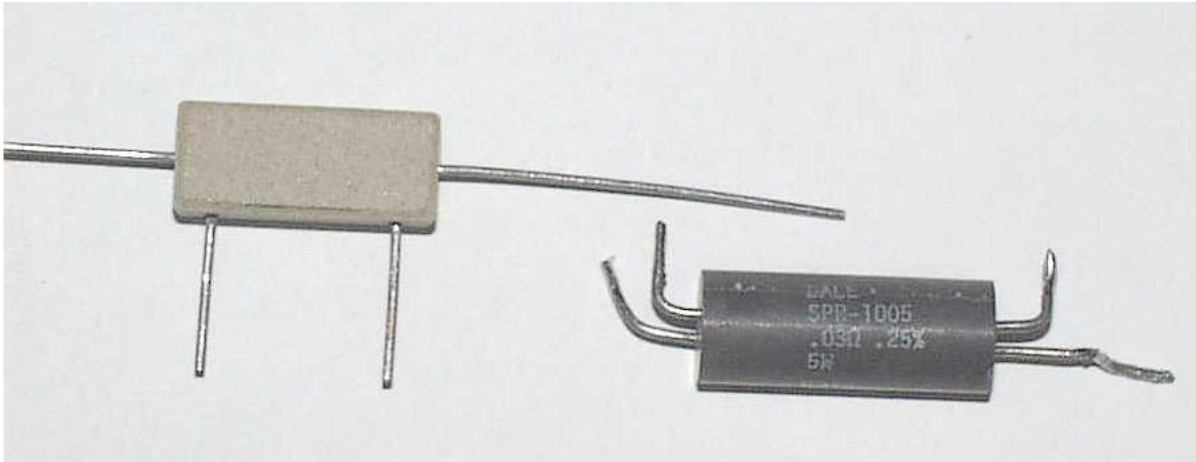

Fig. 2. Four-wire or Kelvin sense resistors. The low resistance values cause parasitic lead resistance to degrade accuracy and require that separate drive and sense terminals be brought out of the package.

Kelvin-sensing resistors are also made smaller, in foil-on-substrate form, as shown in Fig. 3. (The quarter is for size comparison.) The smaller, inner leads are four-wire Kelvin sense leads while the wider, outer leads are the drive leads.

Fig. 3. Small metal-film resistors on a ceramic substrate (with a quarter shown for size). The outer terminals are the drive pair and the smaller inner terminals are for sensing voltage across the accurate 1% resistance.

To summarize, current-sense resistors for power circuits usually must have a low TC, relatively high power rating, and be accurate at low resistance values. Such resistors are commercially available in multiple package choices and at moderate prices.

If prices are too high for a given application, instead of buying them, consider making sense resistors out of manganin wire, or for the lowest cost, use a copper circuit-board trace compensated by scaling the TC of a silicon junction. Thermal calculation is required to determine its thermal resistance and how wide and long to make the trace, for it directly affects the *TCR*.

Parasitic Series Inductance—Frequency Sweep Measurement

Sense resistors are usually within a feedback control loop with its own dynamics. An adequate sense-resistor model includes parasitic series inductance. The terminal leads (or terminal traces for surface-mount resistors) contribute an inductive element. In non-power circuits, this inductance is often of no consequence because it forms a time constant that is very small. When resistance is also small in value, the time constant, $\tau = L/R$, becomes large enough so that the corresponding frequency, $1/\tau$ decreases to within a decade of the loop bandwidth of the power circuit and affects the loop dynamics.

It is not difficult to encounter parasitic inductances in power sense resistors in the 50- to 200-nH range. This is too small in value to measure accurately on common impedance or RLC meters, but can be measured conveniently on the lab bench using an oscilloscope. A 100-mΩ sense resistor with 100-nH series inductance has a time constant of 1 µs. A 10-mΩ resistor will have a time constant of 10 µs, or a break frequency of about 16 kHz, within the bandwidth of many power-circuit feedback loops. The series RL combination has an impedance of

$$Z_s = s \cdot L_s + R_s = R_s \cdot \left(s \cdot \frac{L_s}{R_s} + 1 \right)$$

where $s = j \cdot \omega$ for steady-state frequency-response analysis (and $\omega = 2 \cdot \pi \cdot f$, where f is frequency in Hertz). The additional zero at $\omega_z = 1/(L_s/R_s)$ introduces another pole in a current-amplifier closed-loop response if it is in the feedback path. Consequently, the zero cannot be ignored and some estimate of the parasitic inductance becomes worthwhile.

In the following sections, we will examine two possible methods for series-inductance measurement, based on the frequency and time domains, in that order.

Parasitic Inductance Measurement: Frequency Sweep Method

The test circuit shown in Fig. 4 can be easily built on a lab bench and used to measure the parasitic series *L*. Typical sine-wave or function generators have 50-Ω outputs and can be used as the waveform source for this measurement.

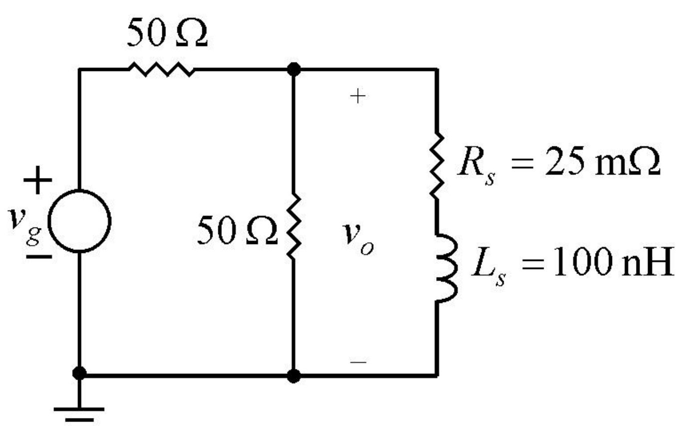

Fig. 4. Sense-resistor test circuit for measuring parasitic series inductance, L_s in a sense resistor of resistance, R_s.

The driving voltage source, v_g has 50-Ω output resistance, and its terminating resistance is 50 Ω. They form a combined series resistance in parallel of R_g = 25 Ω. This Thevenin equivalent source can be transformed into a Norton current source of v_g/R_g in parallel with Norton resistance, R_g. The transfer function of this circuit is

$$Z_o(s) = \frac{v_o(s)}{v_g(s)/R_g} = \frac{v_o(s)}{i_g(s)} = R_g \parallel (R_s + s \cdot L_s) = (R_s \parallel R_g) \cdot \frac{s \cdot \frac{L_s}{R_s} + 1}{s \cdot \frac{L_s}{R_s + R_g} + 1}$$

In steady-state, $s = j \cdot f$ (scaled for f instead of ω), and the frequency-response plot for the magnitude is shown in Fig. 5.

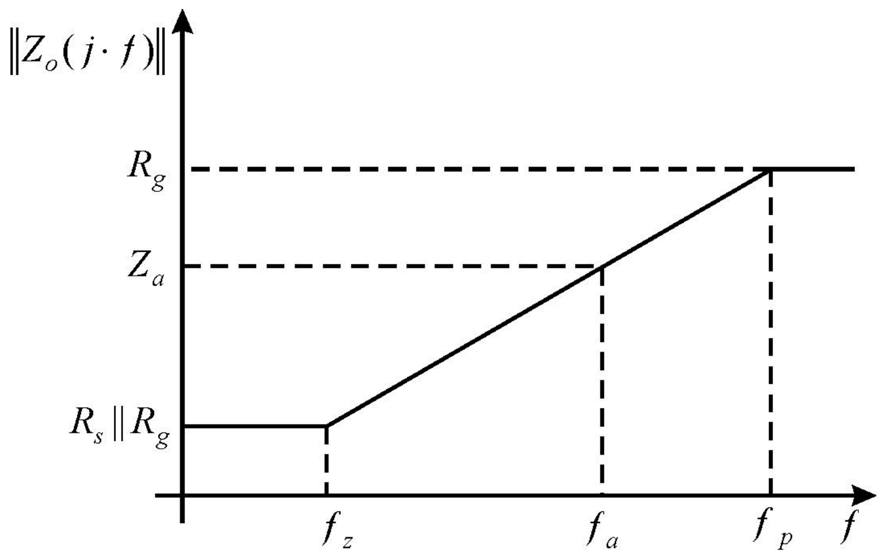

Fig. 5. Magnitude plot of the Fig. 4 test circuit. At 0+ Hz, resistance across v_o is $Z_o \approx R_s \,||\, R_g \approx R_s$ and at high frequency ($f \to \infty$), $Z_o = R_g$. At a test frequency, f_a placed between the zero at f_z and pole at f_p, the impedance is Z_a.

For typical sense resistors, $R_g >> R_s$. Then

$$Z_o(s) = \frac{v_o}{i_g} \cong R_s \cdot \left(s \cdot \frac{L_s}{R_s} + 1 \right), R_g >> R_s$$

By frequency-sweeping the generator, the frequency, f_a at which the amplitude increases over the low-frequency or *quasistatic* (0+ Hz) value of R_s by a times, to $Z_a = a \cdot R_s$, is substituted into $Z_o(s)$ and solved for

$$L_s = \frac{(a-1) \cdot R_s}{2 \cdot \pi \cdot f_a}$$

A value of $a >> 1$ avoids the rounded frequency-response curve near the break frequency and is close to the break frequency asymptote. For about a half decade, $a = 3$. Sweep the generator upward in frequency until the measured amplitude is 3 times that of its unchanging, low-frequency value. Then substitute generator frequency, f_a into the above equation, and

$$L_s = \frac{R_s}{\pi \cdot f_a}, \; Z_a = 3 \cdot R_s$$

Several factors which limit the usefulness of this method are

1. Resistance (besides reactance) increases with frequency from the *skin effect*, the self-induced eddy-current effect within a conductor. Above a frequency at which the effective depth of current penetration into the conductive material is reduced, resistance increases. Thin-film resistors manifest skin effect at higher frequencies than bulk resistors because skin depth exceeds their thin conductive dimension at lower frequencies, causing no change in their effective resistance.

2. The source cable, if not terminated into 50 Ω, will set up standing waves which degrade amplitude measurement accuracy. If possible, set up the test at the generator output terminals, as shown in Fig. 6, but preferably not with a ×10 scope probe, as shown in Fig. 6. The better scheme is to use a 50-Ω-terminated cable, as shown in Fig. 7. A better circuit construction, especially for chip resistors, is to use a GR (General Radio, now IET Labs) line insertion unit, to preserve the 50-Ω cable impedance throughout the test circuit.

Fig. 6. The Fig. 4 test circuit constructed at the output of a function generator. The parallel half-loop of wire is the sense resistor. Although a ×10 probe is sensing v_o, it is better to use a cable terminated in 50 Ω at the oscilloscope. For this circuit, R_g = 50 Ω, not 25 Ω, because it lacks a terminating 50-Ω resistance.

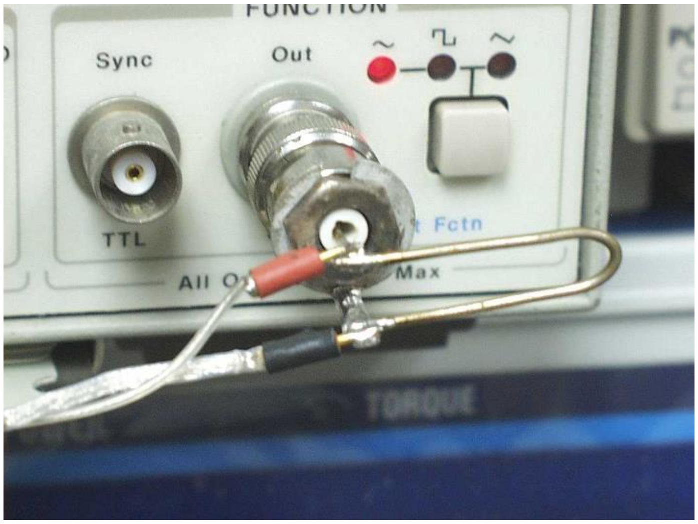

Fig. 7. Preferred method for connecting the oscilloscope input to a sense resistor, with a 50-Ω cable terminated in 50 Ω at the scope. For a higher-frequency connection, trim the cable test leads back to where the cable is coaxial.

3. The oscilloscope "probe" should be a 50-Ω coaxial cable terminated at the scope end in 50 Ω as in channel 1 and not as shown as connected to channel 2 in Fig. 8. The generator must be terminated in its source resistance at the end of a cable with the same characteristic impedance (50 Ω). This termination should be at the oscilloscope input. This results in an equivalent source resistance of 25 Ω, which is still 1000 times that of the 25-mΩ sense resistor.

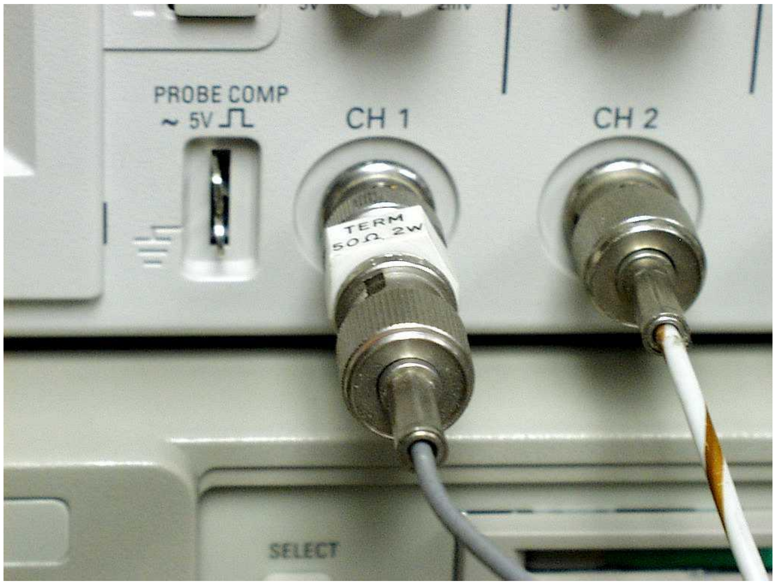

Fig. 8. The test circuit at the scope end is terminated in 50 Ω by a BNC barrel terminator, on channel 1. Channel 2 shows how not to do it.

Despite these measurement precautions (which are generally worth knowing even for power circuits), this frequency-response method is of limited applicability. The value of L_s in the above formula depends on R_s, which increases with frequency once the skin depth is less than the conductor radius or thickness. For 18-gauge wire, the resistance is already around several times the low-frequency value at 100 kHz. For thin-film resistors, which are becoming more common, the technique has some merit, but is still affected by eddy-current variation in R_s.

A test setup with the 25-mΩ Manganin resistor (in Figs. 6 and 7) resulted in a ×3 increase in amplitude at 550 kHz. That calculates to be a value of L_s of 15 nH, which is believable as I expected a value of around 20 nH based on geometry. Yet even in power-circuit design, for what seems as far-removed as resistors, it is not entirely possible to avoid magnetics!

To the relief of some circuit designers, the influence of magnetics effects is not much. By taking the skin effect into account, this sense-resistor modeling method can be applied. The extent of skin-effect is quantified by the *penetration ratio*, ξ. For round copper wire of conductor radius r_c, the skin depth is δ and the penetration ratio, which can be thought of as the wire conductive radius in units of skin depth, is

$$\xi_r = \frac{r_c}{\delta} \approx \left(\frac{r_c}{73.5\,\text{mm}}\right) \cdot \sqrt{\frac{f}{\text{Hz}}}$$

In a slightly different form that introduces the frequency, f_δ at which $r_c = \delta$ instead,

$$\xi_r = \sqrt{\frac{f}{f_\delta}}\ ,\ f_\delta = \left(\frac{73.5\,\text{mm}}{r_c}\right)^2 \cdot \text{Hz}\ ,\ \text{Cu, 80 °C, round wire}$$

In this form, f_δ can be added as a column in a wire table and ξ_r calculated from the above equation. With ξ_r, skin-effect compensation is calculated as the frequency-dependent resistance factor,

$$F_R = \frac{R(f)}{R_0} \approx \frac{\xi_r}{2},\ \xi_r >> 1$$

where R_0 is the quasistatic resistance, as measured with a DMM. For some wire gages that apply for power sense resistors, f_δ is given in the following table.

Table. Frequency (f_δ) at which wire conductor radius (r_c) equals skin depth (δ) for different wire gauges.

r_c in AWG	f_δ, kHz, Cu, 80°C
10	3.226
11	4.071
12	5.132
13	6.453
14	8.153
15	10.278
16	12.985
17	16.340
18	20.608
19	25.980
20	32.629
21	41.225
22	52.103

At f = 550 kHz for the 18 AWG resistor, ξ_r = 5.166 and F_R = 2.58. Then

$R_s = F_R \cdot R_0 \approx (2.58) \cdot (25\ \text{m}\Omega) = 64.6\ \text{m}\Omega$

This is the correct value for R_s(500 kHz) and affects the value of L_s which is based on a constant $R_s = R_0$. With $F_R >> 1$, the frequency-domain method of finding L_s is either more complicated or limited.

Parasitic Series Inductance—Pulse Response Measurement

A better series-inductance measurement technique uses a pulse generator. In the time domain, R_s can be eliminated from its effect on the L_s measurement, and this is a big advantage over the previous method. Using the same setup as before except with a pulse generator (PG) as the source, the rising edge of the generator voltage pulse is adjusted to have a rate of 10 V in 50 ns, or 200 V/µs. Then the source, which is still considered to be a current source (because $R_g >> R_s$), drives the sense resistor with a current ramp of

$$\frac{\left(\frac{dv_g}{dt}\right)}{R_g} \approx (200\ \text{V}/\mu\text{s})/(50\ \Omega) = 4\ \text{A}/\mu\text{s}$$

The PG source resistance of 50 Ω is in series with both the 50-Ω terminating resistance on the scope and the much smaller Z_s being measured. The 50-Ω termination is approximately an open circuit. The inductance is found from the v-i relationship for inductance;

$$L_s = \frac{v_L}{\left(\frac{di}{dt}\right)}$$

where v_L is the constant inductor voltage driven by the current ramp. When driven by a Tektronix PG508 pulse generator, the waveforms of Fig. 9 were observed. Trace A is the open-source voltage without sense resistor; trace 2 is the response with resistor, though only approximately time-aligned with the (stored) A waveform. (Note the × probe attenuation setting for channel 2 because it has no probe and is terminated in 50 Ω.)

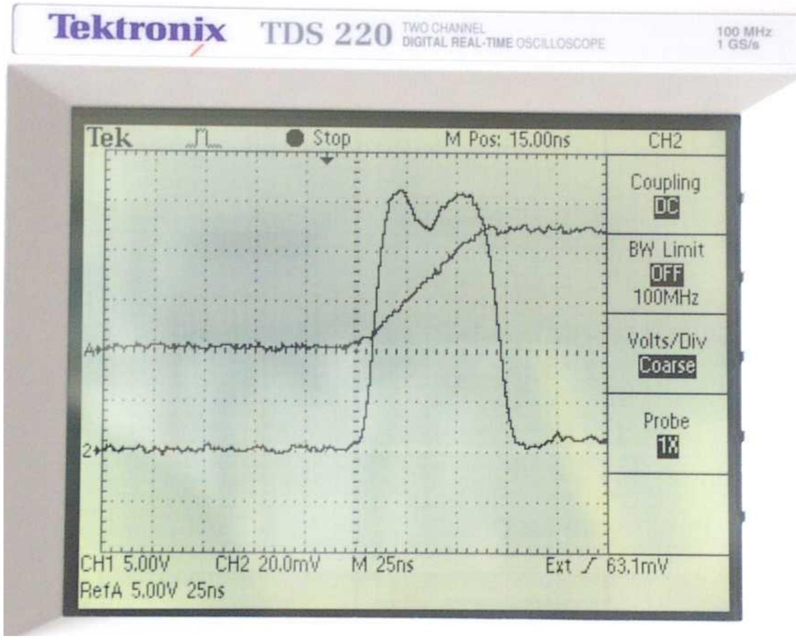

Fig. 9. Trace A is the voltage waveform applied to the test circuit from a 50-Ω source. Trace 2 is the voltage across the sense resistor series inductance, proportional to the derivative of the current.

The voltage, v_o steps up to v_L, an ideally constant value which averages about 100 mV on ch 2, at 20 mV/div. The inductance is calculated from the measured voltage to be

$$L_S = \frac{100\,\text{mV}}{4\,\text{A}/\mu\text{s}} = 25\,\text{nH}$$

This is a reasonable value based on geometry and is somewhat higher than that of the frequency-domain method.

The current ramp also drops voltage across R_s, causing a ramp to be superimposed on v_L. However, it is not observable in the waveform because the inductance voltage is much greater. The source pulse (trace A) flattens after the ramp to a constant voltage, leaving a constant-current drive of R_s that on ch 2 shows about 6 mV (≈ ¼ div) rise above the level to the left of the inductor pulse, where the source current is zero. The value of R_s is calculated from the voltage-divider formula, now including the terminating 50 Ω which forms a ×½ divider with the PG source resistance of 50 Ω. V_g (= 12.5 V) drives R_g = 25 Ω in series with R_s to produce about 6 mV across it. Then

$$\frac{V_S}{V_g} = \frac{1}{2} \cdot \frac{R_S}{R_S + R_g} \Rightarrow R_s = \frac{R_g}{\frac{V_g}{2 \cdot V_s} - 1} = \frac{25\,\Omega}{\frac{12.5\,\text{V}}{2 \cdot (6\,\text{mV})} - 1} = 24.0\,\text{m}\Omega$$

For this measurement, the resulting value of R_s is 24 mΩ, –4% that of the 25 mΩ measured with an RLC meter.

Accurate measurement of the series inductance of small-value resistors is somewhat difficult. An approximate value can be measured, however, using a simple test setup, with a pulse generator and an oscilloscope. This results in the small values not measurable by most RLC meters. A network analyzer can provide more accurate measurements at a much higher price.

About The Author

Dennis Feucht has been involved in power electronics for 35 years, designing motor-drives and power converters. He has an instrument background from Tektronix, where he designed test and measurement equipment and did research in Tek Labs. He has lately been working on projects in theoretical magnetics and power converter research.

For more information on current sensing, see How2Power's Design Guide, locate the "Design area" category and select "Test and Measurement". Also try a keyword search on "current sensing".