# CURRENT SHUNTS

Electric current can be measured directly, through the physical effects of the current itself, or indirectly, by measuring the resultant voltage as the current flows through a suitable, well-defined impedance, usually called a current shunt.

Direct techniques are exemplified by the torque or force produced on a current carrying coil in a magnetic field, such as in classical D'Arsonval meters and electrodynamometers. Other approaches have used the force on an "iron vane" produced by a current carrying coil, the expansion of a "hot wire" carrying a current to move an indicator needle, and the heating of a thermocouple junction to produce a voltage. The latter case is arguably an indirect measurement, illustrating the somewhat arbitrary nature of the distinction.

Even electrochemical methods have been used to measure current, such as the transfer of mercury across an electrolytic gap in a glass capillary tube, or the mass of a deposited metal in an electrolytic cell after a fixed time interval. These might more accurately be considered amp–hour meters, as they measure the time integral of current.

Indirect techniques are typified by the use a of low-value resistive "shunt" to convert current into a corresponding voltage. Although resistances of any value might be used to measure small currents, shunt resistances are typically in the milliohm to microohm range for currents above one amp. The use of current shunts is increasingly common with the advent of modern Digital Volt Meters (DVM), which are essentially voltage-measuring devices.

## CURRENT SHUNT DEVELOPMENT

Early moving coil meters were capable of measuring currents from microamps to several amps. For higher currents, a low-value "shunting" resistance was placed in parallel with the meter, bypassing a large and well-defined fraction of the current around the meter and increasing the measurable current. For large currents, the shunt would be mounted outside the meter housing for better cooling and to allow the shunt and meter to be physically separated for convenience.

It became relatively standard to design panel meters and shunts to operate with 50 mV at full meter scale, while precision "laboratory" grade shunts were more often 100 mV full scale. This standardization allowed the same meter movement to be used for nearly any current, requiring only a suitable shunt resistance and meter scale. Modern shunts for DVMs are usually made in values divisible by a power of ten, allowing current to be read directly from voltage simply by shifting the decimal point; for example, the conductance in a 100 $\mu\Omega$ shunt is 10 A/mV, or 10,000 A/V.

Current shunts may be used for both dc and ac currents, although the design of shunts for high-frequency (kHz to MHz) ac currents is more involved than for dc or low-frequency ac current.

## DC CURRENT SHUNT DESIGN CONSIDERATIONS

The principal considerations in dc and low-frequency ac shunts are accuracy and stability, which are more difficult to achieve with low resistances. The power dissipation in high-current shunts is also a complication, as shall be seen.

### Contact Resistances

Contact resistance is a major factor in defining and accurately measuring resistances below 1 Ω. Current shunts are usually designed with two connections for current, and two additional connections for sensing shunt voltage (or "potential"), as

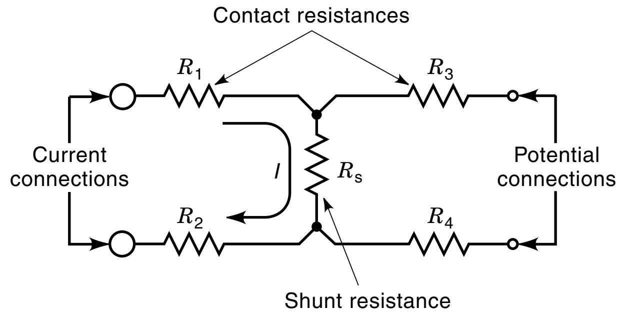


**Figure 1.** Separate shunt current and potential connections are used to minimize errors due to contact resistances.

shown in Fig. 1, often termed a four-terminal "Kelvin" connection. Measurement errors due to the indefinite current contact resistances ($R_1$ and $R_2$ in Fig. 1) are eliminated, as only the voltage across the shunt resistance $R_s$ is sensed. Voltage-sense contact resistances $R_3$ and $R_4$ form a resistive divider with the voltmeter, but these resistances are usually negligible, compared with the input resistance of a voltmeter, and are thus not a significant source of error.

### Power Dissipation Effects

Although the voltage drop is relatively low on a current shunt, the power dissipation can be quite significant; for example, a 1000 A, 100 mV shunt will dissipate 100 W at the rated current. The resulting temperature rise can cause two sources of error: a change in shunt resistance, and a thermal electromotive force (emf), or voltage where dissimilar metals join (a thermocouple effect).

A net thermal emf can occur when the two ends of the shunt are not at the same temperature, resulting in a shunt voltage measurement error. A thermal emf may add to or subtract from the shunt resistive voltage, depending on the algebraic sign of the thermal emf relative to the resistive voltage with a dc current.

These errors are much less when measuring ac currents, as the thermal emf adds to the resistive voltage during one half of the ac cycle, and subtracts during the other half cycle. The result is a complete cancellation of the thermal emf with an average responding ac voltmeter. If the dc component is not blocked to an rms responding voltmeter, the measurement error is still only 1% with a 10% thermal emf, and a 0.01% error with a 1% thermal emf component.

### Selecting a Shunt Material

Pure metals have the low resistivity desirable for shunts, but the temperature coefficients of resistance (TC, or TCR) are high, ranging from about 3 m$\Omega$/($\Omega\cdot$°C) for platinum to 6 m$\Omega$/($\Omega\cdot$°C) for nickel (0.3%/°C and 0.6%/°C, respectively). Alloys of metals typically have higher resistivities but lower TCRs—occasionally much lower. The thermal emf can range from less than one to tens of $\mu$V/°C. The approximate electrical properties of selected metals and alloys are given in Table 1; resistivity and TCR may vary significantly with impurities, alloy composition and degree of work hardening or annealing.

Nichrome is a common resistance alloy with high temperature capability and good corrosion resistance, but the TCR and thermal emf are both relatively high (see Table 1). Nichrome is also physically hard and difficult to work, and cannot be soft (tin/lead) soldered, but must be welded or hard (silver) soldered. Constantan has a much lower TCR than Nichrome, but the thermal emf effect is even stronger.

In 1889, Edward Weston discovered that alloys of copper, manganese, and nickel have very low temperature coefficients. Alloys with 10% to 13% manganese and 2% to 4% nickel are trade-named Manganin. Still the principal choice for shunts, Manganin is available from Harrison Alloys, Inc. in Harrison, New Jersey (800) 526-1256 and (201) 483-4800, and Carpenter Technologies in Reading, Pennsylvania (800) 694-6543 and (610) 208-2000. Typical changes in resistance with temperature for Manganin are shown in Fig. 2. A low positive TCR occurs at low temperatures, with a "turnover" to a negative TCR at elevated temperatures. Operation near

**Table 1. Selected Properties of Electrical Metals and Alloys**

| Metal | Typical Composition | Resistivity $\mu\Omega\cdot$cm (20 °C) | TCR ppm/°C | Thermal emf[a] $\mu$V/°C | Ref. |
|---|---|---|---|---|---|
| Copper | Cu100 | 1.7 | 3,900 | +2.86 | 1 |
| Silver | Ag100 | 1.6 | 3,800 | +2.73 | 1 |
| Aluminum | Al100 | 2.8 | 3,900 | −0.42 | 1 |
| Tin | Sn100 | 11.5 | 4,200 | +0.20 | 1 |
| Lead | Pb100 | 22 | 4,300 | 0.0 | 1 |
| Solder | Sn63, Pb37 | 14.7 | | | 2 |
| Brass | Cu60, Zn40 | 6.4 | 1,000 | +0.8 | 1 |
| | Cu70, Zn30 | 8.4 | 2,000 | | 1 |
| Nichrome | Ni58.5, Fe22.5 Cr16, Mn3 | 100 | 400 | +21.3 | 1 |
| Constantan | Cu60, Ni40 | 44.1 | +2/+33 | −39.9 | 1 |
| Manganin | Cu84, Mn12, Ni4 | 45 | +6/−42[b] | +1.38 | 1 |
| | Cu83, Mn13, Ni4 | 48.2 | +15/−15[c] | | 2 |
| | Cu86, Mn10, Ni4 | 38.3 | +15/−15[d] | | 2 |

[a]Referenced to lead.
[b]12–100 °C.
[c]15–35 °C (wire alloy).
[d]40–60 °C (shunt alloy).

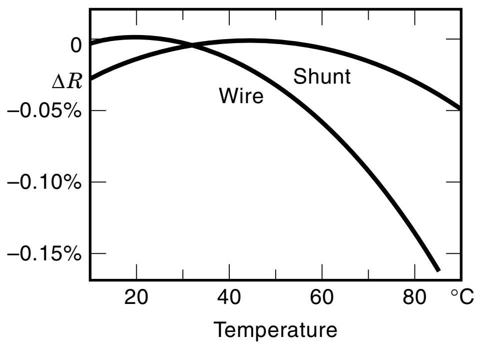


**Figure 2.** Manganin shunt alloy has a minimum resistance change at elevated temperatures to allow for significant power dissipation, while manganin wire is optimized for room temperature operation of "laboratory standard" reference resistors.

the TC turnover point is desirable for the highest temperature stability.

Manganin resistance wire is used for precision resistor standards from about one ohm to thousands of ohms. Power dissipation is minimal in this application, so the alloy content is adjusted to have the minimum resistance variation around room temperature. On the other hand, current shunts are usually required to dissipate significant heat (as noted previously), so Manganin, for current shunt applications, has an elevated TC turnover temperature of 40 to 60 °C.

Manganin can be hard (silver) of soft (tin-lead) soldered, and has a thermal emf compatible with the all of the metals and alloys commonly used in electronics: copper, brass, bronze, silver, gold, tin, and tin–lead solder. Proprietary alloys, with even lower TC, have been developed for high-precision shunts and resistance standards, such as Zeranin 43 (Cu87, Mn7, balance unspecified) by Isabellenhütte in Germany.

### Shunt Resistance

The theoretical resistance of a shunt is calculated from the formula:

$$R_s = \rho l / A \qquad (1)$$

where:

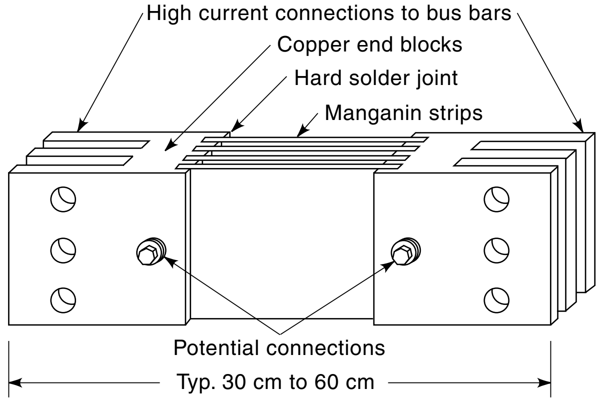


**Figure 3.** Shunts for currents of hundreds to thousands of amps are designed for mounting on one or more bus bars, with good heat dissipation to the bus bars and ambient air.

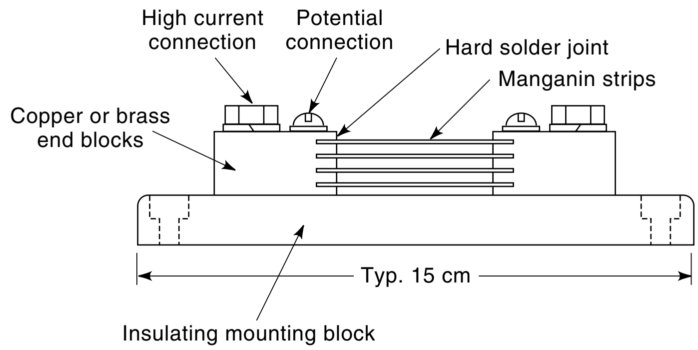


**Figure 4.** Current shunts in the 20 to 200 A range are typically designed for panel mounting.

$R_s$ = shunt resistance;
$\rho$ = resistivity;
$l$ = effective shunt length; and
$A$ = total area of shunt.

This formula is only a close approximation when applied to the shunt material, as there is a small resistance contribution from the end blocks, leads, or other terminations. In addition, the resistivity of the shunt material will vary somewhat from published values, due to residual impurities and processing variations. Precision shunts are designed to have an initial value slightly below the desired value. A calibration adjustment to the desired value is achieved by carefully removing small amounts of shunt material, which decreases the effective area and increases resistance.

## DC CURRENT SHUNT CONSTRUCTION

The construction of typical shunts is shown in Figs. 3, 4, and 5. The high current shunt of Fig. 3 is designed to be bolted directly to bus bars carrying hundreds to thousands of amps. Multiple strips of Manganin are mounted in parallel for low resistance, with spaces between strips for cooling air flow. Copper end pieces are used to minimize their contribution to resistance, and to help conduct heat from the shunt into the bus bars.

A typical 20 to 200 A meter shunt is shown in Fig. 4. The basic construction is similar to that of high-current shunts, but is physically smaller, and wire or heavy cable current connections are often used instead of bus bars. In this case, the

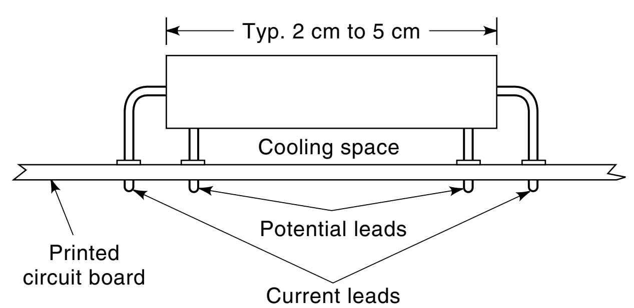


**Figure 5.** Low-current shunts for 1 to 20 A may be mounted on a printed circuit board. A cooling space between the shunt and board is recommended when power dissipation exceeds one watt.

shunt is mounted on an insulated base for mounting on a panel. Copper end pieces are preferred for higher currents, while brass may be used at lower currents.

Low-current shunts (below about 10 A) may be designed for printed circuit board (PCB) mounting, such as shown in Fig. 5. Current shunts with only two terminals may be used for lower currents when high accuracy is not required; these may have wire leads and resemble a conventional resistor in all but resistance value. Low-value "chip resistor" shunts are also finding increasing use for surface mounting on a PCB or other substrate, and these are also often two terminal devices.

The accuracy of Manganin shunts can range from 1% to 0.001% or less. Obtaining the highest accuracies requires careful design, fabrication, annealing, and aging of the shunt, with operation restricted to a limited temperature range. Precision shunts may be forced-air cooled, in order to improve accuracy or current rating, although an oil bath is more common. The oil bath may be stirred and/or water cooled, in order to keep the temperature change to a few degrees or less.

## HIGH-FREQUENCY AC CURRENT SHUNT DESIGN

There are two principal limitations to the accuracy of shunts at higher frequencies; inductive effects, eddy current skin, and proximity effects. Inductive effects are usually the dominant high-frequency error source, with eddy current effects only of concern in essentially "noninductive" current shunts at very high frequencies.

### Shunt Inductance at High Frequencies

A typical parasitic inductance of 10 to 50 nH has little impact on the impedance of a 100 Ω or 1,000 Ω resistor below a few hundred megahertz. However, even 10 nH of equivalent series inductance (ESL) severely limits the higher frequency accuracy of a 10 mΩ shunt; the impedance increases by 1% at 22.6 kHz and 10% at 73 kHz. For wattmeter applications, the phase shift between shunt current and voltage is even more important, particularly for low-power-factor measurements, where the system voltage and current are nearly 90° out of phase. For the above shunt, a 1° phase shift occurs at 2.8 kHz, increasing to 10° at 28 kHz.

The effective ESL of a shunt is the sum of intrinsic inductance of the shunt and mutual inductance coupling between the current input and voltage output leads, although it is not always practical, or even meaningful, to try accurately to separate the two components of inductance.

Accurate current measurement with shunts at high frequencies requires that ESL or its effects be minimized; the use of wirewound resistors for low-current shunts should be avoided for this reason. The intrinsic inductance of conventional low-frequency shunts (Figs. 3 and 4) is not readily reduced, but the larger ESL effects of mutual inductance between current and potential leads can often be reduced by a factor of two to ten, by bringing one or both of the potential sense leads back along the shunt body to a common point, and exiting from the shunt with twisted wires or coaxial cable. Minimizing the inductive area of the current leads would also reduce ESL, in principle, but is usually impractical or ineffective, due to the larger size and relative inflexibility of the conductors.

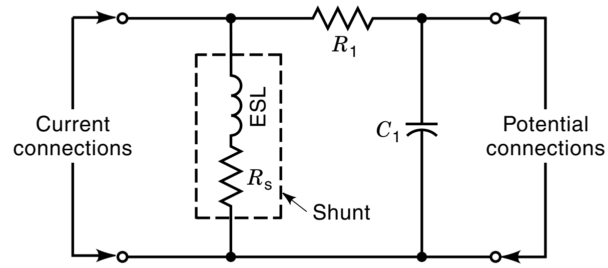


**Figure 6.** The high-frequency accuracy of a current shunt is typically limited by the equivalent series inductance (ESL). The accurate bandwidth can be extended by compensating the shunt's ESL/$R_s$ frequency response zero with an $R_1 \times C_1$ resistor–capacitor pole (see text).

### Shunt ESL Compensation

The effective bandwidth of a shunt can be extended significantly by compensating or canceling the inductive effects with either of two techniques. The first approach is to use a resistor–capacitor ($R$–$C$) network, as shown schematically in Fig. 6. The $R_1$–$C_1$ network of Fig. 6 is used to create a response pole at the same frequency as the $R_s$–ESL zero of the shunt resistance and inductance, which occurs when

$$(R_1)(C_1) = \mathrm{ESL}/R_s \tag{2}$$

This technique can extend the bandwidth of a shunt by orders of magnitude, although some care must be taken in the values of the compensating resistor and capacitor, to minimize other parasitic effects. Compensation resistor values in the range of 30 to 1,000 Ω are recommended; ESL of the resistor may limit bandwidth at lower values, while parasitic capacitance becomes a factor at resistances above this range. Using compensation capacitances below 47 to 100 pF increases sensitivity to loading by other parasitic capacitances, while the ESL of capacitors above about 1 nF may limit the upper frequency of accurate compensation. The ESL of leaded resistors and capacitors is typically 10 to 20 nH, while that of surface mount "chip" devices is significantly lower, at 1 to 5 nH. The parasitic capacitance of a $\frac{1}{4}$ W leaded resistor is about 0.4 pF, while that of a 1206 size chip resistor is less, at about 0.1 pF.

An alternative approach to ESL compensation is to use negative mutual inductance coupling ($-L_m$) between the current and potential leads, as shown schematically in Fig. 7; when $-L_m$ is numerically equal to ESL, the effects of the

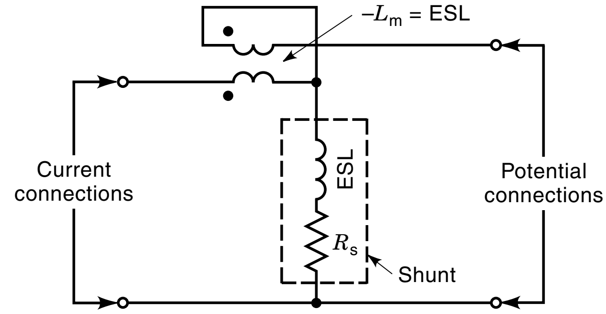


**Figure 7.** The high-frequency bandwidth of a current shunt can be extended by compensating or canceling the shunt's equivalent series inductance (ESL) with a negative mutual inductance ($-L_m$).

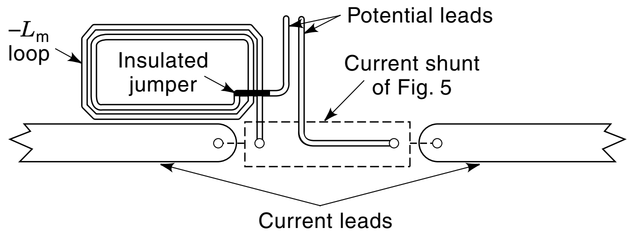


**Figure 8.** An example of $-L_m$ compensation of shunt ESL, achieved by routing of a potential sense lead back through the magnetic field near a current lead.

shunt ESL are canceled. In this approach, the potential sense leads are arranged in the magnetic field of the current leads, in such a way as to create an induced voltage, which cancels the inductive voltage of the shunt. An illustration of this approach is shown in Fig. 8, for the printed circuit board mounted shunt resistor of Fig. 5.

The ESL of a shunt is sensitive to both shunt construction and current and voltage lead placement, so these must both be well controlled and consistent. The ESL compensation must then be calibrated, by comparing the output voltage with the input current at frequencies above and below the $R_s$–ESL zero frequency.

Although the frequency response of the shunt can be measured with various compensations, the simpler time domain approach shown in the test circuit of Fig. 9 is recommended. A voltage pulse from a generator is applied to a noninductive resistor (typically 50 Ω), which is in series with the (relatively low-impedance) shunt. The current in the 50 Ω resistor will have essentially the same waveform as the voltage, so the shunt compensation is adjusted to match the sensed shunt voltage waveform to the input voltage waveform. In the following discussions, the pulse will be assumed to have a smooth risetime and a flat top or plateau, although this is not essential for ESL compensator calibration.

Without compensation, the ESL of a current shunt will cause a large voltage overshoot during the pulse rise time. The overshoot may be tens to thousands of times higher than the steady-state voltage, depending on the shunt resistance, shunt ESL, and the pulse rise time.

When a shunt is undercompensated with an $R$–$C$ network, the shunt voltage will still overshoot the desired value, as

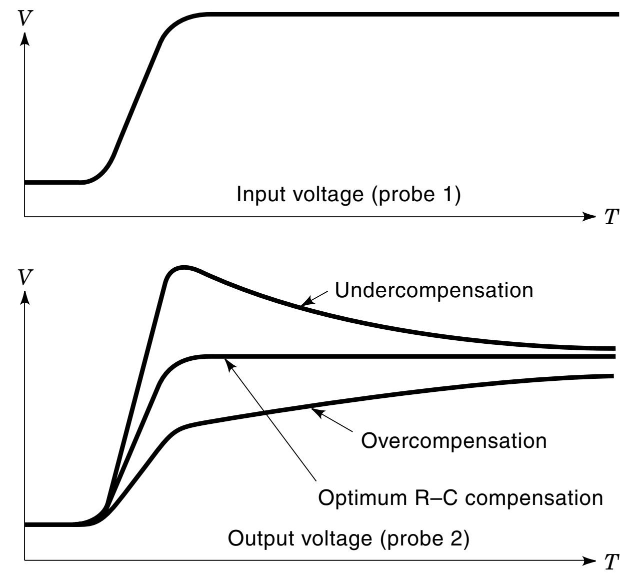


**Figure 10.** The effects on output voltage waveform of over- and undercompensation of shunt ESL with an $R$–$C$ network (as in Fig. 6) are shown for a pulse input using the test setup of Fig. 9. The voltage and time scales are arbitrary.

illustrated in Fig. 10, and then decay to the final voltage at the $R$–$C$ compensation time constant. Overcompensation will cause the initial rise in voltage to be less that the desired value, with a longer rise to the final value, due to the greater $R$–$C$ time constant. Correct compensation will produce a shunt voltage waveform essentially identical to the input pulse voltage waveform.

Full ESL compensation with negative mutual inductance also produces a shunt waveform equal to the input waveform, but under- and overcompensation produces different waveforms, as illustrated in Fig. 11. Undercompensation produces a similar shunt voltage overshoot as with $R$–$C$ compensation, but the voltage falls back to the final value immediately after the input rise is completed. Significant overcompensation will cause the sensed shunt voltage to initially undershoot, or swing in the opposite direction, with an immediate rise to the final value at the end of the input current rise.

Thus both resistor–capacitor and negative mutual inductance ESL compensation of a shunt can produce an output voltage waveform that matches the input current. The voltage

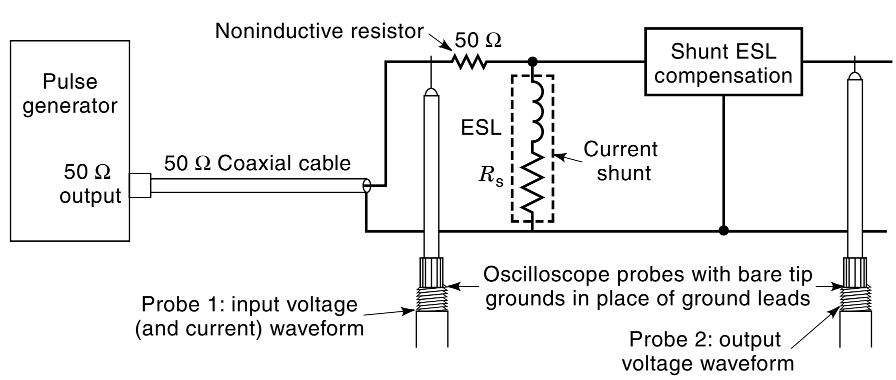


**Figure 9.** A recommended test setup for calibrating shunt ESL compensation, with either an $R$–$C$ pole or negative mutual inductance. The compensation is adjusted to achieve the same waveforms on oscilloscope probes 1 and 2.

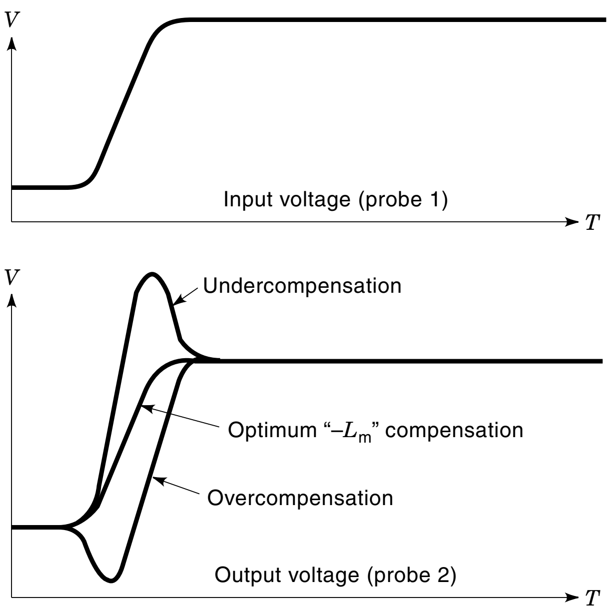


**Figure 11.** The effects on output voltage waveform of over- and undercompensation of shunt ESL with negative mutual inductance (as in Fig. 7) are shown for a pulse input using the setup of Fig. 9. The voltage and time scales are arbitrary.

waveforms due to over- and undercompensation however, are quite different.

## Noninductive Shunts

Shunts can be constructed to appear noninductive at the sense voltage, although there is always some finite inductance in series with the current terminals. These shunts effectively utilize negative mutual inductance ($-L_m$) compensation, but in an automatic way, which does not require calibration.

**Coaxial Shunts.** Coaxial shunts are the more widely known noninductive shunt construction. A cylinder of resistive material is incorporated in a coaxial arrangement with current and potential leads, with the potential leads brought out in a "flux free" region. One such construction is illustrated in Fig. 12.

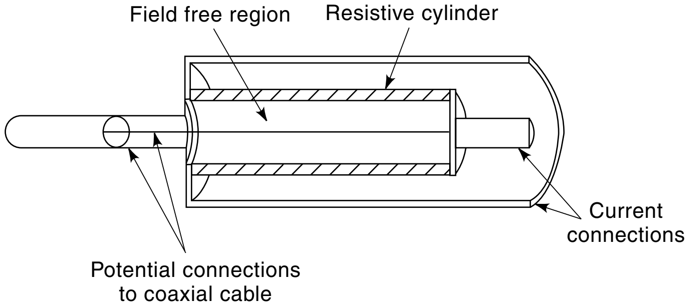


**Figure 12.** Cross-section of a "noninductive" coaxial shunt, which achieves automatic $-L_m$ compensation of the input ESL by bringing the potential sense leads out through a flux free region inside the resistive cylinder.

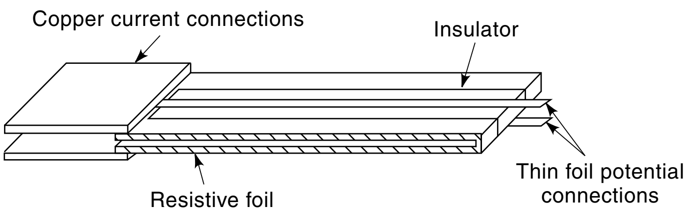


**Figure 13.** Construction of a "parallel plate" current shunt. The potential leads are brought out through a nearly flux-free region, to produce an essentially noninductive shunt.

The coaxial shunt has a nearly zero sense voltage ESL, as none of the magnetic field between the coaxial conductor and shunt material is coupled to the sense leads. However, the coaxial shunt can be inconvenient to use in many applications, particularly when coaxial conductors are not already in use. The two-terminal "insertion inductance" may also be significant (perhaps tens of nanohenries), partly due to the magnetic field between the coaxial cylinders, and partly due to the transition joint required to noncoaxial conductors.

**Parallel-Plate Current Shunt.** An alternative construction for a noninductive shunt is shown in Fig. 13. A relatively thin and annealed Manganin foil may be folded over at the midpoint, with thin insulation (not shown) between the foils. The ends of the foil are soft- or hard-soldered to copper current terminations, which may be left parallel or folded apart at right angles, depending on the application. Thicker or unannealed foils or strips will have to be soldered together, in place of the fold.

Narrower voltage sense lead foils are also soldered to the copper current terminations, and are brought back along the center of the Manganin foils or sheets, as shown in Fig. 13, with thin insulation between the sense leads and the Manganin. Closely spaced foil sense leads may be continued to the shunt voltage sensing point, or a transition made to coaxial cable or a twisted pair of wires.

The parallel plate current shunt can have a sense lead effective ESL nearly as low as the coaxial shunt; Shunts have been constructed with an ESL of less than 50 pH. The insertion inductance may be an order of magnitude lower than that of a coaxial shunt (perhaps several nanohenries or less) when used with noncoaxial conductors.

If the propagation delay along the sense leads of a high-frequency shunt approaches 6 to 8% of a cycle at the highest ac frequency, or a quarter of the 10 to 90% rise time with a pulse current, then transmission line termination techniques should be used to minimize reflections and the resultant waveform distortion and measurement error. If the voltmeter has a high input impedance, a "source termination" at the shunt is advised, using a series resistor equal to the transmission line impedance. This will be 50 or 75 Ω for most coaxial cable, and typically in the range of 100 to 200 Ω for a twisted pair of wires. If the voltmeter input impedance matches the transmission line impedance (typically 50 Ω for RF applications), a source termination is, generally, not necessary.

## Eddy Current Effects in Shunts

Skin effect is classically thought of as the nonuniform current distribution of an isolated conductor at high frequencies, due

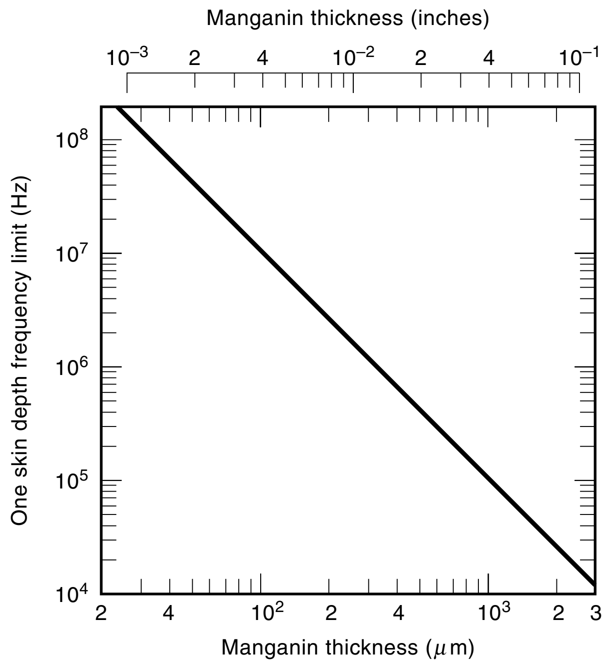


**Figure 14.** The maximum "accurate" frequency of a coaxial or parallel plate current shunt is limited by skin effects. The "one-skin-depth" frequency limit for Manganin increases rapidly with thinner material.

to eddy currents induced in the conductors. In a round conductor at high frequencies, the current is uniform around the surface, while the current falls off exponentially beneath the surface. The "skin depth" is defined as the depth below the surface where the current amplitude is $1/e$ (36.8%) of the surface value. If all of the current were flowing uniformly in the first skin depth of thickness, the effective resistance would be the same. The skin depth varies as the inverse square root of frequency, but also increases as the square root of resistivity.

When two or more conductors carrying high-frequency current are brought near to each other, the current distribution on the surface will change, which is termed a "proximity effect." Currents flowing in opposite directions in two parallel cylinders tend to concentrate on the facing surfaces. The converse effect occurs when the currents flow in the same direction, as the currents begin to avoid the facing surfaces. At the risk of oversimplification, this effect can be envisioned as parallel currents "repelling", while opposing currents "attract".

The concept of skin effect as "the current distribution in an isolated conductor," however, is misleading. For example, the high-frequency current in an isolated bus bar or thin ribbon is not uniform on the surface, but tends to concentrate on the outside edges and corners. The term "skin effect" should be limited to the exponential decay of current normal to the conductor surface; that is, the current distribution perpendicular to the conductor surface.

It is recommended that the current distribution around the conductor be considered a single-conductor proximity effect, caused by the mutual repulsion of parallel current filaments in the conductor. This conceptually explains the uniform surface current distribution in a round conductor and the nonuniform current in a flat conductor, where the current filaments "pile up" at the edges and corners.

The value of this distinction can be appreciated when relatively wide and thin sheets, plates, or bus bars are placed close together and carry high-frequency current in opposite directions, as in the parallel plate current shunt of Fig. 13. The proximity effects now force the high-frequency current to be uniform across the width of the sheets or plates (although only on the facing surfaces if the sheets are thicker than one skin depth).

The skin effect causes the impedance of coaxial or parallel plate current shunts to rise as the thickness approaches a skin depth. The excess impedance is a distributed resistance-inductance effect, asymptotically rising as the square root of frequency with the voltage leading the current by 45°. As such, the skin effect cannot be compensated by a single pole, but would require a distributed $R$–$C$ network or the equivalent.

Thus, skin effect represents a more fundamental upper limit to a shunt's useful frequency than ESL, and is best dealt with by using thinner foils. (The relatively high resistivity of Manganin, about 25 times that of copper, is useful in increasing skin depth and raising the frequency of skin effect impedance increases.) For a high-frequency current penetrating a flat or tubular conductor from one side (as in the parallel plate and coaxial shunts, respectively), the impedance increases about 10% when the conductor is one skin depth thick. For Manganin sheets or tubes with current flow on one surface, the "one skin depth" limit on frequency versus Manganin thickness is given in Fig. 14.

## BIBLIOGRAPHY

1. C. D. Hodgman (ed.), *Handbook of Chemistry and Physics.,* 41st ed., Cleveland, OH: Chemical Rubber Publishing Co., 1960.

### *Reading List*

J. A. Ferreira, W. A. Cronje, and W. A. Relihan, Integration of High Frequency Current Shunts in Power Electronic Circuits, *IEEE Trans. Power Electron.,* **10**: 32–37, 1995.

A. J. Schwab, *Hochspannungsmesstechnik,* New York: Springer Verlag, 1981.

C. M. Johnsson and P. R. Palmer, Current Measurement using Compensated Coaxial Shunts, *IEE Prec.-Sci. Meas. Technol.,* **141** (6): 471–480, 1994.

S. Sevensson, *A Precision Wattmeter for Non-sinusoidal Conditions,* Chap. 5, School Electrical and Computer Engineering, Chalmers University of Technology, Technical Report no. 223L, 1996.

W. Kohler, *A New Shunt for the Measurement of Low Level ac Currents,* Paper 64.01, Dresden, Germany; 7th Int. Symp. High Voltage Eng., August 26–30, 1991.

O. B. Oliveira, M. T. Silva, and P. A. Chagas, *Design Construction and Performance Evaluation of Shunts for the Measurements of High Impulse Current,* Paper 64.02, Dresden, Germany: 7th Int. Symp. High Voltage Eng., August 26–30, 1991.

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