

Simplified Current-Sensing Circuits

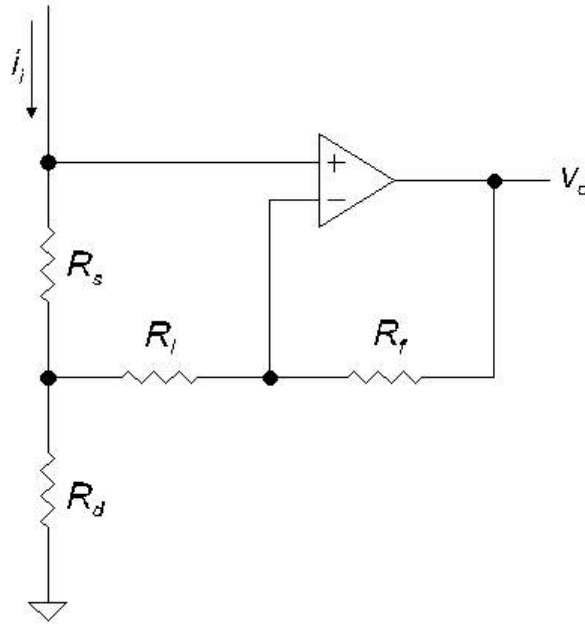
by Dennis L Feucht

Non-isolated current sensing can often be accomplished with a sense resistor followed by an amplifier that scales the output for desired volts/ampere. This article addresses design tradeoffs between using a differential amplifier and a single-ended, non-inverting op amp as the sense amplifier.

If the full-scale current is large the sense resistor is made small to minimize power dissipation and interaction with the sensed circuit. Small-valued resistors can have significant parasitic lead resistance. If it is in series with the accurate sense resistance, R_s , the sensed voltage will be larger by the parasitic resistance voltage drop. The usual solution to this problem is to use a 4-terminal resistor, where a pair of *sense* terminals connected across the desired resistance is brought out separately from the other pair, the *drive* terminals. The sense terminals become the inputs to a differential amplifier, which amplifies only the desired quantity, $i_i \cdot R_s$. This is the optimal approach to sense-resistor circuit design. It is also unnecessary in some applications and more expensive than using a single-ended-input (non-differential) amplifier.

Compensated Single-Ended Sensing

In close circuit-board layouts using two-terminal sense resistors, a single-ended amplifier can be used without appreciable sensing error if the sense resistor has negligible additional resistance to ground in series with it, with respect to the amplifier output. The op amp circuit, shown below, has parasitic resistance, R_d , in series with R_s .



The R_s connections cannot avoid the parasitic resistance of the sense resistor leads themselves. Sensing error caused by board trace resistance can be minimized by running separate sense traces to the sense-resistor pads. (If parasitic resistance of the sense resistor itself is significant, a 4-terminal resistor might instead be required.) The above circuit amplifies the voltage across R_s with a gain of $R_f/R_i + 1$. But this gain, at v_o , is with respect to the junction of R_s , R_d , and R_i , and not ground.

If $R_d \ll R_s$, then negligible error occurs when v_o is taken with respect to ground at the output. This can be achieved by a short, wide circuit-board trace between the bottom of R_s and the ground side of v_o . In good

design direct connection is made instead from the output ground to the bottom terminal of the sense resistor -- the *sense* return node -- for both R_i and the amplifier output. In some circuits, however, the output connection cannot be run back to the sense return but must be connected to the system ground. Then the question of interest is whether the error introduced by R_d can be cancelled by suitable choice of amplifier gain.

Ideally, the transfer function -- a transresistance -- of the sense circuit is:

$$\frac{v_o}{i_i} = R_m = \left(\frac{R_f}{R_i} + 1 \right) \cdot R_s$$

By including R_d , it is:

$$\frac{v_o}{i_i} = \left(\frac{R_f}{R_i + R_d} + 1 \right) \cdot R_s + R_d$$

For $R_i \gg R_d$, the first term -- the amplifier gain term -- is not significantly affected. R_d , the second term, adds directly to the transresistance as an error. R_d in the gain term reduces the transresistance, R_m , but its addition as the second term increases it. With the right choice of R_f or R_i , the opposing effects of R_d cancel. By equating the above transresistance to the ideal expression and solving for R_d , then either $R_d = 0$ (the trivial solution) or:

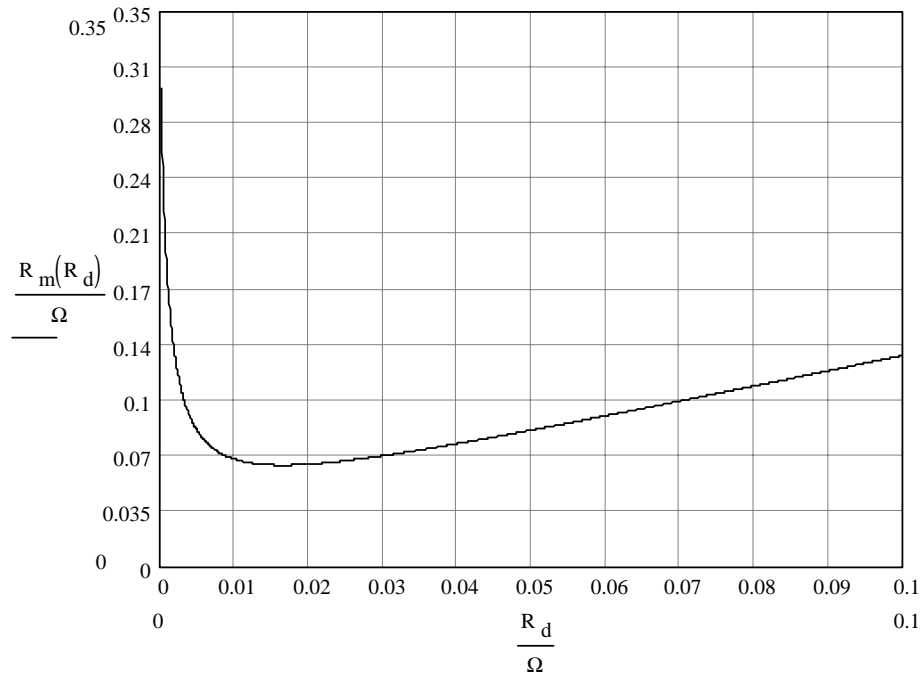
$$R_d = \frac{R_f}{R_i} \cdot R_s - R_i$$

Solving for (positive) R_i that will achieve the desired transresistance:

$$R_i = -\frac{R_d}{2} + \sqrt{\left(\frac{R_d}{2} \right)^2 + R_f \cdot R_s}$$

This formula is applied by example to a circuit with the following values: $R_s = 30 \text{ m}\Omega$, $R_f = 10.2 \text{ k}\Omega$, $R_i(\text{ideal}) = 1.13 \text{ k}\Omega$, and an estimated $R_d = 20 \text{ m}\Omega$. Then the compensating value of $R_i = 17.48 \text{ }\Omega$. This is a rather low value and not practical for maintaining low op amp output current, nor a desired gain of 10. However, with larger R_s values, practical R_i values result. For $R_s = 1 \text{ }\Omega$, then $R_i = 101 \text{ }\Omega$. This compensation technique, therefore, is applicable for low full-scale current values. For high-current sensing, it is not as applicable.

The following graph clearly shows the two effects of R_d , resulting in a minimum in the transresistance function. However, for this minimum to appear, the values of R_f and R_i were reduced from the example values in $\text{k}\Omega$ to $\text{m}\Omega$ instead.



The above equation for R_d can be expressed in R_m by adding R_s to each side of the equation, forming the ideal R_m on the right side, and solving for R_i . Then:

$$R_i = R_m - R_d - R_s$$

For the example, for a desired $R_m = 0.30 \Omega$, then $R_i = 0.25 \Omega$, too small generally to use in practice. What seemed like a hopeful compensation scheme theoretically is dashed upon the rocks of practical parameter values. It also assumes that a sufficiently accurate value for R_d is known for which we are compensating by using ground at the amplifier output. From the transresistance equation, it is evident that for the typical case, $R_i \gg R_d$ and the contribution of R_d is dominantly that of the R_d term, a linear contribution.

Minimizing the Effect of R_d

Because R_d is parasitic and consequently inexact in known value, another possible design goal is to minimize the effect of R_d upon the transresistance. The condition under which this occurs is:

$$\frac{d}{dR_d} \left(\frac{v_o}{i_i} \right) = \left(1 - \frac{R_s \cdot R_f}{(R_i + R_d)^2} \right) = 0$$

Taking the derivative and setting it equal to zero results in:

$$R_f \cdot R_s = (R_i + R_d)^2$$

Solving for R_i :

$$R_i = \sqrt{R_f \cdot R_s} - R_d$$

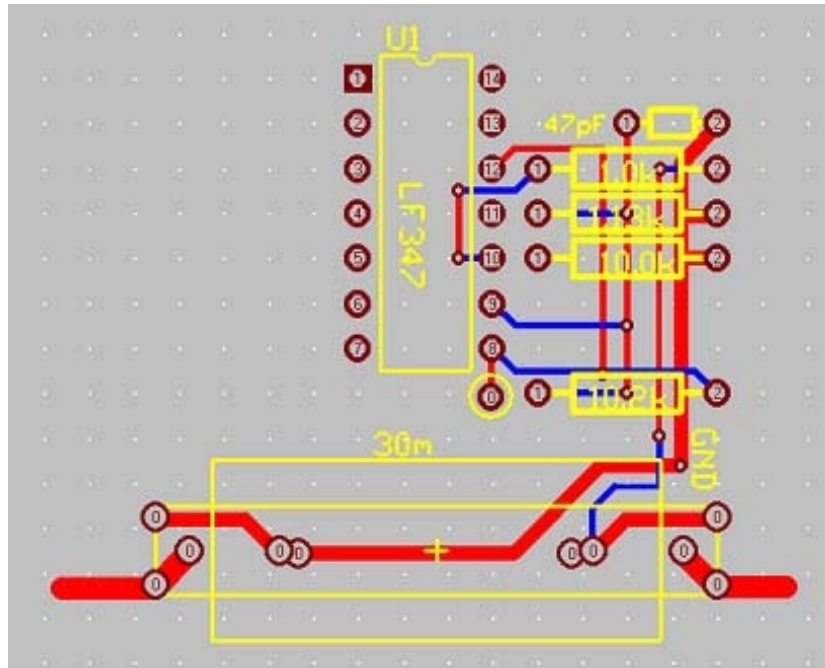
Because R_d is not well known the first term must be made to dominate the maximum estimated value for R_d . By making R_f sufficiently large (and given R_s), then sensitivity to R_d and op amp output current are both minimized. Using parameters of the previous example, the value of R_i is still quite small and not much different: 17.49Ω . If R_f were $102 \text{ k}\Omega$ instead, then $R_i = 55.3 \Omega$, a more workable value, but with an amplifier voltage gain of about 200.

At this point, the creative engineer regards failure as a stepping-stone to a better idea. An engineering maxim that can be applied for this occasion is: "If you can't fix it, feature it." The cause of the problem is R_d , of course. What if, instead of trying to minimize R_d , it is made larger instead? What if an accurate resistor is placed in series with the parasitic R_d with a value that dominates it? For high-current sensing, this intentional R_d cannot be made too large. But in many cases, R_d contributes only a small amount of error, and additional resistance in the high-current input loop can be tolerated. Then adding to R_d solves the problem.

Suppose the parasitic R_d is only around 1 m Ω in the example circuit. It introduces 1/31 or about +3 % error. If an additional R_d is introduced with a value 50 times greater, or 50 m Ω , then the R_f or R_i values are recalculated from the transresistance equations above using an R_d of 51 m Ω . These values need not be kept low to minimize the effect of R_d because it has been stabilized and has become more accurate.

Single-Ended Sense-Amplifier Board Layout

If different sense and output grounds do not allow for the desired accuracy, then the output ground return must be connected to the input sense return node, not the system ground. The drive ground return containing R_d is then no longer critical in board layout.



A layout using a four-terminal resistor and one of the four op amps of an LF347 is shown above. Pin 10 is an op amp + input, 9 the – input, and 8 the output. The 1.0 k Ω resistor in series with the + input is not needed for a FET-input amplifier but is included for bias-current compensation in case a pin-compatible BJT-input op amp is substituted. The 47 pF capacitor is for dynamic compensation. Sensed current flows from right to left through the sense resistor drive terminals connected to the red (top-layer) traces at the bottom of the layout.

The inner terminals of the 5 W, 30 m Ω sense resistor (an IRC product) are the sense terminals, and the left one is the sense return. Note that it is the node to which the ground side of R_i (1.13 k Ω) is attached and should be the return accompanying the amplifier output, v_o , at pin 8. This particular layout was removed and modified slightly from a commercial pulse generator (Innovatia PG1) in which it has performed as expected (and desired).

Closure

A ground-referenced transresistance amplifier can be used to accurately amplify input current in cases where the full-scale current is not large (and R_s is) or when parasitic R_d is large enough to excessively degrade accuracy but small relative to R_s . In the latter case, an accurate and stable additional R_d can be added to dominate the wily parasitic R_d .

In all cases, transresistance accuracy depends upon an accurate R_s for sensing. A 4-terminal Kelvin sense resistor will cost more than a two-terminal resistor and might even be a custom part in a product design to which it does not add distinctive competitive advantage -- an undesirable part to have customized. But it simplifies the entire sensing problem by providing separate sense terminals. Surface-mount sense resistors can in some cases be made four-terminal by board layout, with two separate traces per terminal (connected to the inside center of the pads), thereby eliminating the need for a four-terminal resistor.

For those cases in which the single-ended amplifier does not suffice, and grounding constraints do not allow the methods described here to be applied, then a differential amplifier is required. The differential input senses directly across R_s while the output is referenced to a node other than the sense-return node -- usually the system ground at the input to whatever the sense amplifier drives.

