The Instrumentation Amplifier Handbook

Including Applications

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Instrumentation Amplifiers

Overview

The term "instrumentation amplifier" is properly used to describe a category of true differential-input amplifiers that emphasize high common mode rejection (CMR) and accuracy. Although both instrumentation amplifiers and difference amplifiers use op amps as basic architectural "building blocks", they are distinctly different from their op amp cousins.

Op amps are "single-ended" and they are usually intended to operate in a variety of applications- with their feedback determining their functions. Instrumentation amplifiers and difference amplifiers are used primarily to provide differential gain and common mode rejection. Employing feedback from output to input is not intended.

In some instances this term has been widely misused and this has created confusion as to the correct definition of an instrumentation amplifier (IA). In the early days of monolithic operational amplifiers, one well-known vendor referred to their new precision op amp as an instrumentation amplifier. What they meant to say was that it was an "instrumentation-grade" op amp.

In addition, large laboratory bench-top amplifiers and even traveling- wave tube (microwave) amplifiers have been called instrumentation amplifiers. It is not surprising, then, that so much confusion exists about what an IA really is and what it does.

Most common IAs are one of three types: the simple "Difference Amplifier", the "Two Op Amp Instrumentation Amplifier", and the "Classical Three Op Amp Instrumentation Amplifier" architecture. As we shall see, these three architectures are interrelated but their performance differs in certain important aspects. For now, let's just think of the IA as a "black box" differential amplifier.
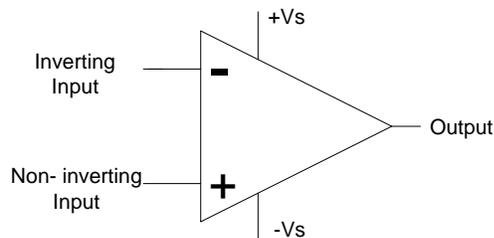
Figure 2- 1. A "Black Box" Representation of an Instrumentation Amplifier.

The Wheatstone Bridge Sensor

To better understand the instrumentation amplifier and why its high common-mode rejection is so important, let's take a look at one of the most common transducers in use today—the Wheatstone Bridge. While the usual way of depicting the bridge circuit is shown the diagram of Figure 2a, it can be redrawn (Figure 2b) to show that the bridge is nothing more than two voltage dividers driven by a single voltage (Vex) or current (Iex) excitation source.
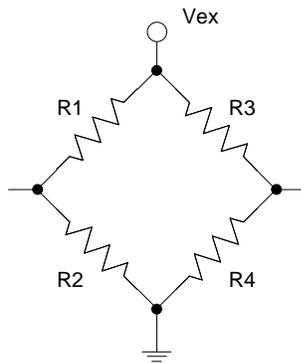
Figure 2-2. (a.) Conventional Bridge Circuit.   (b.) Redrawn Bridge Circuit.
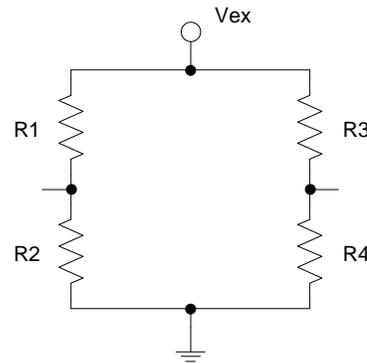
Let's look at an example of a Wheatstone Bridge sensor with zero inputstimulus (pressure, temperature, force, etc.):

$$V_{ex} = +10\text{V}$$

At zero applied force the sensor bridge is balanced and

$$R_1 = R_2 = R_3 = R_4 \quad \text{so:}$$

$$I_1 = \frac{V_{ex}}{R_1 + R_2} \qquad I_1 = 1\text{mA}$$

$$I_2 = \frac{V_{ex}}{R_3 + R_4} \qquad I_2 = 1\text{mA}$$

$$V_1 = I_1 \times R_2 \qquad V_1 = +5\text{V}$$
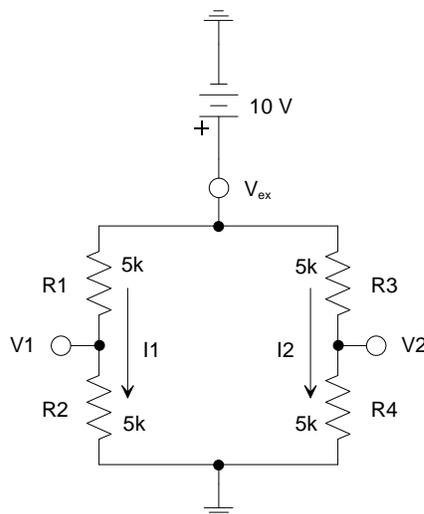
$$V_2 = I_2 \times R_4 \qquad V_2 = +5\text{V}$$

Figure 2-3. Balanced Bridge Generates CMV But No Differential Output Voltage.

With no stimulus applied to the sensor, all arms of the bridge are equal and R1= R2= R3= R4. The current in one side of the bridgeis :

$$I_1 = \frac{V_{ex}}{R_1 + R_2}, \qquad I_1 = \frac{10V}{5k + 5k}, \qquad I_1 = 1mA, \tag{1}$$

On the other side of the bridge:

$$I_2 = \frac{V_{ex}}{R_3 + R_4}, \qquad I_2 = \frac{10V}{5k + 5k}, \qquad I_2 = 1mA \tag{2}$$

The output voltage (V1) on one side of the bridge is:

$$V_1 = I_1 \times R_2, \qquad V_1 = 1mA \times 5k, \qquad V_1 = 5V \quad \text{with respect to ground.} \tag{3}$$

Similarly, the output voltage (V2) on the other side of the bridge is:

$$V_2 = I_2 \times R_4, \qquad V_2 = 1mA \times 5k, \qquad V_2 = 5V \quad \text{with respect to ground.} \tag{4}$$

The sensor output is the voltage difference (ΔV) between the two sides of the bridge. At zero applied stimulus, the bridge differential output voltage is zero:

$$\Delta V = V_1 - V_2, \qquad \Delta V = 5V - 5V, \qquad \Delta V = 0V \tag{5}$$

The problem for our amplifier is thatΔV is measured by subtracting (taking the difference of) one large voltage V2 from another large voltage V1. Since these large voltages (V1 and V2) appear on both sides of the bridge, they are "in common", while the desired bridge output voltage (ΔV) –perhaps only a few microvolts appears as a differential output measured between the sides ("legs") of the bridge.

The Wheatstone Bridge was an early example of a "ratiometric" measurement. Ironically, pioneer physicists were not inconvenienced by the presence of common mode voltage on the bridge. The only sensitive measuring instrument available to them was the galvanometer, which was simply connected to each side of the Wheatstone Bridge. Fortunately, this also provided a floating measurement of the bridge differential output voltage. The entire galvanometer was at the voltage divider potential.

Modern ground-referenced amplifiers have greatly increased the sensitivity of bridge measurements over those made by the old galvanometer method but, since they are not "floating" like the old galvanometers, they have also introduced a common mode voltage (CMV) limitation to the measurement. Ground-referenced IAs and difference amplifiers have specified operating limits on CMV– linear operation is possible only within these limits and permanent damage can occur if operated beyond the "Absolute Maximum" device ratings.

A stimulus applied to the sensor will change the resistance of one or more resistors (arms) of the bridge. Most sensor designs now employ a "four active arm" design as it maximizes the sensitivity and linearity of the bridge sensor. Early sensor designs employed only one or two active arms but they are not frequently encountered in modern systems.

In the four active arm type of bridge, force is applied so that when R1 increases, R4 also increases and R2 and R3 both decrease. In this way, the voltage on one side of the bridge increases while the voltage on the other side decreases a corresponding amount.

Let's consider an example where a small stimulus is applied to the sensor, causing the bridge to be unbalanced, so that R1= R4= 4.999K and R2= R3= 5.001K.

Since the total resistance on each side of the bridge (R1+R2 on one side and R3+R4 on the other) remains constant at 10K, the current in each side will still be 1mA. From Equations (3) and (4) above:

V1= 5.001V ,   and V2= 4.999V.

Using Equation (5), the bridge differential output voltage is ΔV = 2mV.
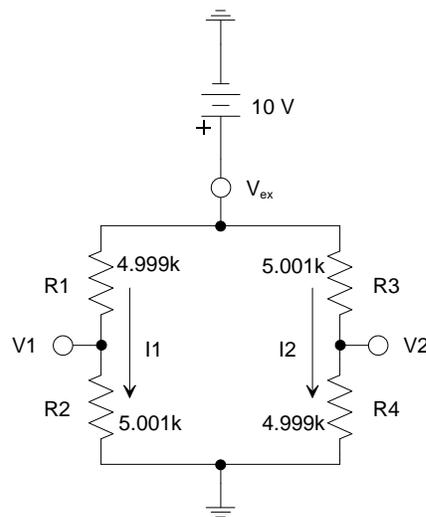
$$V_{ex} = +10\text{V}$$

With applied force the sensor bridge is unbalanced and

$$R_1 = R_4 \quad \text{and} \quad R_2 = R_3 \quad \text{so:}$$

$$I_1 = \frac{V_{ex}}{R_1 + R_2} \qquad I_1 = 1\text{mA}$$

$$I_2 = \frac{V_{ex}}{R_3 + R_4} \qquad I_2 = 1\text{mA}$$

$$V_1 = I_1 \times R_2 \qquad V_1 = +4.999\text{V}$$

$$V_2 = I_2 \times R_4 \qquad V_2 = +5.001\text{V}$$

Figure 2- 4. Unbalanced Bridge Generates A Differential Output Voltage.

Calculating the CMV reveals that although the bridge is unbalanced its CMV has not changed:

$$CMV = \frac{V_1 + V_2}{2}, \qquad CMV = \frac{5.001\text{V} + 4.999\text{V}}{2}, \qquad CMV = 5\text{V} \qquad (6)$$

The task of our differential amplifier is to provide adequate gain (G) to amplify the bridge sensor's 2mV output (Vin) to whatever level is required by the following stage (Vout).

In this example, let us choose a level of 1V full-scale:

$$G = \frac{V_{out}}{V_{in}}, \qquad G = \frac{1\text{V}}{2\text{mV}}, \qquad \text{so the required amplifier gain } G = 500\,\text{V/V} \qquad (7)$$

Error Sources

Real-world instrumentation amplifiers are not ideal devices and the error contributions of their non-ideal parameters must be considered.

Serious measurement errors can occur if an amplifier exhibits poor common mode rejection. These CMR errors are caused by a shift in the amplifier's input offset voltage (V_{os}) due to the applied CMV. Common mode rejection is defined by:

$$CMR = 20 * \log \frac{CMV}{\Delta V_{os}} \qquad (8)$$

To illustrate the importance of amplifier common-mode rejection and its contribution to measurement error, let's assume an amplifier Common Mode Rejection specification of 80 dB. By rearranging Equation (8), the CMR error can be calculated:

$$\Delta V_{os} = CMV \times 10^{-\left(\frac{CMR}{20}\right)}, \qquad \Delta V_{os} = 5\text{V} \times 10^{-\left(\frac{80\text{dB}}{20}\right)}, \qquad \Delta V_{os} = 5\text{V} \times 10^{-4},$$
$$\Delta V_{os} = 500\text{uV}$$

Clearly, a 500uV error on a 2mV signal is not acceptable; our amplifier needs far higher CMR. An amplifier with a 100 dB CMR specification will reduce the input offset shift due to CMV to 50uV, which may be a more acceptable error of 2.5%. Reducing this error by another decade will require another 20 dB of CMR from the amplifier.

Instrumentation amplifiers are optimized for the high precision requirements of this type of application. Not every instrumentation amplifier can achieve this high level of performance, however. Only the best devices can achieve CMR specifications of 120 dB—a high figure indeed. A premium performance device like the Burr-Brown INA128 is specified to have 120 dB minimum and 130 dB typical Common Mode Rejection at the high gain required by our sensor's small output signal.

Other error terms such as input offset voltage and drift, power supply rejection (PSR), and input bias current will also contribute to the amplifier's total measurement error. Needless to say, we wish to minimize these measurement errors so we must select an amplifier with adequate performance specs in all areas.

Power supply rejection errors are caused by a shift in the amplifier's input offset voltage (V_{os}) due a *change* in supply voltage. Power supply rejection is defined by:

$$PSR = 20*\log\frac{\Delta V_s}{\Delta V_{os}}\qquad(9)$$

Power supply rejection ratio (PSRR) is also frequently specified in a data sheet (instead of PSR) as a direct change in V_{os} per volt of supply voltage change:

$$\pm\Delta V_{os}/\Delta V_s,\quad\text{usually as}\quad \mu V/V.\qquad(10)$$

To illustrate amplifier power supply rejection and its contribution to measurement error, let's assume that we have a +/-15V amplifier that's being operated on +/-12V supplies and has a power supply rejection specification is 80 dB. By rearranging Equation (9), the PSR error can be calculated:

$$\Delta V_{os} = \Delta V_s \times 10^{-\left(\frac{PSR}{20}\right)},\quad \Delta V_{os} = (15\text{V} - 12\text{V})\times 10^{\left(\frac{-80\text{dB}}{20}\right)},\quad \Delta V_{os} = 3\text{V}\times 10^{-4},\quad \Delta V_{os} = 300\mu\text{V}$$

Once again we find that a 300uV error on a 2mV signal is not acceptable; our amplifier needs far higher PSR. An amplifier with a 100 dB CMR specification will reduce the input offset shift due to CMV to 30uV.

Input offset voltage is an error source that can be easily calculated. V_{os} is an input-referred specification (like CMR and PSR) so the IA offset voltage can simply be added to the differential signal voltage at the instrumentation amplifier's input. Input offset voltage drift is the change in V_{os} per degree Celsius change in ambient temperature. Drift error is treated in the same manner as input offset voltage error. Both signal and all input-referred errors (such as V_{os}, drift, CMR, and PSR) are multiplied by the gain of the amplifier but at the amplifier's output, the ratio of the signal to errors remains constant, however.

Instrumentation amplifiers may have input-referred specifications that contain gain dependent equations rather than the simpler and more familiar single value specs found in op amp and diff amp data sheets. There is a good reason for this as we will see later.

Input bias current (I_b) flowing through the amplifier's source resistance generates an additional offset voltage that must be added to the input errors. High impedance signal sources can cause serious offset problems with bipolar transistor input IAs. In these applications low bias current FET-input amplifiers are recommended to minimize I_b errors with very high source resistances.

The circuit shown in Figure 5 can model input errors. Total IA error is the sum of all individual errors.
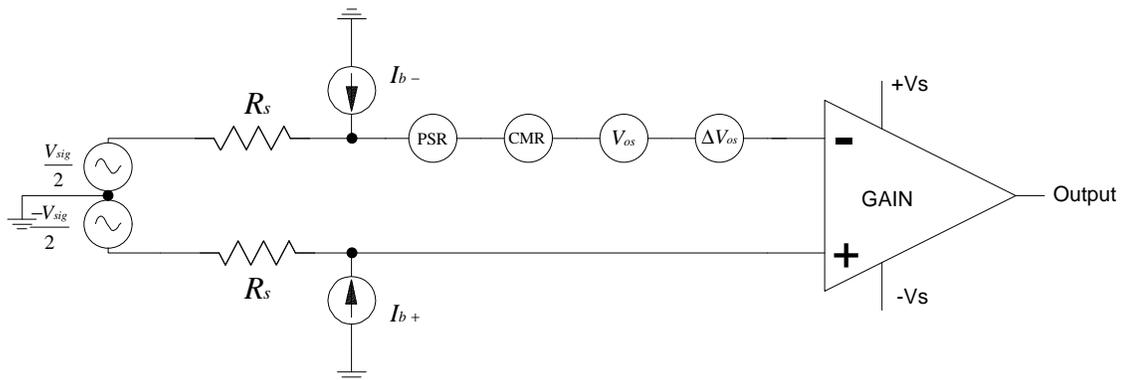
Figure 2- 5. A Model Of Input- Referred Errors.

An error analysis is performed by considering the amplifier's operating temperature range, power supply regulation, common- mode voltage, source resistance, and gain. For a worst- case (albeit pessimistic) analysis, use the min/max specs in the amplifier's data sheet. An analysis of more likely "real world" errors—since all parameters are unlikely to be at their spec limit at the same time—is performed with the data sheet "typical" specs.

How reliable the "typical" specs are depends, to some degree, on the vendor's integrity but as most characterization measurements are made when the product is first put into production, any process changes incorporated in the course of normal manufacturing may have skewed the statistical distribution from the original tests . Bear in mind that using "typical" specifications entails some risk; typicals are not guaranteed because no semiconductor manufacturer can guarantee the statistical distribution of his yields will remain invariant for all time. Only the data sheet min/max specs are guaranteed.

Amplifier Topologies

The Difference Amplifier

Let's take a look into our "Black Box". The simplest form of instrumentation amplifier is the Difference Amplifier[1], an op amp with four precision resistors as shown in Figure 6.
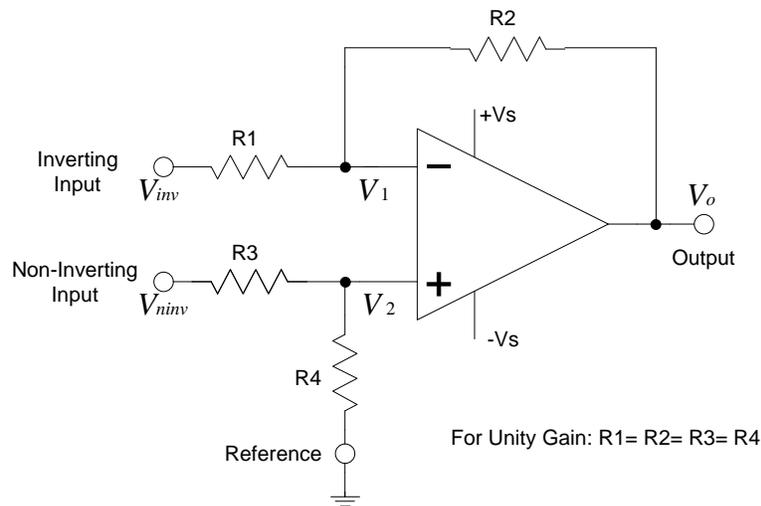
Figure 3- 1. The Unity- Gain Difference Amplifier.

The input (R1) and feedback (R2) resistors of the diff amp in Figure 6. form the familiar op amp unity- gain inverter (since R1= R2) configuration. Therefore, if we ground the other (non-inverting) input, the gain at this input is:

$$V_o = V_{inv} \times \frac{-R_2}{R_1} \qquad \text{and since R1= R2:} \quad V_o = V_{inv} \times (-1) \quad \text{or:} \qquad V_o = -V_{inv} \qquad (12)$$

The non- inverting op amp input is connected to a voltage divider formed by R3 and R4. The voltage at this point (V2) is:

$$V_2 = V_{ninv} \times \frac{R_4}{R_3 + R_4} \qquad \text{and since R3= R4:} \quad V_2 = V_{ninv} \times \frac{1}{2} \quad \text{or:} \quad V_2 = \frac{V_{ninv}}{2} \qquad (13)$$

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[1] Operational Amplifiers- Design and Applications: Tobey, Graeme, Huelsman. McGraw- Hill 1971. pp. 202

If we analyze the inputs separately, the amplifier analysis is a little easier to understand. So, let's ground the inverting input. Now we can recognize the resulting circuit as a non- inverting amplifier:

$$V_o = V_2\left(1 + \frac{R2}{R1}\right) \qquad \text{and since R1= R2:} \qquad V_o = 2V_2 \qquad (14)$$

but as shown in equation (13):

$$V_2 = \frac{V_{ninv}}{2} \qquad \text{so:} \qquad V_o = 2 \times \frac{V_{ninv}}{2} \qquad \text{or} \qquad V_o = V_{ninv} \qquad (15)$$

Now we have found that the amplifier gain at the non- inverting input is unity (+ 1V/V) and we have already found that the amplifier gain at the inverting input is unity- gain inverting (-1V/V). Combining Equations (12) and (15) yields the result:

$$V_o = V_{ninv} - V_{inv} \qquad (16)$$

By definition, common mode voltage is equal at each input, $V_{ninv} = V_{inv}$, which makes $V_o = 0$. The amplifier has completely rejected (another way of stating that the common mode gain is zero) the CMV appearing at its inputs! The ideal IA, therefore, amplifies only the small differential signal at its inputs while completely rejecting a much larger common mode voltage also appearing on those same inputs.

Resistors cannot be perfectly matched so: $\qquad \dfrac{R_1}{R_2} \neq \dfrac{R_3}{R_4} \qquad$ which places a limitation on achievable CMR. An analysis will show that the resistor matching required to achieve acceptable CMR in a diff is surprisingly difficult! Using the equations above, assume a resistor matching of 0.01%. How much CMR can be achieved? We've achieved 80 dB CMR but there are two important conditions that must be met. If the source impedance is not perfectly equal at each input and the resistance from the Reference Pin to ground is not zero, the CMR will be seriously degraded.

Why is this? Simple– remember that the source resistance appears in series with R1 and R3 and the Reference Pin ground return resistance appears in series with R4. These additional resistances can degrade the resistor ratio matching. Don't forget the resistor temperature coefficients (TC)-- the resistor ratios must be maintained over the amplifier's operating temperature range, too.

What seems so simple to build with a handful of discrete parts turns out to be far more difficult and expensive than it is at first glance. Monolithic diff amps use on- chip laser- trimmed thin-film nichrome resistor networks that exhibit far better matching and TC tracking that a discrete PC board design. For equivalent CMR performance, the monolithic diff amp offers lower cost and smaller size than a discrete approach.

Real world diff amps have one outstanding disadvantage over the two and three op amp configuration IAs: low input impedance. A non-zero-input resistance will draw current from the signal source, degrading its accuracy and linearity. This low input resistance, coupled with

the severe limitation of requiring equal source resistances to prevent CMR degradation, makes the two and three op amp instrumentation amplifier more attractive than the diff amp in many applications.

Make no mistake, though; diff amps have their place. Wherever you have low impedance sources, high common mode voltages, or severe cost restraints, the simple low cost diff amp can be your best choice.

The "Classical" Three Op Amp Instrumentation Amplifier

You may well ask "Since the limitations of the diff amp are input resistance related, can't we put a buffer amplifier on the inputs to solve these problems?" The answer is "Yes! In fact, that's the whole idea behind the three op amp instrumentation amplifier." Not only will we resolve the diff amp limitations and achieve very high input impedance, these input buffer amplifiers can also provide voltage gain.

Here's how it's done:
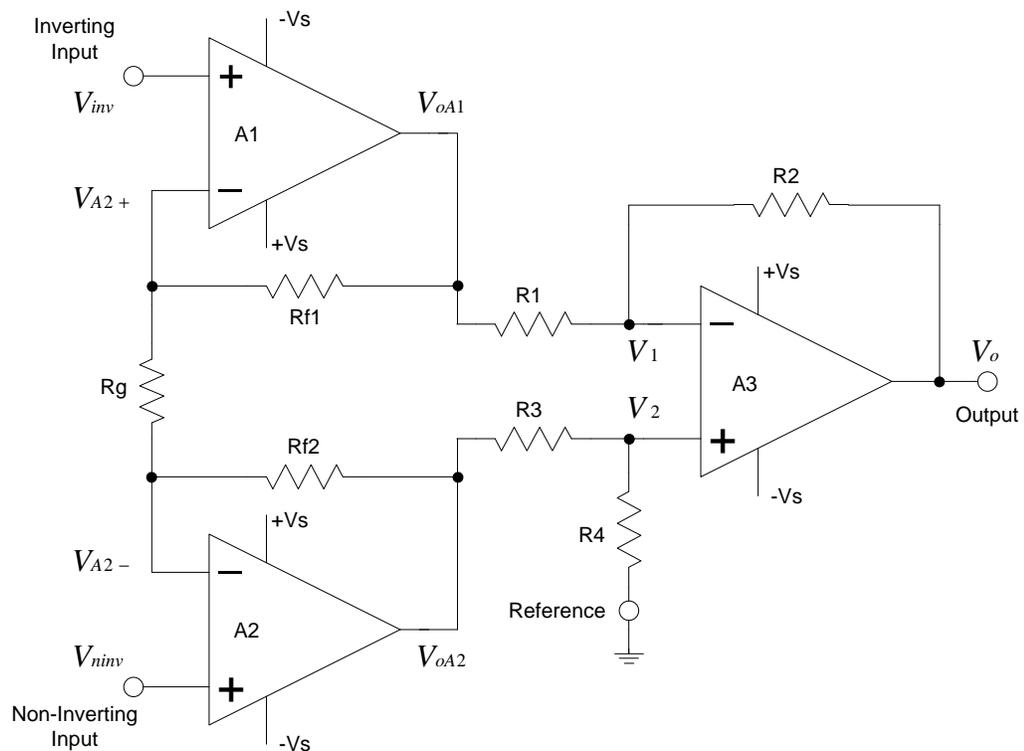
Figure 3-2. The "Classical" Three Op Amp Instrumentation Amplifier.

Operational amplifiers A1, A2, and the feedback network Rg, Rf, and Rf form a differential-input differential-output amplifier that drives an output stage difference amplifier (A3 plus R1, R2, R3, and R4).

The output diff amp's function in the three op amp IA is to provide common mode rejection. The input amplifiers may provide gain[2] but no CMR; A1 and A2 are both non-inverting operational amplifiers so any CMV at their inputs is amplified by +1 and appears at their respective outputs. CMR is provided solely by the output difference amplifier– we'll look at this later.

As A1 and A2 are non-inverting op amps, their input impedance is very high [3], thus eliminating any undesirable source loading and removing the strict requirement of having equal source impedances at each input. In fact, except for having to consider input bias current effects, the IA's inputs do not affect the signal source at all.

To understand how this IA works, let's once again analyze each input and assume that the non-inverting input is grounded while a signal of 1V is placed on the inverting input. For simplicity we will assume that all resistors in the IA are 1k ohm and that we have ideal op amps. This results in the circuit of Figure 8. If the non-inverting input of A1 is 1V, the open-loop gain of the amplifier forces its differential voltage to zero, so its inverting input must also must be 1V.

In a similar manner, since the non-inverting input of A2 is grounded (0V), then its inverting input must also be 0V. This places a 1V potential across Rg, which is 1k ohm, so the current through Rg has to be 1mA. In an ideal op amp, no current flows in the amplifier's inputs, so the 1mA current in Rg must also be flowing through Rf1, causing a 1V drop across Rf1. As we have seen, the bottom end of Rg is grounded, so the voltage on the right side (in our drawing) of Rf is: 1V+1V. Thus the output of A1 is 2V. This is consistent with the familiar non-inverting op amp gain equation:

$$A_V = 1 + \frac{R_{f1}}{R_g} \qquad \text{or} \qquad A_V = 1 + \frac{1\text{k}}{1\text{k}} \qquad \text{so} \qquad A_V = 2\text{V}/\text{V} \qquad (17)$$

Amplifier A1's output will be: $V_{oA1} = V_{inv}\left(1 + \frac{R_{f1}}{R_g}\right)$ or $V_{oA1} = 1\text{V}\left(1 + \frac{1\text{k}}{1\text{k}}\right)$ and $V_{oA1} = 2\text{V}$

The 1mA flowing through Rg does not flow into the inverting input of A2; it therefore flows through Rf2 and develops a 1V drop across it. The left end of Rf2 is at ground potential, so its right end , which is the output of A2, is at -1V.

[2] Depending on the value of Rg.

[3] Example: OPA128 typical input impedance: 10^{10} & 2pF differential, 10^{11} & 9 pF common-mode

Again, this is consistent with the inverting op amp gain equation:

$$A_V = -\left(\frac{R_{f2}}{R_g}\right) \quad \text{or} \quad A_V = -\left(\frac{1k}{1k}\right) \quad \text{so} \quad A_V = -1V/V \quad (18)$$

Amplifier A2's output will be: $V_{oA2} = (1\,V) \times \left[-\left(\frac{R_{f2}}{R_g}\right)\right]$ or $V_{oA2} = (1\,V) \times \left[-\left(\frac{1k}{1k}\right)\right]$

so $\quad V_{oA2} = -1V$
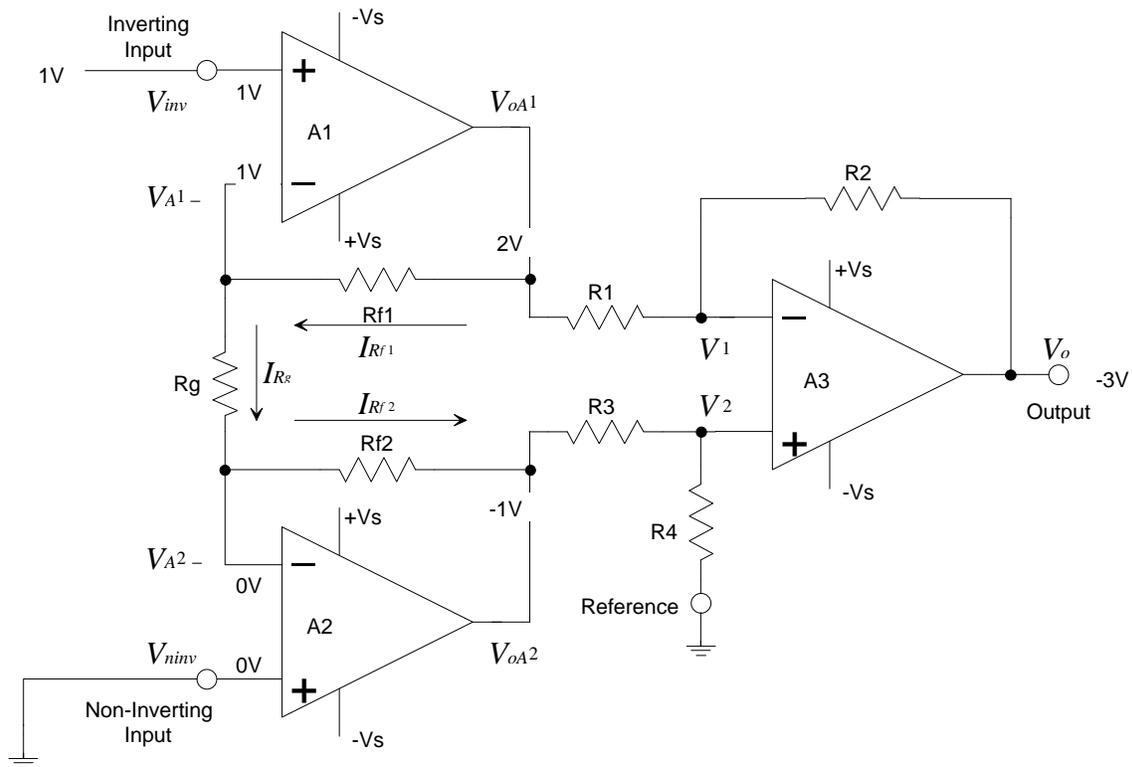
Figure 3- 3. Three Op Amp Instrumentation Amplifier Gain Analysis

The differential output of A1 and A2 is:

$$\Delta V_o = V_{oA1} - V_{oA2} \quad \text{or} \quad \Delta V_o = 2V - (-1V) \quad \text{or} \quad \Delta V_o = 3V \quad (19)$$

If we let R1/R2= R3/R4, using Equation (16), the diff amp stage gain is -1V/V, so the IA output is -3V.

Thus the overall gain equation for a three op amp instrumentation amplifier with a unity gain difference output stage is:

$$A_V = 1 + \frac{2R_f}{R_g}. \quad (20)$$

The buffer amplifier stage is perfectly symmetrical so this analysis can be applied to the other (inverting) input with identical results.

If Rg is open, both A1 and A2 are reduced to functioning only as unity gain buffers and the overall IA voltage gain is simply that of its difference amplifier stage: (usually) unity gain (1 V/V).[4]

So far we've shown that this IA configuration has gain, but does it also have CMR? Let's take a look.

Keeping things simple again, let's put 1V on each input of the IA. The circuit is shown in Figure 9.
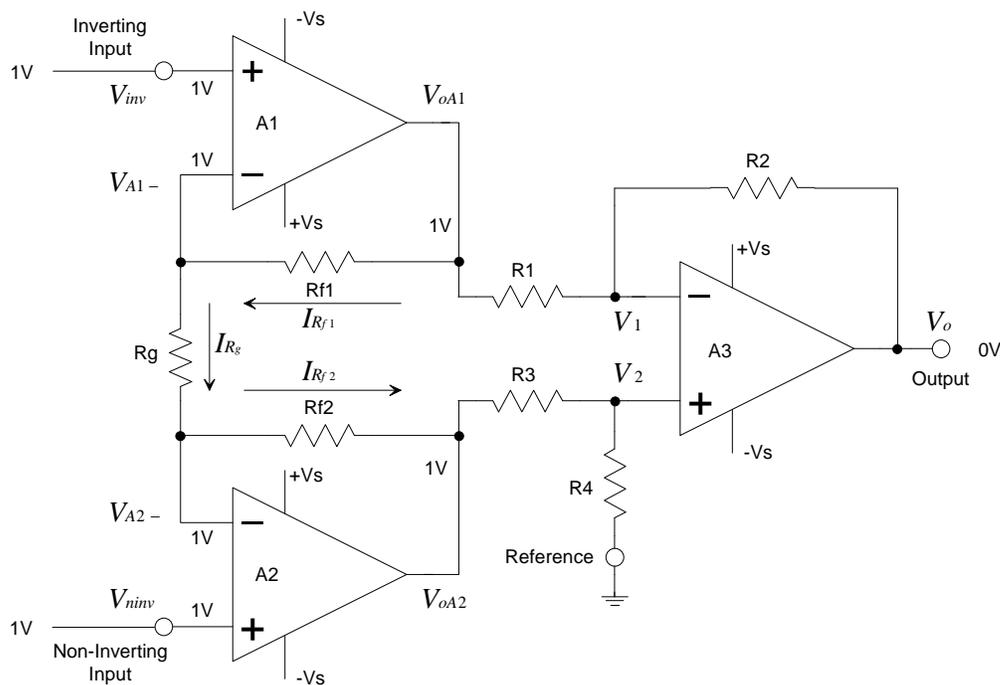
Figure 3- 4. Three Op Amp IA Common Mode Voltage Analysis

Open loop gain of A1 and A2 will reduce their respective differential input voltages to zero, so $V_{A1-} = V_{inv}$ and $V_{A2-} = V_{ninv}$. Inspection of the circuit shows that there is 1V on each end of Rg. Since the net voltage across Rg is zero, the current flowing through it (I_{Rg}) must also be zero. Reminding ourselves that no current can flow into the inputs of an ideal op amp, we see that if $I_{Rg} = 0$, then the currents I_{Rf2} and I_{Rf1} must be also zero because the same current flows through all three resistors.

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[4] This is usually true but there are a few exceptions where the diff amp has gain. An INA106 has a gain of 10V/V.

Applying our gain equations: amplifier A1's output voltage is:

$$V_{oA1} = V_{inv}\left(1 + \frac{R_{f1}}{R_g}\right) - V_{A2-} \quad \text{so} \quad V_{oA1} = 1\text{V}\left(1 + \frac{1\text{k}}{1\text{k}}\right) - 1\text{V} \quad \text{therefore:} \quad V_{oA1} = 1\text{V} \quad (21)$$

and amplifier A2's output voltage is:

$$V_{oA2} = V_{A1-} \times \left[-\left(\frac{R_{f2}}{R_g}\right)\right] + V_{ninv}\left(1 + \frac{R_{f2}}{R_g}\right) \quad \text{SO} \quad V_{oA2} = 1\text{V} \times \left[-\left(\frac{1\text{k}}{1\text{k}}\right)\right] + 1\text{V}\left(1 + \frac{1\text{k}}{1\text{k}}\right) \quad (22)$$

therefore: $V_{oA2} = -1\text{V} + 2\text{V} = 1\text{V}$.

But both A1 and A2 have output voltages equal to the input CMV. We haven't accomplished any CMR at all in the input buffer amplifier stage! All is not lost– both output voltages are equal, so the following stage– the diff amp– will give us the needed common mode rejection in this type of instrumentation amplifier.

Common mode rejection is provided solely by the diff amp, so in very low impedance sensor unity gain applications, a three op amp IA doesn't necessarily have better CMR than a simple diff amp. In applications where high gain is required, the Classical IA configuration wins "hands down". It's no coincidence that this IA topology has become so popular.

If you have not been deterred from building your own diff amp with discrete components, you may even be tempted to try the same thing with a three op amp instrumentation amplifier. A bit of advice: don't. It is more difficult than it may seem at first.

Internal vs. External Gain Setting Resistors

In addition to all of the strict requirements of resistor matching and TC tracking, this IA design adds a few more: unless the feedback resistors of A1 and A2 are exactly equal ($R_{f1} = R_{f2}$), their unequal gains will degrade CMR ($V_{oA1} \neq V_{oA2}$). In addition, the feedback resistor to gain set resistor ratio ($\frac{R_{f1}}{R_g}$ and $\frac{R_{f2}}{R_g}$) must be very accurate and this ratio must not change over temperature. Otherwise, the overall gain accuracy of the IA will be compromised. Still think you want to build your own IA?

In fact, this gain set resistor ratio is a limitation on the accuracy that can be achieved by a monolithic IA which uses an external resistor to set its voltage gain. An external resistor will not match the temperature coefficient of the internal on-chip thin film resistors (in the amplifiers' feedback) because of its physical separation and also possibly due to their different resistive materials . For applications that require the highest gain accuracy and gain stability, choose an amplifier having internal on-chip thin film resistor gain-setting networks.

The internal gain-set instrumentation amplifiers INA131 and INA141 offer gain TCs as low as ±10ppm/C maximum[5] – an order of magnitude better than the best "external resistor" amplifiers can achieve.

"Internal resistor" amplifiers lack only the flexibility of choosing an arbitrary voltage gain [6]. Of course, an external resistor can be added in series with the internal gain resistor to reduce gain but its TCR mismatch will degrade the amplifier's gain stability. If the external resistor is much smaller than the internal gain resistor (less than about 10%), this may be acceptable.

The Two Op Amp Instrumentation Amplifier

The last instrumentation amplifier that we will look at is the two op amp type shown in Figure 10.
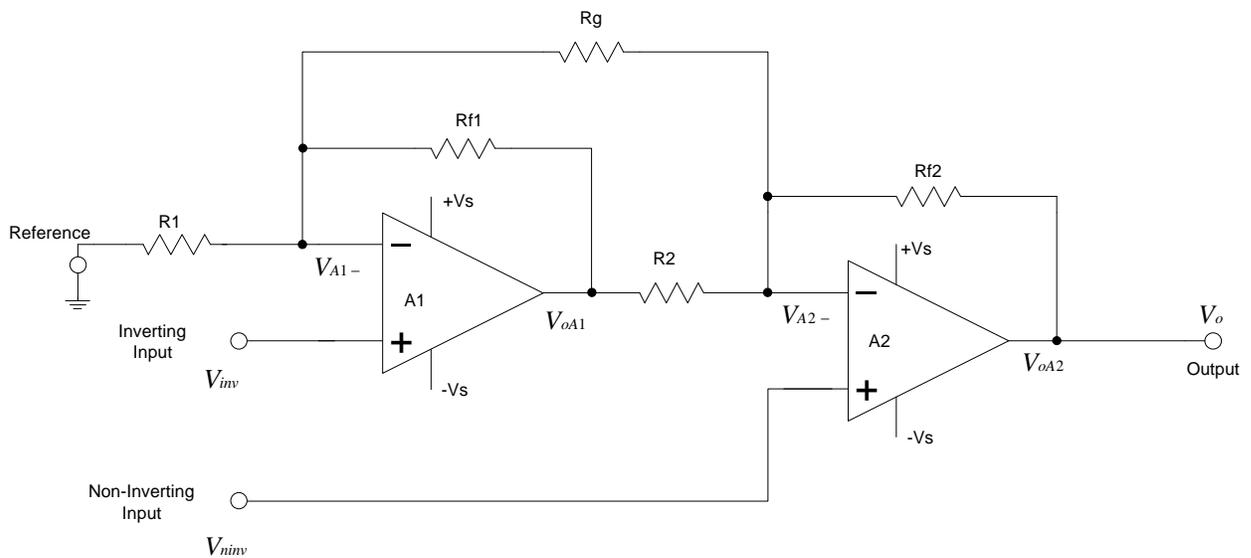
Figure 3-5. The Two Op Amp Instrumentation Amplifier

In this simpler IA, we have fewer op amps and precision resistors, both of which reduce the IC die size and manufacturing cost. The input impedance is still as high as in the three op amp configuration, however, as we are still using non-inverting op amps as the IA inputs. There is one feature that <u>has</u> been sacrificed for circuit simplicity: unity gain (+1 V/V) is not possible with the two op amp instrumentation amplifier [7].

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[5] INA131 and INA141 have typical gain TC specs of only +/- 5ppm/C and +/- 2ppm/C respectively.

[6] INA131 has a gain of 100 V/V while the INA141 has gains of 10 and 100 V/V.

[7] The two op amp instrumentation amplifier is limited to a minimum gain of 2 V/V or more, depending on the ratios of the internal resistors Rf1/R1 and Rf2/R2.

If we analyze this IA's differential gain in the same manner as we have before, we'll start by grounding the IA non- inverting input and applying a 1V signal to the IA inverting input. For the purposes of simplifying this analysis, we will assume all resistors to be 1k ohm. This will result in the conditions shown in Figure 11:
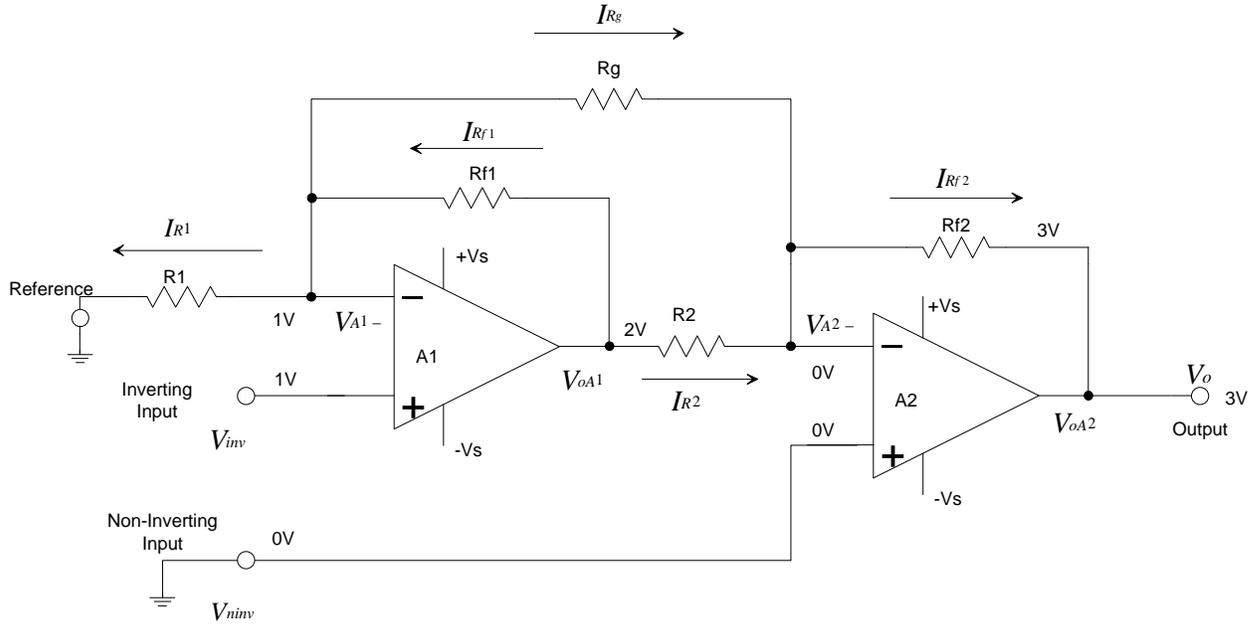
Figure 3- 6. Two Op Amp Instrumentation Amplifier Gain Analysis.

Due to the loop gain of A1, both input voltages of A1 must be equal: $V_{inv} = 1V \therefore V_{A1-} = 1V$. One end of resistor R1 is connected to the Reference pin which is grounded. It then has a potential difference of 1V so by Ohm's Law:

$$I_{R1} = \frac{V_{A1}}{R_1} \qquad \text{or} \qquad I_{R1} = \frac{1V}{1k} \quad \text{so} \qquad I_{R1} = 1mA \tag{23}$$

Due to the loop gain of A2, both input voltages of A2 must also be equal: $V_{ninv} = 0V \therefore V_{A2-} = 0V$. Now we see that one end of Rg is at 1V and the other is at 0V.

$$\text{Therefore:} \qquad I_{rg} = \frac{V_{A1-} - V_{A2-}}{R_g} \qquad \text{or} \qquad I_{rg} = \frac{1V - 0V}{1k\Omega} \qquad \text{so:} \qquad I_{rg} = 1mA \tag{24}$$

Since no current can flow into (or out of) the inputs of an ideal op amp, the sum of the currents flowing through R1 and Rg1 must be supplied by a current that flows through Rf1. In other words:

$$I_{Rf1} = I_{R1} + I_{Rg} \qquad \text{and} \qquad I_{Rf1} = 1mA + 1mA \qquad \text{or} \qquad I_{Rf1} = 2mA \tag{25}$$

Now we can calculate the voltage across Rf1:

$$V_{Rf1} = I_{Rf1} \times R_{f1}, \qquad V_{Rf1} = 2\text{mA} \times 1\text{k}\Omega \qquad \text{or} \qquad V_{Rf1} = 2\text{V} \tag{26}$$

We have already found that there is 1V on the left end of R1 ($V_{A1-} = 1\text{V}$) so the 1v drop across Rf1 is added to arrive at the voltage on the right end of Rf1. This voltage is supplied by amplifier A1's output.

$$\text{So:} \qquad V_{oA1-} = V_{A1-} + V_{Rf1} \qquad \text{and} \qquad V_{oA1-} = 1\text{V} + 2\text{V} \ \text{ so } \ V_{oA1-} = 3\text{V} \tag{27}$$

We already know that $V_{A2-} = 0\text{V}$ so the current in R2 can be easily calculated:

$$I_{R2} = \frac{V_{oA1} - V_{A2-}}{R_2} \qquad \text{and we know that} \qquad I_{R2} = \frac{3\text{V-0V}}{1\text{k}\Omega} \qquad \text{so} \qquad I_{R2} = 3\text{mA} \tag{28}$$

Once again, the sum of the currents flowing in R2 and Rg flow through Rf2. Calculating I_{Rf2}, we see

$$I_{Rf2} = I_{R2} + I_{Rg} \qquad \text{and} \qquad I_{Rf2} = 3\text{mA} + 1\text{mA} \qquad \text{so} \qquad I_{Rf2} = 4\text{mA} \tag{29}$$

The summing junction of A2 is at ground potential, so by calculating the voltage across Rf2 (V_{Rf2}) we can find the output voltage of A2 (V_{oA2}):

$$V_{Rf2} = I_{Rf2} \times R_{f2} \qquad \text{and} \qquad V_{Rf2} = 4\text{mA} \times 1\text{k}\Omega \qquad \text{or} \qquad V_{Rf2} = 4\text{V} \tag{30}$$

But $V_{oA2} = V_{A2-} - V_{Rf2}$ so $V_{oA2} = 0\text{V} - 4\text{V}$ therefore $V_{oA2} = -4\text{V}$ (31)

We now know the output of the IA is -4V for an inverting input of 1V; our instrumentation amplifier in this example has a voltage gain of -4 V/V.

In the two op amp IA topology, internal resistors must be such that: $R_1 = R_{f2}$ and $R_2 = R_{f1}$ (we'll see why later). A generalized gain equation [8] for the two op amp instrumentation amplifier can be expressed as:
$$A_V = -\left(1 + \frac{R_1}{R_{f1}} + 2 \times \frac{R_{f2}}{R_g}\right) \tag{32}$$

We've now found that this type of IA does have voltage gain; but does it also exhibit common mode rejection? Our circuit analysis could be done as follows: we will assume that all of the resistors are 1k ohm and that we have 1V on the IA inverting input. These conditions are the same as our previous gain analysis example.

Now let's apply 1V to the IA non-inverting input as well, thus generating a 1V common mode voltage.

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[8] J.G. Graeme, *Applications Of Operational Amplifiers; Third-Generation Techniques*, McGraw-Hill, 1973, pp. 56.

The circuit diagram of the CMR analysis is shown in Figure 12.
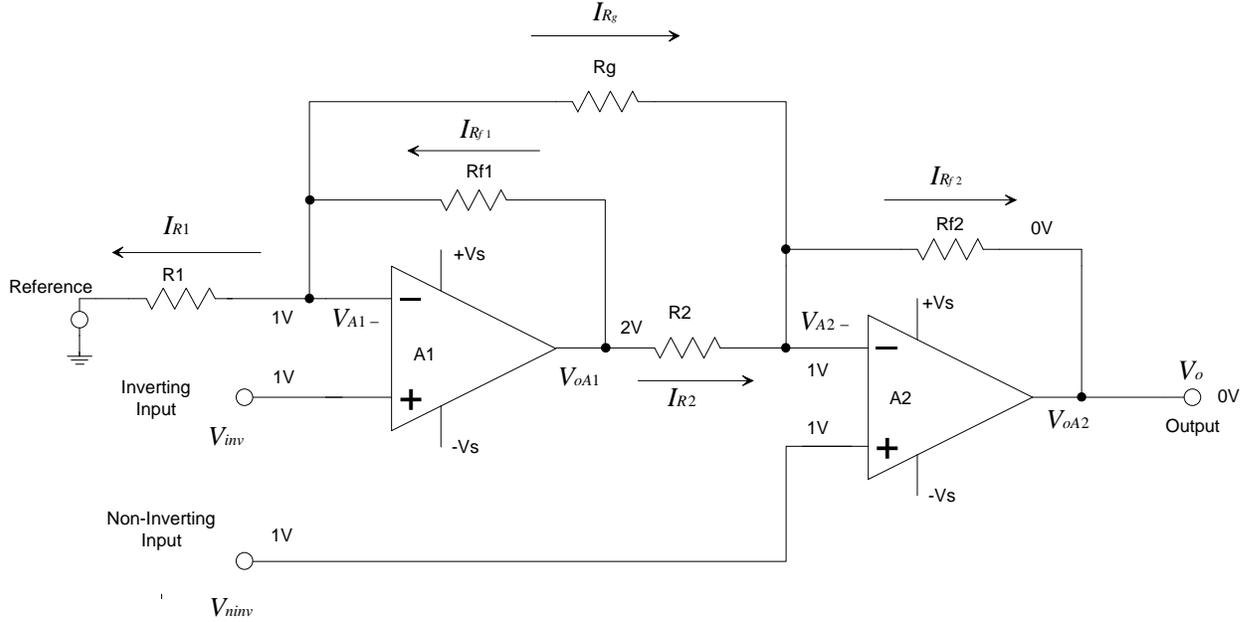
Figure 3- 7. Two Op Amp IA Common Mode Rejection Analysis.

In the gain analysis, we determined that V_{A1-} is 1V and that I_{R1} is 1mA. As we now have 1V on the non- inverting IA input, the loop gain of A2 and its associated feedback will force its summing junction to be 1V as well. By Equation (24) we find that the current through Rg is zero since there is 1V on each end of Rg. By Eqn. (25), it is found that I_{Rf1} is 1mA and therefore the voltage across A1's feedback resistor (V_{Rf1}) is 1V. We have already found that there is 1V on the left end of Rf1 ($V_{A1-}=1V$) so the 1V drop across Rf1 is added to arrive at the voltage on the right end of Rf1. This voltage is supplied by amplifier A1's output.

So: $\qquad V_{oA1-} = V_{A1-} + V_{Rf1}$ $\qquad$ and $\qquad V_{oA1-} = 1V + 1V$ $\qquad$ so $\qquad V_{oA1-} = 2V$.

Since no current flows in Rg, it may be completely ignored. This allows us to greatly simplify the analysis of A2. Notice that R2 and Rf2 allow A2 to be an inverting op amp with a gain of :

$$A_{vi} = -\frac{R_{f2}}{R_2} . \qquad\qquad\qquad (33)$$

Substituting known values: $\qquad A_{vi} = -\frac{1k\Omega}{1k\Omega}$ $\qquad$ resulting in $\qquad A_{vi} = -1V / V$.

The output of A2 due only to this inverting gain path is:

$$A_{oA2} = V_{oA1} \times -1V / V \qquad\qquad \text{or} \qquad A_{oA2} = 2V \times -1V / V , \quad \therefore A_{oA2} = -2V . \qquad (34)$$

This is only a partial result, as we have haven't yet considered the 1V on the IA's non-inverting input. At this input we can see that the same components (R2, Rf2, and A2) form a non-inverting op amp with a voltage gain of: $A_{vni} = 1 + \dfrac{R_{f2}}{R_2}$.

Substituting, we have: $A_{vni} = 1 + \dfrac{1k\Omega}{1k\Omega}$ so $A_{vni} = 2V/V$. (35)

The output of A2 due only to the non-inverting gain path is:

$A_{oA2} = 1V \times A_{vni}$ or $A_{oA2} = 1V \times A_{vni}$, $\therefore A_{oA2} = 2V$. (36)

We can find the resultant output due to both gain paths by combining Equations (34) and (36):

$A_{oA2} = (-2V) + (2V)$ so $A_{oA2} = 0V$.

We've just proven that the two op amp instrumentation amplifier has the ability to reject common mode voltages. At last!

This type of IA is no exception to the rule that its resistors must be very well matched and their TCs must track closely to achieve high amplifier gain accuracy and CMR performance. This IA topology does have the advantage of having fewer high precision resistors to contend with and fewer op amps. For the user, this means that a two op amp IA is usually– but not always-- lower cost and has lower quiescent current than a classical three op amp design.

That being the case, why not just adopt the cheaper two op amp IA for all applications? For one thing, remember that this IA topology has a minimum gain limitation (in our example, if Rg is left unconnected the minimum voltage gain [9] is 2V/V) and its CMV operating range is different from the classical topology.

Because of differences in the noise gain of the two op amps, the two op amp IA does not preserve symmetry of its common mode error well at higher frequencies and, as a result, its CMR falls off faster with frequency than a classical three op amp amplifier. Due to the equal noise gains of the two input op amps in a three op amp IA, their bandwidths are well matched and there is little unmatched "residual error" to degrade CMR at high frequency.

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[9] Instrumentation amplifiers such as the INA122 and INA126 have a minimum gain of 5V/V.

Avoiding Instrumentation Amplifier Pitfalls

Input Bias Current Effects

The effect of amplifier input bias current on its input signal source impedance should not be overlooked. This is one of the most important things to check when selecting an amplifier and if a serious blunder is made, the resulting mismatch between the IA and its source will create large errors in the circuit.

Let's investigate a "horrible example" of a bad source-to-IA mismatch:

Attempting to design a low noise accelerometer preamplifier, our neophyte designer thumbs through a catalog and finds the INA103 Low Noise, Low Distortion Instrumentation Amplifier. "Great," he thinks to himself, "it has less than 1nV per √Hz noise, its super low distortion means it also has excellent linearity, and it has great CMR. I couldn't ask for a better amplifier!" He has just fallen into a very common trap– his accelerometer is a self-generating *piezoelectric* type and it (the INA103) is a completely unsuitable choice. A very high load impedance is required at the accelerometer output to preserve its low-end frequency response and this impedance is the cause of serious problems.

Here– in Figure 13-- is his preamp circuit; let's see how much trouble he is in.
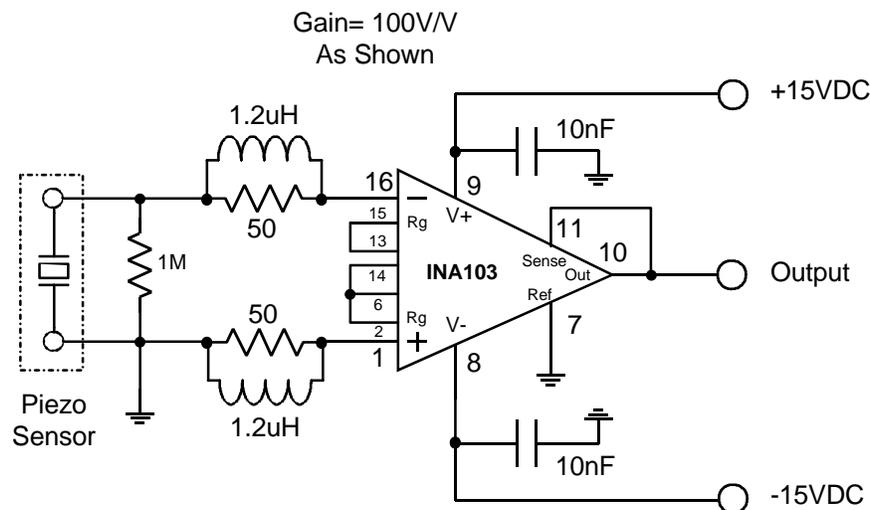
Figure 4-1. An Unsuitable IA Choice For A High Impedance Transducer.

As already pointed out, very high load impedance is required at the accelerometer output to preserve its low- end frequency response. This is because of the capacitive nature of a piezoelectric transducer-- it can be modeled as an AC current source in series with a capacitor. Extending the transducer & amplifier low frequency response means that the transducer capacitance and its load resistor RC time constant must be as large as possible.

The accelerometer used in this example required a 1 megohm load resistor to meet his low frequency response requirement. The 1 megohm load resistor is where the problem is, An INA103 is specified to have an input bias current of 2.5μA typical and 8μA maximum. The inverting input (-) bias current (V_{OS}) must be drawn through the negligible resistance (perhaps 10 ohms) of an RL network[1] in series with the 1 megohm load resistor.

The resulting voltage drop creates an offset at the amplifier's (-) input:

$$V_{OS-} = I_{b-} \times R_L, \quad \text{with } I_{B-} = 2.5mA \text{ and } R_L = 10 + 10^6 \quad \text{or} \quad R_L \cong 10^6 ohms \quad \text{(xx)}$$

And we get an inverting input offset of: $V_{OS-} = (2.5 \times 10^{-6} A) \times (10^6 ohms)$ or $V_{OS-} = 2.5V$

That's an input offset of *2.5 volts* ! (Offset generated by the 10 ohm resistance in the other input is small enough to be ignored.) This huge offset is amplified by the INA103's gain of 100, resulting in the amplifier's being driven into the rail. Clearly, *this* circuit won't work at all.

Our neophyte designer then may attempt to balance the impedances seen by each IA input in order to cancel the bias current offsets. To which we reply "Nice try, but no cigar!" Looking at the offset created by adding a 1 megohm resistor from the non-inverting (+) input to ground, we first see that the input bias current of the (+) input is not equal to the input bias current of the (-) input.

This bias current difference– offset current (I_{os})-- is specified as 30nA typical and 500nA maximum. The (+) input bias current is therefore:

$$I_{B+} = I_{B-} + I_{OS} \quad \text{so:} \quad I_{B+} = 2.5 \times 10^{-6} A + 30 \times 10^{-9} A \quad \text{and:} \quad I_{B+} = 2.53 \times 10^{-6} A$$

Which gives an offset at the (+) input of: $V_{OS+} = (2.53 \times 10^{-6} A) \times (10^6 ohms)$ or $V_{OS+} = 2.53V$

Now we have an offset of 2.5V on one input and 2.53V on the other. The instrumentation amplifier's common mode rejection will allow only the voltage difference (ΔV) to be amplified.

Thus he is amplifying a much smaller input signal than in his first try:

$$\Delta V = V_{OS-} - V_{OS+} \quad \text{or} \quad \Delta V = 2.53V - 2.5V \quad \text{so} \quad \Delta V = 0.03V \quad \text{or} \quad 30mV.$$

---

[1] The RL network improves the stability of the INA103.

Now, when multiplied by a gain of 100, we see an output voltage of "only" 3V– still unacceptable but at least it is better than last time. Before we proceed to show why even heroic measures such as applying 3VDC to the IA reference pin in an attempt to zero this 3V output offset won't make this amplifier/transducer work, we can point out a math shortcut.

Although the analysis of the effect of balancing input impedances to reduce offset voltage was calculated for each input independently, we can, instead, simply consult the data sheet and use Input Offset Current rather than each individual input's Bias Current. This shortcut works, provided of course, that the offset voltage at each input is not so high that it exceeds the amplifier's common mode voltage range.

We could have calculated ΔV this simpler way:

$$\Delta V = I_{OS} \times R_L \quad \text{or} \quad \Delta V = 30 \times 10^{-9}\, A \times 10^6\, ohms \tag{xx}$$

and we obtain the same answer: 30mV at the INA103 input.

Although we have shown that this circuit has a large DC offset, it may be tempting to argue that since a piezoelectric transducer output rolls off at low frequency (it has no DC response), this circuit will still be acceptable if the amplifier's output is simply AC coupled through a capacitor to its following stage. *Not so!*

A calculation of the input noise voltage resulting from the amplifier's high input bias current will reveal why this "low noise" amplifier is not suited to high source impedances. With very low impedance sources, the INA103's input voltage noise density (e_n) of 1nV/√Hz will predominate but as the source impedance becomes higher, the amplifier's current noise density (i_n) of 2pA/√Hz will generate significant noise voltage (e_{in}) across its source resistance:

$$e_{in} = i_n \times R_L \quad \text{and since} \quad i_n = 2\, pA / \sqrt{Hz} \quad \text{and} \quad R_L = 1 megohm$$

$$e_{in} = (2 \times 10^{-12}\, A\sqrt{Hz}) \times 10^6\, ohms \quad \text{or} \quad e_{in} = 2 \times 10^{-6} V / \sqrt{Hz} \tag{xx}$$

This is 2μV/√Hz ; it is _far_ larger than the 1nV/√Hz that our clueless designer had expected. In fact, the total amplifier noise will be slightly higher than 2μV/√Hz since we must add the amplifier's input voltage noise spectral density to it. In this case, adding the voltage noise contribution to the current noise contribution will result in only a small increase in noise.

 Noise voltages add vectorially due to their being uncorrellated. Our total noise will thus be:

$$e_T = \sqrt{\left[(e_n)^2 + (e_{in})^2 \right]} \quad \text{so} \quad e_T = \sqrt{\left[\left(10^{-9} V / \sqrt{Hz}\right)^2 + \left(2 \times 10^{-6} V / \sqrt{Hz}\right)^2 \right]} \quad \text{and}$$

$$e_T \cong 2 nV / \sqrt{Hz} \tag{xx}$$

Since the voltage noise density due to the amplifier's current noise density (in this example) is so much larger than the amplifier's input voltage noise density, the resultant noise is still only slightly more than $2\mu V/\sqrt{Hz}$.

If the balanced input resistance scheme is attempted, the amplifier's total noise will be even higher! In this case, there is $2\mu V/\sqrt{Hz}$ of uncorrellated noise at each input. Common mode rejection cannot reject completely uncorrellated input noise. The two input noise voltages will add arithmetically rather than vectorially (being multiplied by $\sqrt{2}$). Similarly, differential amplifier inputs cannot subtract uncorrellated input noise; rather is the same as simply adding two equal noise voltages. Here, the amplifier's CMR is also adding these same uncorrellated input noise voltages. The total equivalent input noise (e_{EQ}) due to the noise at the inverting input (e_{T-}) and at the non-inverting input (e_{T+}) is:

$$e_{EQ} = \sqrt{\left(e_{T-}\right)}^2 + \sqrt{\left(e_{T+}\right)}^2 \tag{xx}$$

thus we have $\quad e_{EQ} = \sqrt{\left(2\times10^{-6}V / \sqrt{Hz}\right)}^2 + \sqrt{\left(2\times10^{-6}V / \sqrt{Hz}\right)}$

and so the total input noise of the balanced impedance amplifier is $4\mu V\sqrt{Hz}$. But we are not finished-- this must be multiplied by the amplifier's gain of 100V/V, so the output voltage noise density will be $400\mu V\sqrt{Hz}$.

Finally, we must multiply the IA output noise voltage density by the square root of the amplifier's bandwidth to obtain the RMS noise (E_n) seen at the output. The INA103 is a wideband amplifier-- it has a -3dB bandwidth of 800kHz in a gain of 100V/V.

$$E_n = (400\times10^{-6}V / \sqrt{Hz}) \times \sqrt{800\times10^3\,Hz} \tag{xx}$$

so $\quad E_n = 0.35777V \quad$ or 358mV rms at the output.

And, since the amplifier's equivalent noise bandwidth is a factor of 1.57 times its -3dB bandwidth (assuming a single-pole response), we must again multiply to find our final value:

$$E_{rms} = E_n \times 1.57 \tag{xx}$$

so the noise seen at the INA103 output is 562mV rms. Even worse, the peak-to-peak noise will be about six times higher!

It now seems obvious that the INA103-- while it is an excellent choice for low source impedances-- performs poorly when it is misapplied to a very high impedance transducer.

A FET input instrumentation amplifier would be a far better choice for this application. Repeating these calculations with an INA111 or INA121 should be convincing.

A simple rule of thumb:

1. Consider your source impedance when selecting an amplifier.

2. High source impedances require JFET or CMOS amplifiers.

3. Very low input noise (E_n) amplifiers require low source impedances.

Avoiding the "Reference Pin" Trap

The reference pin of a difference amplifier– whether it is a stand-alone device or whether it is the third op amp in a three op amp IA– and the reference pin of a two op amp IA allow a designer great flexibility. The most common use for this pin is for generating output voltage offsets. Unlike "trim" pins, the reference pin allows very large offsets (up to several volts) to be generated without degrading the amplifiers input voltage drift.

It is necessary to drive this pin with a "zero ohm" source resistance so that the amplifier's CMR is not compromised. In the analysis, common mode rejection was shown to depend on a very close resistor matching in their internal network. Additional resistance between the reference pin and ground will generate a resistance mismatch and CMR will suffer. After all, $R = R + R_S$ only if $R_S = 0$.

The source resistance of whatever is connected to the reference pin must be as close to zero as possible. In an attempt to generate an offset voltage, this rule is sometimes overlooked as in the abortive attempt shown in Figure 14. Here is a graphic example of "How *Not* To Do It." In this circuit, a 1k pot is used to generate an offset voltage at the INA122 reference pin.

The high resistance to ground seen at the reference pin ruins the CMR performance of the IA. Even when the offset adjust pot wiper arm is turned toward the end connected to ground, CMR is degraded. Remember, though, that the potentiometer end resistance is not zero– it can be surprisingly high. A typical potentiometer end resistance specification is "1 ohm or 2%, whichever is higher." So the end resistance of the 1k pot could be as high as 20 ohms– high enough to cause a measurable reduction of amplifier CMR.
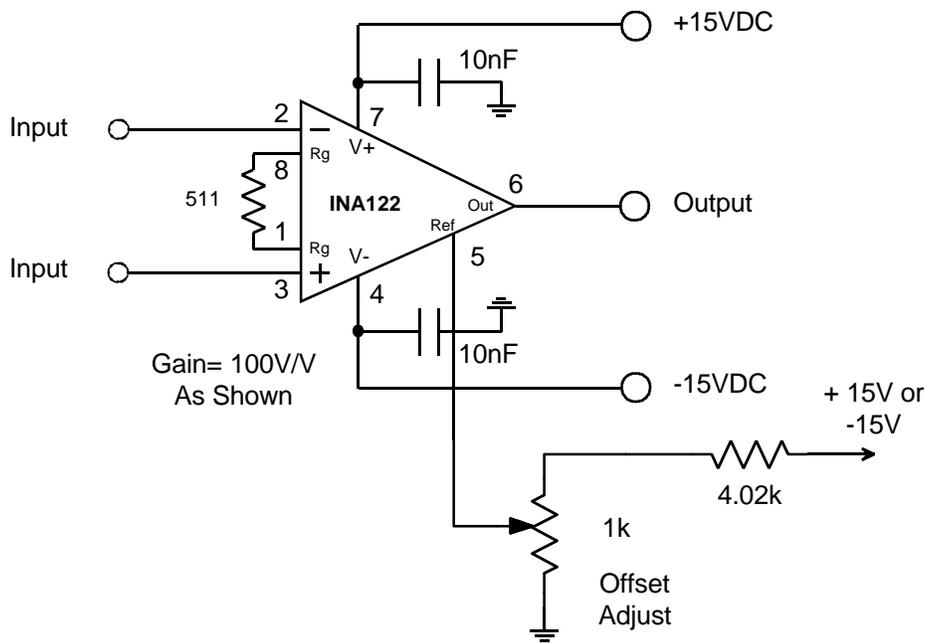
Figure 4- 2. How To Ruin Your CMR By Driving The IA Reference Pin Incorrectly.

To determine how much CMR is lost by the addition of resistance between the reference pin and ground requires knowledge of the amplifier's internal resistor network values. Then an analysis can be performed and a precise answer can be formulated. This is hardly ever worth the effort involved; it is more important to just keep in mind that resistor match is critical to amplifier CMR and any extra resistance at the reference pin will create resistor mismatch.

To appreciate how critical the resistor matching is, refer to Table 1. This illustrates just how close the required IA resistor match must be to achieve different levels of CMR.

20dB	40dB	60dB	80dB	100dB	120dB
1%	0.1%	0.01%	0.001%	0.001%	0.0001%

Table 1. Resistor % Match Required To Achieve CMR.

A far better approach to creating an offset is to insert an op amp buffer between the pot and the reference pin. Negative feedback brings the op amp output impedance down to _well_ under one ohm. This small amount of resistance in series with the IA's internal resistor network will not cause the amplifier' CMR to be adversely effected.
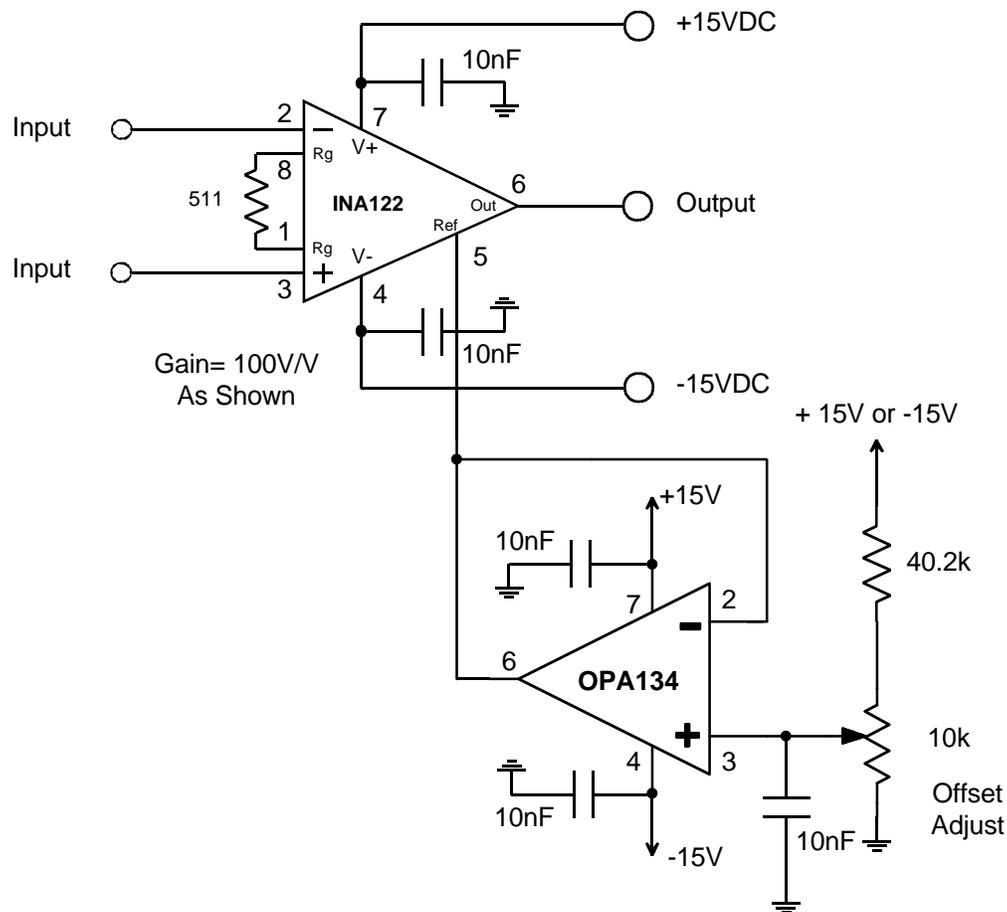
Figure 4- 3. Offsetting An Instrumentation Amplifier By Driving Its Reference Pin Correctly.

Since the op amp non- inverting input has an extremely high input resistance (10^{13} ohms), the resistors can (optionally) be increased in value to minimize supply current. As a further benefit, an RC low- pass filter can be formed by adding a capacitor to the pot wiper output. This will improve power supply noise rejection and assure a low noise reference voltage.

Large offsets can be generated by the circuit shown in Figure 15. If only a few millivolts of offset is required (such as might be required to trim the effect of IA input offset voltage to zero), the Offset Adjust pot can be reduced to 100Ω.

If a large offset voltage with very high stability is required, the 40.2K resistor in Figure 15. can be tied to the output of a + 10V voltage reference (such as a REF102) instead of being connected to either supply. To create a negative voltage reference, a + 10V reference can be inverted by the op amp. A current source (REF200) could also be substituted for the 40.2k resistor. The REF200 current source's versatility allows it to be tied to either supply and thereby generate either a positive or negative voltage on the pot wiper arm.

A Common Mistake: Floating Inputs

Another common mistake made by many first- time IA users is shown in Figure 16. This circuit uses two silicon PIN photodiodes to detect a modulated LED light beam. For best speed of response and linearity, the detectors are reverse biased. An instrumentation amplifier is used to amplify the difference between the two detectors' outputs. Capacitors AC couple the signals to the IA inputs while blocking the detectors' DC bias voltage.

The circuit is missing one thing-- without a bias current return path, the IA bias current creates a huge offset and its output is driven to the rail.

Consider Ohm's Law: The input offset $V_{os} = I_b \times R_{in}$ where I_b is the IA input bias current (1nA) and R_{in} is the IA input resistance (10^{10} ohms) in parallel with the 0.47μF capacitor's insulation resistance (about 30,000 megohm • microfarads for a metallized polypropylene film capacitor).
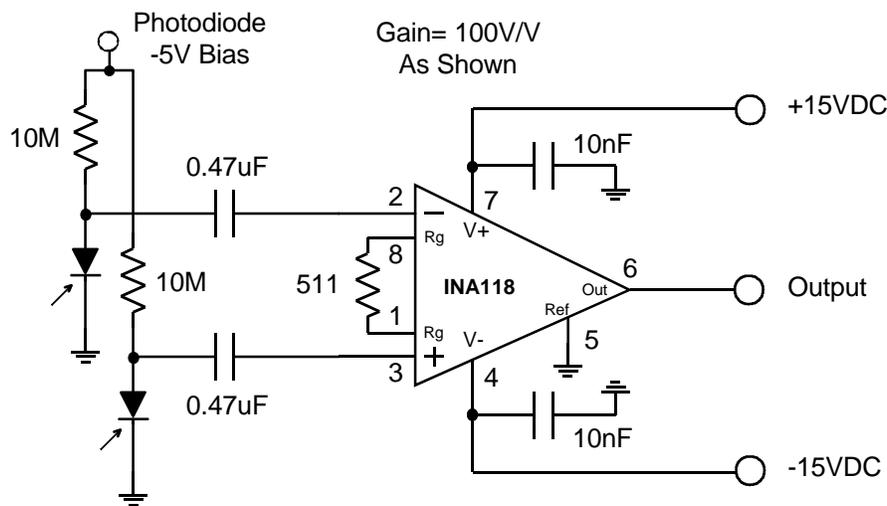
Figure 4- 4. What's Wrong With This Circuit? Hint: Where Does The INA118 Input Bias Current Come From?

Fortunately, this circuit can easily be salvaged. If the two AC coupling capacitors were simply removed and replaced with a jumper, this circuit will be OK. Now the photodiode DC bias voltage will appear on each input, but this is no cause for alarm– the slightly less than -5V bias at each input is well within the INA118's acceptable common mode voltage (CMV) range when it is operating on 15V supplies. See the section on calculating CMV for further details.

The high CMR of the IA will reject the diode DC bias voltage as well as any ambient light common to both detectors.

A second approach (Figure 17.) would be to add a 10 megohm (or higher) resistor from each IA input to ground. This preserves the coupling capacitors' DC blocking function and so the circuit will completely reject detector bias voltage and steady- state ambient light. Rejection of noise from amplitude modulated ambient light--such as the output of fluorescent lamps-- will still depend on the instrumentation amplifier's CMR.
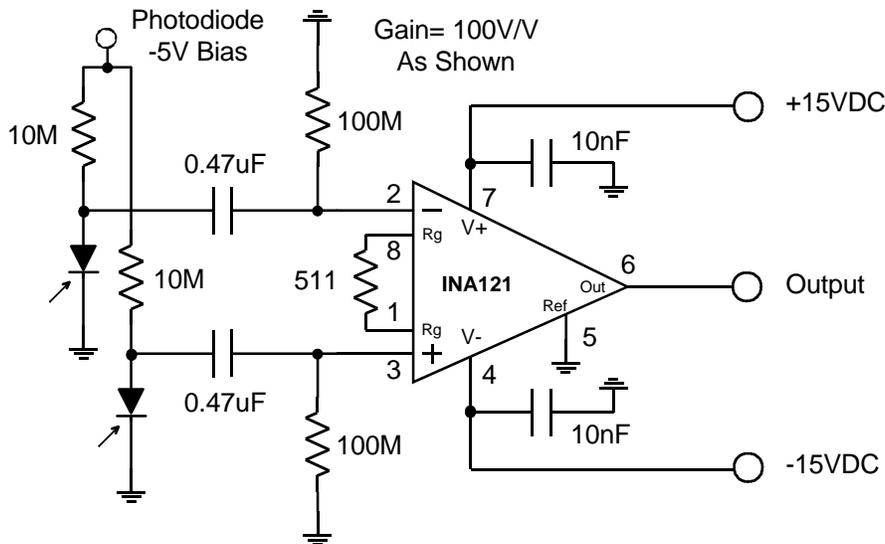
Figure 4- 5. Adding Bias Current Return Resistors Solve The Problem.

Strong power line harmonic output is typical of fluorescent lamps. Modulation at 120Hz and 180Hz is apparent from lamps operated on 60Hz lines. Unexpected optical noise may also be encountered. Solid- state fluorescent lamp ballasts or even Barkhausen oscillation in the tube's plasma discharge sometimes cause high frequency ($\approx$ 50kHz) light output modulation.

To minimize detector loading effects caused by the bias current return resistors, their resistance should be high relative to the detector load resistance (10 megohms in this example) but too-high a value can cause unacceptable input offset voltages due to input bias current. A good solution to this tradeoff is to use a FET- input instrumentation amplifier in place of the bipolar IA used in the original circuit.

An INA121 FET IA will drop into the same socket as the original INA118 bipolar IA, and its input bias current is so low-- 4pA-- that the bias current return resistors may be increased in value to 100 megohms or more, if desired.

Other methods of providing for bias current return are shown in Figure 18. This circuit uses the center tap of the linear variable differential transformer (LVDT) to provide a bias current return path for the IA inputs. Most inductive sources will have a resistance of only a few hundred ohms at most, so this method is effective with virtually any type of IA.

Some sensors may require a specified load resistance (and sometimes a specified capacitance, as well) so its data sheet should be consulted. Any necessary load can be added in parallel with the IA's inputs. Typical input impedance for an instrumentation amplifier is a gigohm or more, so IA source loading will not be a problem!

In some instances, a simple difference amplifier may suffice in a similar circuit. A few hundred ohms pose no problems for a difference amplifier as the center tap will provide an equal source resistance (within a few ohms) to each input. No imbalance occurs, thus the diff amp's CMR is unaffected.
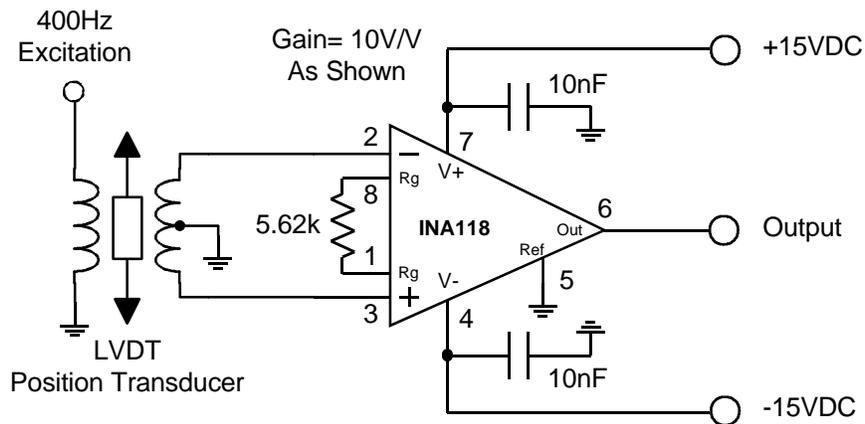
Figure 4-6. Instrumentation Amplifier Input Bias Current Can Also Be Returned To Ground Through An Inductive Source.

Worries about possible detrimental effects of bias current on the sensor magnetic core characteristics can be dismissed. First, the bias currents are low and second, the center-tapped secondary generates an opposing magnetic flux– any net magnetic flux is well below a level that could cause problems in the inductor core.

Those of us who are old enough to remember having to carefully adjust a bias potentiometer to balance a pair of 6L6s or KT-88s in a "push-pull" audio power amplifier can appreciate the fact that opposing currents cancel transformer core magnetic flux– one of the inherent problems with single-ended Class A power amplifiers.

Since Class A amplifiers ran their full plate current through only one winding (the primary) of their output transformer, they required an enormous amount of iron to minimize distortion caused by driving the output transformer core into a nonlinear region of its B-H curve on current peaks. Ah, yes– these old amplifiers are now "high-end" audio designs!

Common-Mode Voltage Limitations

If a designer does not pay careful attention to his choice of instrumentation amplifiers, there may be an unpleasant surprise in store. There are profound external influences on an instrumentation amplifier's operation when a common mode voltage is applied to its inputs. The IA's common mode voltage operating range is influenced by its supply voltage(s), voltage gain, and reference pin voltage.

Operating an IA or diff amp outside of its "envelope" may render it completely useless– the amplifier may have not been damaged, but one or more internal nodes have been driven into a nonlinear region or even into saturation. Amplifier topologies determine their linear envelope and, depending on your particular application, either a 2 or 3 op amp IA may be your best choice.

Common mode input voltage range is the most important consideration in choosing an instrumentation amplifier; other IA specs are secondary. If an amplifier will not function in a particular application, all other considerations are moot.

Determining an instrumentation amplifier's common mode input voltage operating range (its "envelope") is not an easy exercise. A complete circuit analysis must be performed (this can be performed as shown in an earlier chapter) but all of the IA's internal op amps' input and output ranges must be known in order to compute an "envelope." While this may be possible for "do-it-yourself" circuits, most IA vendors do not furnish enough information for an accurate calculation.

Most manufacturers do provide one or two graphs in their data sheets that illustrate the amplifier's CMV operating "envelope" under certain specified operating conditions. This is fine if those conditions represent your particular application but most often they do not. Somehow, we must determine if the amplifiers under consideration will operate properly in our particular application.

Generating a graph of amplifier input CMV range vs. output voltage swing would give us enough information to evaluate the amplifier's operation under the actual operating conditions of our application. This type of graphical representation of CMV range is known as a "Trump Plot[1]."

If all internal op amp node voltage swing limits are known, it is possible (using blood, sweat and tears) to calculate the CMV by hand. But fear not--there are two alternatives to performing this odious task.

A test circuit can be built that allows the amplifier's CMV range to be measured directly. Applying a triangular wave to the IA's differential input while applying a second unrelated triangular wave to the two IA inputs as a common mode signal will allow the amplifier's CMV range to be traced on an oscilloscope. This requires the scope to be driven in an X-Y mode with its X input connected to the amplifier output and its Y input connected to the common mode triangular wave input.

The unsynchronized input signals will generate a family of curves that describe the amplifier's CMV range *under the operating conditions that were imposed on it* . Various supply voltage combinations, amplifier gains, and reference pin voltages can be tried and their effects on CMV range can be quickly visualized and measured.

One cautionary note is in order: even the most advanced digital oscilloscopes seem incapable of properly displaying the family of curves that this method requires. Of all those that were tried, an older Tektronix 7000 series (7834) analog storage scope was found to have the best X-Y display! This is one more reason to keep that "old" analog scope in the corner of your lab. Aliasing is not a problem with analog scopes.

Calculating Common Mode Voltage Range Painlessly

There is an even more painless way of determining an instrumentation amplifier's CMV range– let your computer perform the drudge work of calculating the IA's operating envelope.

A freeware computer program for calculating IA and difference amplifier CMV range can be downloaded from the Burr-Brown Corp Internet website. To use this software, simply download the files and create a new directory (name it something like "CMV_Range") on your hard disk. Place the three files into your newly- created directory and that's all there is to it. Here is what your directory should contain:
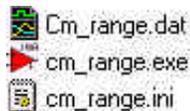
The executable program, cm_range.exe, runs under Microsoft Windows and graphically represents the IA's CMV operating envelop by overlaying plots of each of the IA's internal op amp input and output nodes. Thus both the instrumentation amplifier's overall CMV range is displayed as well as showing where the CMV limitation originates inside the IA. A pull-down menu is provided to select a difference or instrumentation amplifier part number, its supply voltages, voltage gain, and reference voltage.

Positive and negative power supply voltages are entered separately in order to display the effects of asymmetrical supply voltages on the amplifier's CMV range. If the voltage on the amplifier's V- pin is negative, enter that voltage with a negative sign, e.g., -15. Similarly, enter the voltage on the amplifier's reference pin. Lastly, enter the amplifier's voltage gain. If the voltage gain you have entered is lower than its minimum specified gain, the gain is automatically defaulted to that value.

A graph ("Trump plot") of the amplifier's common mode voltage range is calculated and displayed on screen. An example of a "Trump plot" of an INA118 instrumentation amplifier is shown in Figure 19. The selected operating conditions are indicated at the top of the graph: +Vs= 15.00, -Vs= -15.00, Vref= 0.00, and G= 100.00.

Node limits for these operating conditions are calculated and plotted for each of the INA118's three internal op amps. Designations: A1 & A2 are the input op amps and A3 is the output difference amplifier. Overlaying the input and output limits of all three internal op amps allow the overall INA118 common mode voltage range to be clearly seen.

Proper operation is assured if the amplifier is operated anywhere within the large open white area of the plot. Due to internal node limitations, an amplifier cannot be operated outside of its proper operating envelope; that is, within a hatched area of the plot. Under certain conditions, operating an instrumentation amplifier outside of its proper operating envelope can have undesirable consequences– the IA output can not only be inaccurate, it can also have an inverted polarity! See page 40 for a discussion of this behavior.

To illustrate the effect of IA gain on its CMV range, compare Figure 19 (a.) with Figure 19 (b.) where an INA118 is operated under identical conditions except for gain. Reducing the IA gain from 100V/V (a.) to 1V/V (b.) noticeably shrinks its positive CMV range. Careful inspection of the hatching in this region reveals this CMV limitation to be the result of one particular node: the input of A2.

This plot is only applicable to an INA118 operating under the conditions that have been specified. Each different IA or difference amplifier will exhibit its own unique "Trump plot".
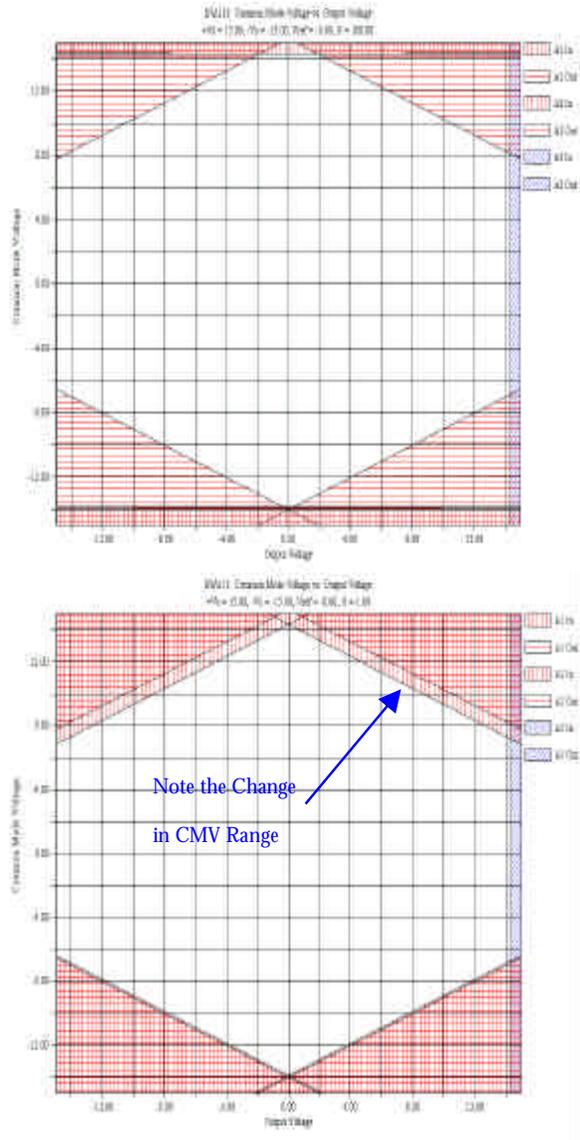
Figure 4- 11. CMV Range Plot Reveals the Effects of IA Gain. (a.) 100V/V, (b.) 1V/V. (INA118 with ±15VDC Supplies)

Three op amp IAs have distinctive CMV plot shapes that are clearly distinguished from those of two op amp IAs. Notice the very different "Trump plot" envelopes of a three op amp INA118 (Figure 20 (a.)) and a two op amp INA122 (Figure 20 (b.)) when both are operated under the same conditions that were shown in Figure 19 (a.).

Three op amp instrumentation amplifiers generate a rhombic envelope (a.) while two op amp IAs generate a very different- looking trapezoidal envelope (b.). This difference in CMV range shape can be used to great advantage for tailoring a CMV range to fit single supply (set V-= 0) and low voltage bipolar supply applications.
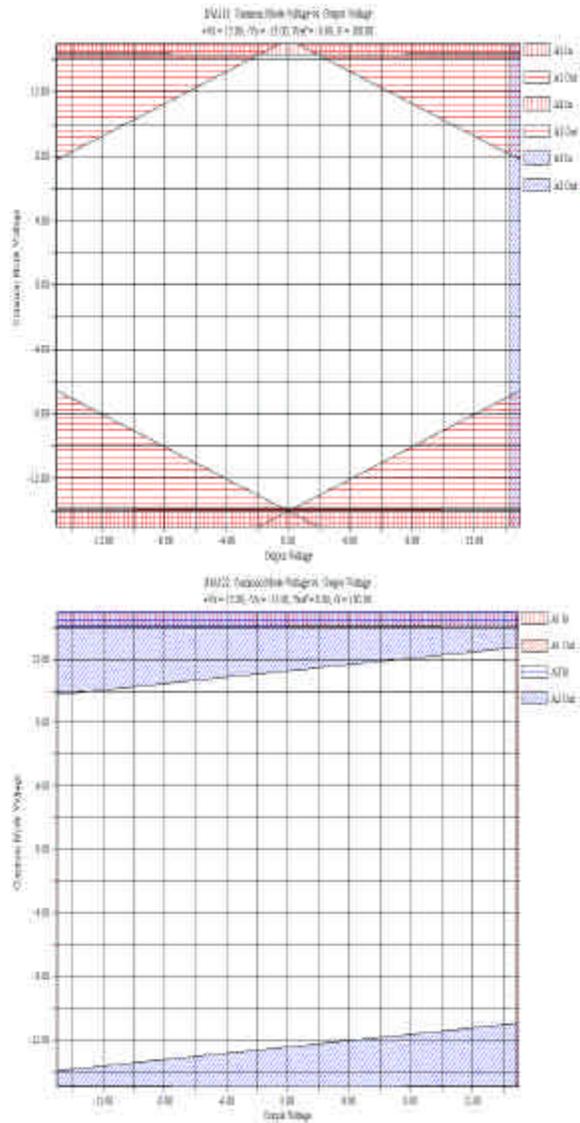
Figure 4- 12. Distinctive CMV Range Shapes. (a.) Three Op Amp IA, (b.) Two Op Amp IA.

Shifting an amplifier's CMV range to suit a particular application can also be accomplished by driving the amplifier's reference pin with a low impedance (!) voltage source. Two and three op amp IAs respond differently to reference pin offsetting and this can be quickly visualized by the computer program. Note the envelope shifts of Figure 21 (a. & b.) with 10VDC applied to both an INA118 and an INA122 compared with the grounded reference pins of Figure 20 (a. & b.).

Although the distinctive envelope shapes remain, they are shifted to the right in both plots. This greatly improves the three op amp IA input CMV range for positive output swings. Conversely, the two op amp IA CMV is little affected.
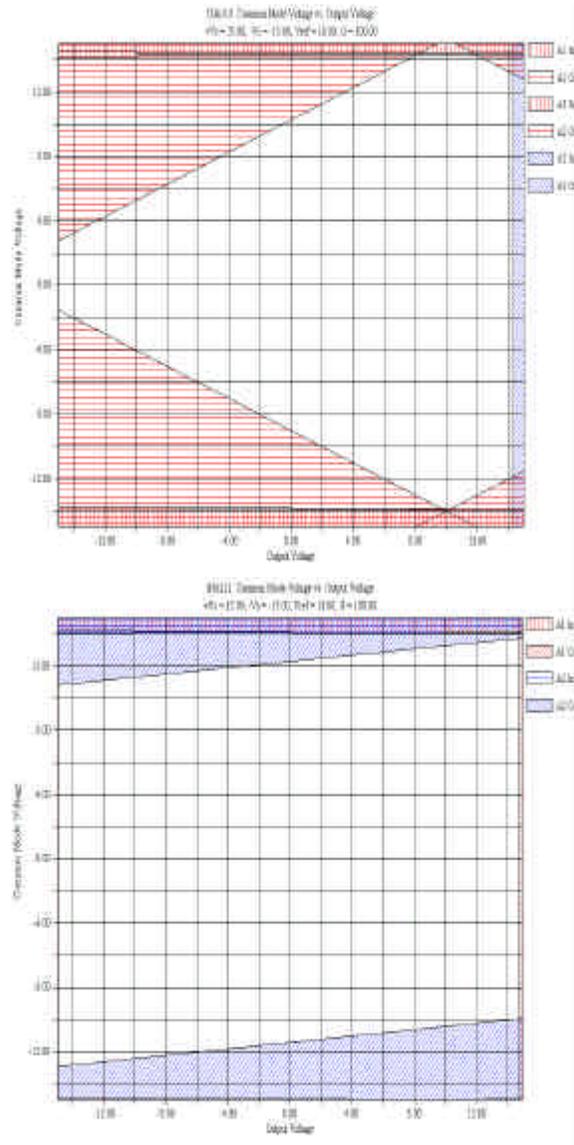
Figure 4– 13. Ten Volts Applied To the Reference Pin Modifies CMV Range. (a.) Three Op Amp IA, (b.) Two Op Amp IA.

Instrumentation and difference amplifier supply voltages will also alter their CMV operating envelopes. Shifting the envelope by employing asymmetrical power supplies can be a useful

technique to match an amplifier's CMV range to an application. "Asymmetrical" power supplies refers to using unequal voltages on the amplifier's V+ and V- pins, such as V+= 20V and V-= 10V.

Single supply operation is also an example of using asymmetrical power supplies. Plotting CMV envelopes for two and three op amp IAs operating on a single +5VDC supply will reveal the ***inherent superiority of the two op amp topology in single supply applications.***

Low voltage bipolar supply applications should likewise be approached with caution. Choose the wring IA topology and things can get ugly (Figure 22.) Operating on ±1.35 volt supplies illustrates this dramatically. A three op amp INA118 would be a poor choice for this application due to its small odd- shaped CMV range while the two op amp INA122 still offers a useable envelope. Choose carefully!
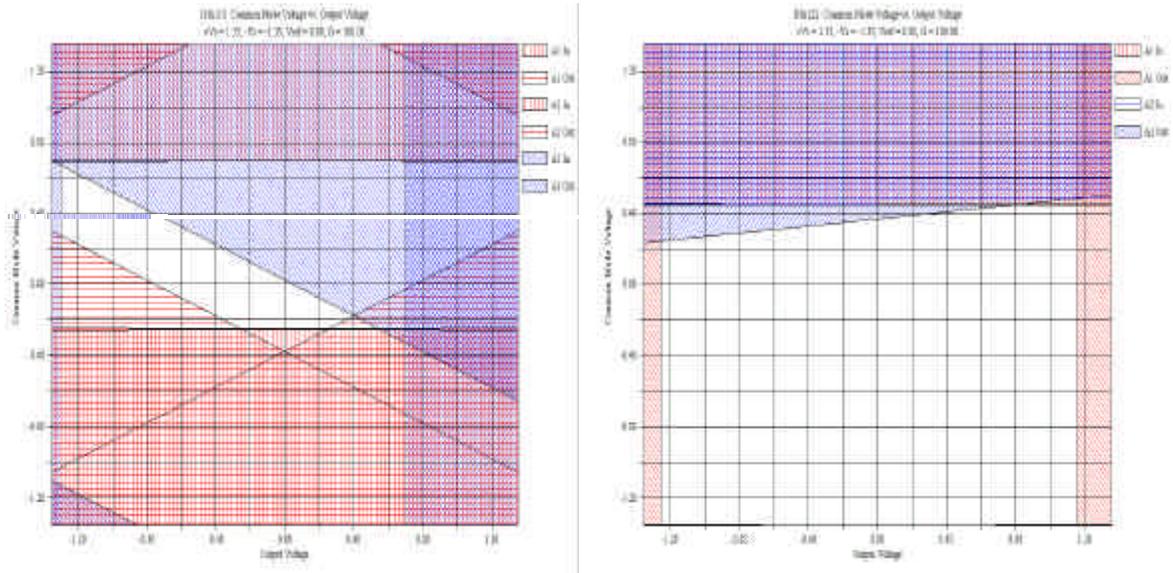
Figure 4- 14. Low Supply Voltages Must Be Used With Caution! (a.) Unusable Three Op Amp CMV Range, (b.) A Larger Two Op Amp CMV Range.

Maintaining "Truth in Output"

There is a very tricky characteristic of three op amp instrumentation amplifiers that users should be aware of. The reason for this characteristic is its topology, so it affects every three op amp IA on the market... if you **exceed** the output swing range of either of the two input amps, you lose all sense of "which way the signal should go" at the output of the third amplifier (the IA output). This phase- ambiguous behavior can be predicted by inspection of IA CMV plots as described in the section "Common- Mode Voltage Limitations" above and in many 3 op amp IA data sheets.

The solution, of course, is to limit the amplifier's input, gain, and CMV to keep its internal nodes from reaching their voltage swing limit. In some systems, it may be possible to monitor the two input op amp outputs[1] with a window comparator made up of two op amps & a handful of resistors. By sensing the two input op amp outputs, the user has at least an indication of whether the instrumentation amplifier output can be trusted.

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1.   These internal nodes are accessible on three op amp IAs such as the Burr- Brown INA101, INA103, and INA115, as well as on programmable gain amplifiers PGA205 & 205 and PGA206 & 207.

Phase reversal in applications such as a servo amplifier can have serious consequences. Oscillation caused by inadvertent positive feedback in a servo system driving a many-ton structure can be dangerous and destructive. Be safe-- carefully check each active component in a closed-loop control system for the possibility of phase inversion under overdriven conditions.

Noise Filtering the IA Input the Wrong Way—Making A Bad Situation Worse

If a signal input line to an instrumentation amplifier is exposed to high levels of radio-frequency interference (RFI), the resulting conducted RFI can cause unexpected problems: DC offset voltage, offset voltage "drift" or inexplicable "jumps" in its DC output.

In many cases, blame for an unstable circuit is placed on its input amplifier when, in fact, that amplifier is not really the cause of the problem. If high amplitude high frequency interference is induced into the cabling between a sensor and a low level amplifier input– the external wiring acting like an antenna– the IA inputs can rectify and generate small DC shifts in the amplifier's operating point. Thus even low bandwidth devices can be adversely affected by signals well into hundreds of megahertz. Once the RF is rectified, it appears as DC. If the offending RF signal is amplitude modulated, its modulation envelope also appears on the amplifier's output.

If RF rectification is suspected, connecting an audio amplifier to the IA output can sometimes identify the interfering RF source. Don't be surprised if you hear a local AM radio station coming through loud and clear! This trick also works well for identifying RFI from TV stations but not for FM broadcast stations. Radar interference can be identified by its buzz.

Some types of IA topologies are less sensitive to RFI rectification than others. The forward-biased emitter-base junction of a bipolar transistor is an efficient RF detector while a reverse biased gate-source junction of a JFET is a relatively poor detector. This is because the JFET gate-source diode must be driven by a large amplitude signal before it can conduct in the forward direction (rectify).

A simple rule of thumb:

1. **JFET- input amplifiers are best in severe RFI applications.**

2. **Filter all input, output, and power leads.**

3. **Shield everything.**

The most effective single thing that can be done to eliminate RFI is to employ a low pass filter at the instrumentation amplifier inputs. Keeping RF out of the instrumentation amplifier inputs is half the battle toward solving conducted RFI problems but it must be done correctly.

Simply adding an RC low pass filter (LPF) on each IA input (Figure 21.) seems at first to be the correct approach. If the RC LPF pole frequencies are low enough, the amplitude of an interfering RF signal will be greatly reduced at the IA inputs and the problem is solved– right?

Unfortunately, this is wrong. There is a limit to how low a pole frequency an input LPF can employ without affecting the IA differential signal bandwidth. If the offending RFI is much higher in frequency than the required signal bandwidth, this approach can work but in order for this approach to succeed the time constant (LPF pole frequency) of R1 &C1 must *perfectly* match that of R2 & C2.
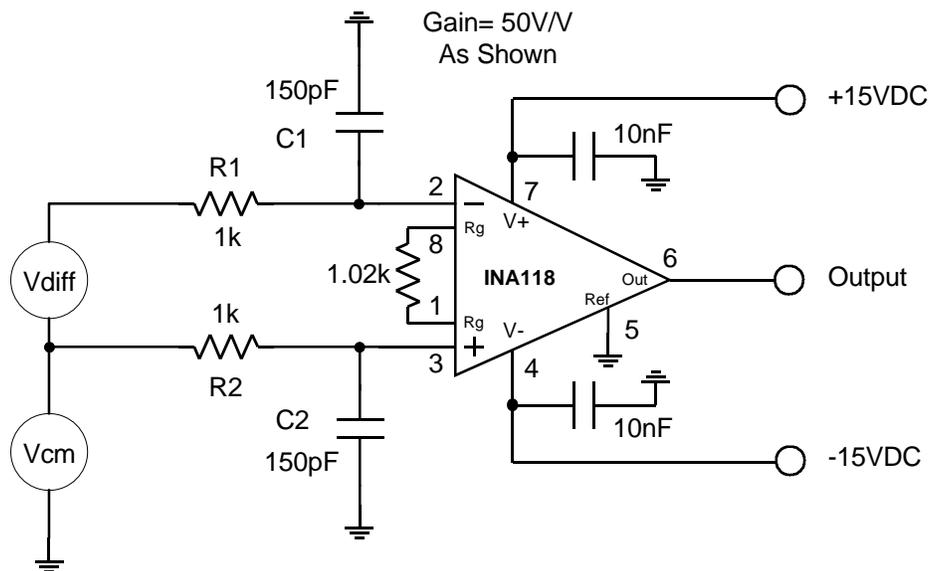
Figure 4- 15. Noise and RFI Filtering– The Wrong Way.

Consider what happens when real- world component tolerances are used in an RFI filter. With 1% resistors and 5% capacitors, there can be worst- case RFI filter component mismatches as shown in Figure 24. Here R1 and C1 are at their upper tolerance limit while R2 and C2 are at their lower tolerance limit. This results in a lower LPF pole frequency in the inverting input than in the non- inverting input. It is this mismatch that causes serious problems.
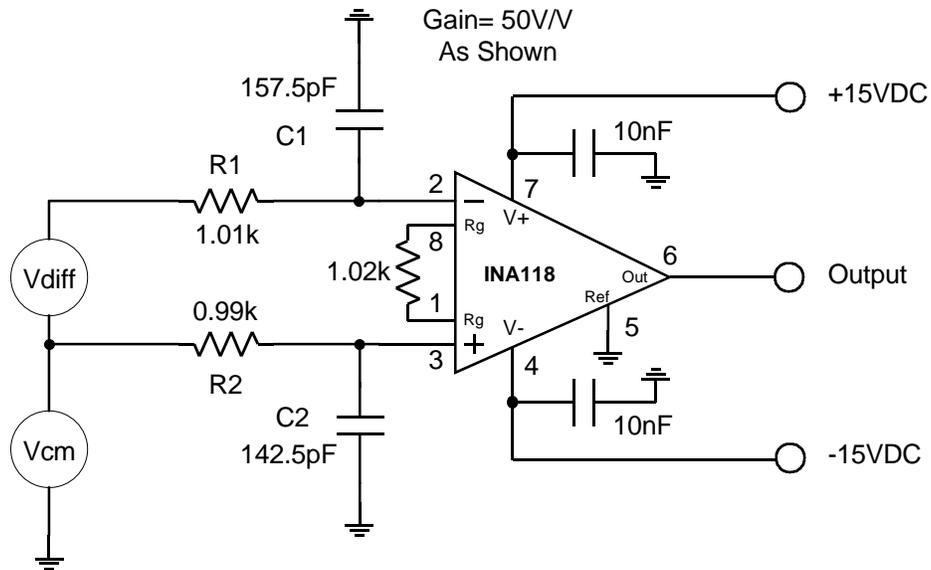
Figure 4‑16. Worst‑Case Mismatched‑Pole RFI Filters With 1% Resistors and 5% Capacitors.

Figure 25 illustrates how the filter mismatch problem occurs. In this graph, a $1V_{RMS}$ common mode signal (V_{cm}) is swept from 100kHz to 1GHz and the amplitude of the signals at the inverting and non‑inverting inputs are is shown. Because the two filter pole frequencies (time constants) are not matched, the input RFI filters then create a differential signal (Figure 26.) from the common mode signal. Thus we have made the RFI problem worse– even an IA with infinite CMR cannot reject the interfering common signal because now a portion of it has been converted to a differential signal. Thus it is now amplified along with the desired input signal.
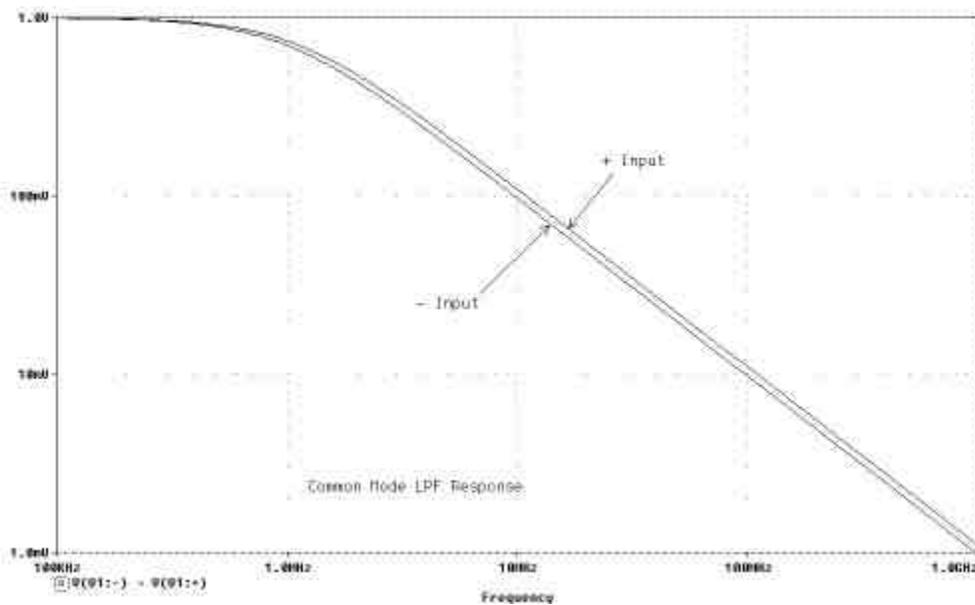
Figure 4‑17. Mismatched Noise Filter Components Create Mismatched CMV Low Pass Filter Poles.
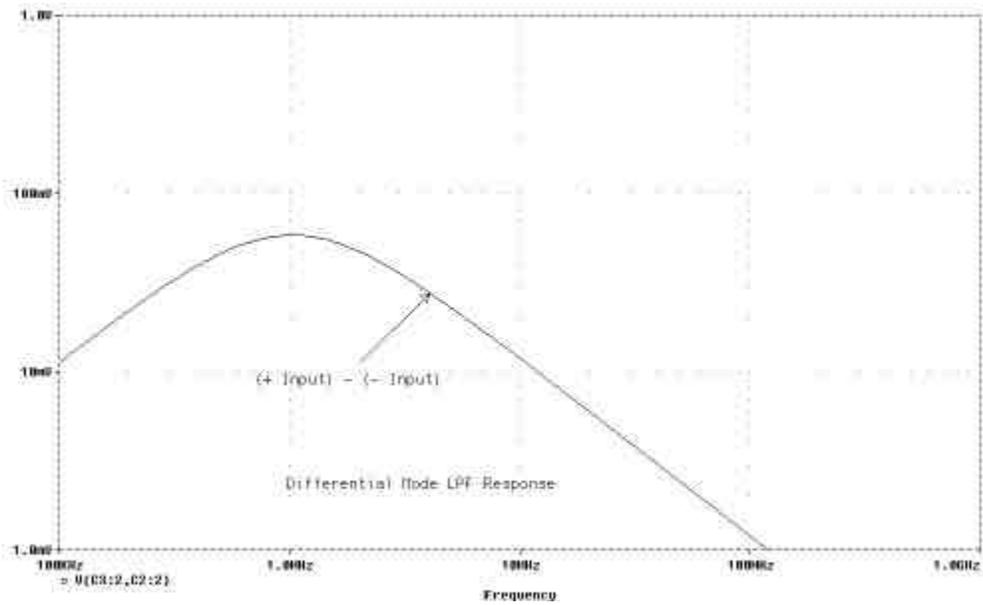
Figure 4- 18. Differential Voltage Created By Mismatched Common Mode Low Pass Filter Poles.

The frequency response of the circuit in Figure 24. is shown in Figure 27. The INA118 response (upper curve) quickly rolls off above about 1MHz because of two factors: the amplifier's gain-bandwidth limitation (about 150kHz BW in a gain of 50V/V) and the RFI filters' pole locations. The lower curve shows the differential input signal that has been created by RFI input filter pole mismatch.
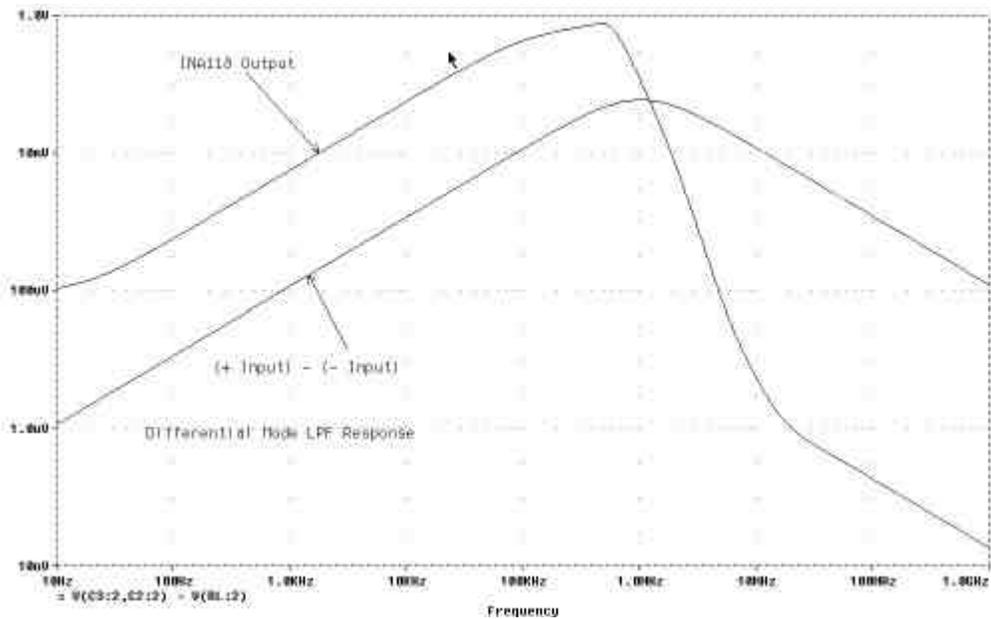
Figure 4-19. CMV Frequency Response of INA118 Instrumentation Amplifier with Mismatched Common Mode Low Pass Filter Poles.

If a passive RC filter is used, the effect of bias current in the resistors must be considered. To minimize input offset voltage, high resistance RC filters will require a very low input bias current instrumentation amplifier such as a FET- input INA121.

If C1 is a trimmer capacitor and C2 is a fixed capacitor (or vice- versa), the two pole frequencies can be tuned to match. For easier tuning, try a fixed capacitor in parallel with a smaller trimmer capacitor. Whichever approach is used, all of the RFI input filter capacitors should have a low temperature coefficient. Polystyrene film, polypropylene film, or COG (NPO) ceramic dielectric types are recommended.

To RFI filter IAs with higher bias current, substitute inductors for the resistors. An RC filter will have low resistance at DC, but its RF impedance will be high if a suitable choice of inductance is made. Be aware that 1% tolerance fixed inductors are not inexpensive. Of course, a variable inductor LC filter can be used in one input which allows it to "tune" its pole frequency to that of a fixed inductor LC filter in the other input. Low TC filter components are needed to maintain the pole matching over temperature.

There is a better way of passive filtering that is not as critical to filter component matching. In addition, it offers better differential- mode filtering than the simple filter of Figure 23. This input RFI filter is shown in Figure 28.
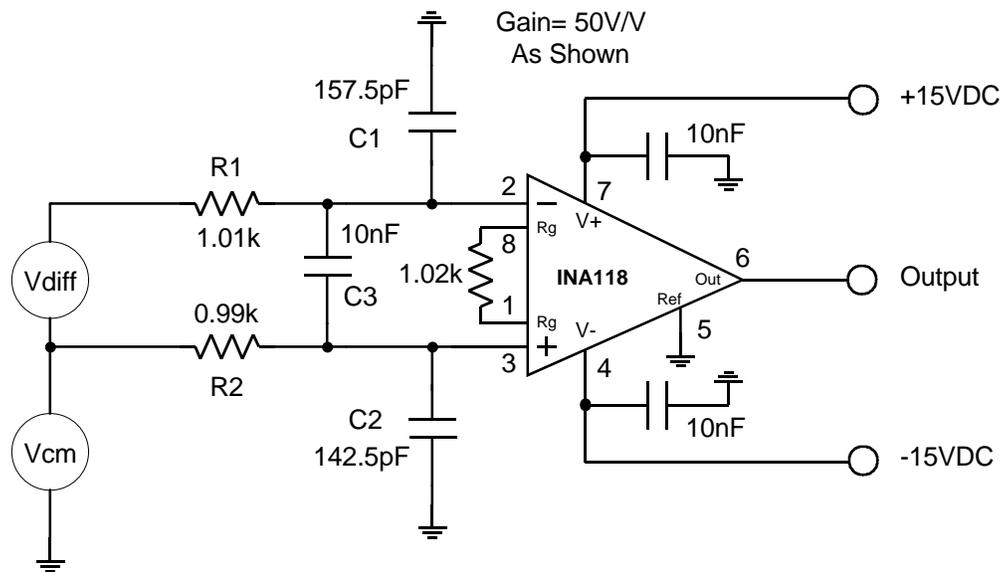
Figure 4- 20. An Improved Method of Instrumentation Amplifier Input Noise and RFI Filtering.

Compare the frequency response of this input filter and instrumentation amplifier (Figure 29.) to the other shown in Figure 27. Even though the R1C1 and R2C2 mismatch remain the same, there is less common mode voltage to differential voltage conversion with this filter.
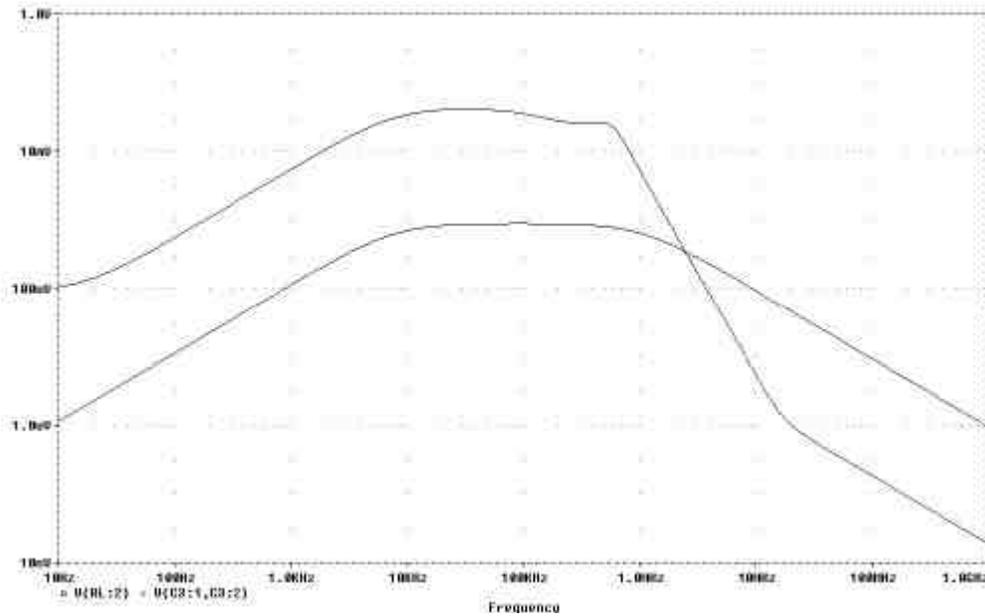
Figure 4- 21. Frequency Response of INA118 Instrumentation Amplifier with Mismatched "Improved" Noise Rejection Filter.

Now that you have been warned not to do it, I'll confess that input filtering can be done successfully if the required signal bandwidth is low. If the input filter RC pole is very low, the differential error signal caused by input filter RC pole mismatch and the higher frequency common mode signal will be so far down the attenuation rolloff curve that it will contribute little to the CMR error.

Application Circuits: General

A +/-20V Input Diff Amp With +/- 200V CMV Applications Circuit
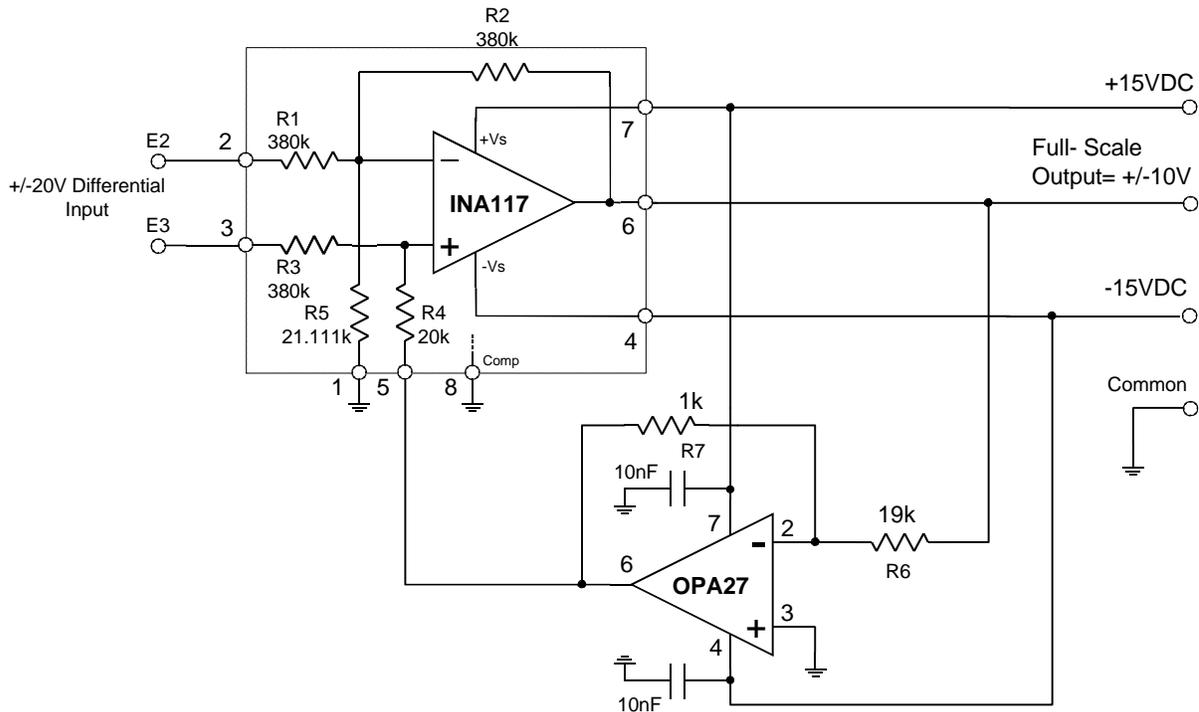
Figure 6-1. An External Op Amp Boosts The INA117 Difference Input Range To +/-20V But Still Handles +/-200 V Common Mode Voltages.

The INA117 monolithic difference amplifier, can accept +/-10V differential input signals with up to +/-200V common-mode, though it operates from standard +/-15V power supplies. Many applications require an amplifier with both the +/-200V common-mode capability and a larger differential input range. An applications circuit showing how to extend the differential input range to +/-20V is shown in Figure 13 above.

The high common mode rejection capability of the INA117 results from the roughly 20-to-1 resistor dividers internally supplied on the inputs of the op amp (see Figure. 13) With that attenuation, a +/-200-V common mode signal is reduced to +/-10 V at the op amp's two inputs. This arrangement rejects the common-mode signal, but passes differential signals at unity gain.

Appropriate resistors in the INA117's internal op amp circuit set the diffamp's gain independently of its common mode rejection ability. But for the gain to remain stable with temperature changes, the ratios of R_1/R_2 and R_3/R_4 must track with ratio $R_1/(R_2$ in parallel with R_3). A precision laser-trimmed metal film resistor network on-chip achieves excellent matching and TC tracking.

Building a similar circuit from a handful of discrete resistors and an op amp is possible but achieving the necessary accuracy from non- matching resistors is difficult at best. The original INA117 circuit, however, limits its differential input range to +/- 10V only because it has unity gain. Its +/-15V power supplies limit the output swing.

Reducing the INA117's gain would increase its differential input range– for example, reducing the gain to 0.5V/V would increase the circuit's differential input range to +/-20V. Reducing the gain just with external resistors may seem like a simple approach, but an external op amp (0PA27) circuit for reducing the gain is a superior method. This preserves the INA117's extremely precise internal-resistor matching, so the circuit's common mode rejection and its drift with temperature remain unchanged. Furthermore, the gain reduction produced by the external op amp circuit actually improves output noise. It would remain unchanged with the simpler resistor- only approach.

Inverting the diff amp output with the 0PA27 and feeding a small amount of its output signal back to its reference pin (pin 5) reduces the diff amp's voltage gain. Even with the added OPA27 op amp in the feedback path, the stability of the resulting circuit is excellent (Figure 14).
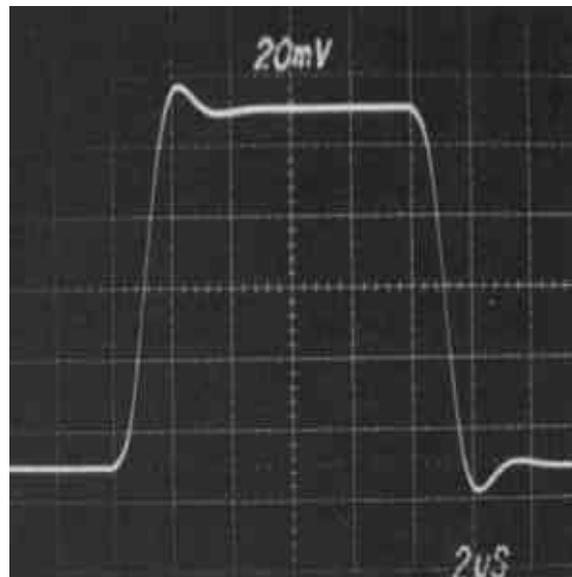
Figure 6- 2. Even With The 0PA27 Op Amp feedback circuit gain of 0.5V/V and l000pF load, the stability of the INA117 circuit is excellent

To better understand the circuit's operation, consider the INA117 to be a four-input device where E_2 is the signal at pin 2; E_3 at pin 3; E_5 at pin 5; and so forth. The output voltage is: $E_0 = E_3 - E_2 + 19E_5 - 18E_1$. With E_1 grounded (equal to 0 V) the reduced differential gain is: $A = 1/[1+19/(R_6/R_7)]$ and for $A = 0.5$, $R_6/R_7 = 19$.

Because of the low output impedance of the 0PA27 circuit, the impedance at pin 5 of the INA117 is low and consequently, the INA117's critical resistor matching, gain, and common-mode rejection are preserved.

To adjust the common-mode rejection for critical applications, add a 10 ohm fixed resistor in series with pin 5 and a 20 ohm variable resistor in series with pin 1. Short pins 2 and 3 together and drive them with a 500Hz square wave. Using a square wave instead of a sine wave test signal allows the AC signal to settle out and makes the DC CMR easier to observe on an oscilloscope and adjust. At high scope gain, trim the circuit to minimum output with the 20 ohm variable resistor. This trimming of the CMR may change the gain slightly. If it does, then adjust the R_6/R_7 ratio to adjust gain. This adjustment will not affect the CMR.

Two and Three Op Amp IA Applications

Show where a two op amp IA would be most suitable for wider CMV range or single supply and another application where a three op amp IA would give better CMR.

Difference Amplifiers In Single-Ended Applications

Although originally intended for differential amplification, commercially available monolithic diff amps can be used for interesting and useful single ended gain applications. Most of these applications take advantage of their on-chip precision resistor network to give precise and stable gains.

Connecting a difference amplifier as shown in Figure 15 (a.) forms a precision inverter (a gain of -1.000V/V). The input and feedback resistors are carefully laser trimmed to obtain a high CMR in the original design but here we are using this precise ratio to generate a very precise and stable gain.

A simple wiring change shown in Figure 15 (b.) and the circuit's voltage gain becomes +2.000V/V. The same ratio resistors which set the inverting gain are now used to set the non-inverting gain. The parallel resistors in series with the input and the op amp non-inverting gain are of no consequence as far as the signal is concerned—they are in series with an input impedance that is on the order of 10^9 ohms, so no error is generated by their presence. These resistors do generate a small offset voltage due to that input's bias current flowing through them but it is beneficial as it compensates the offset generated at the other op amp input by its bias current.
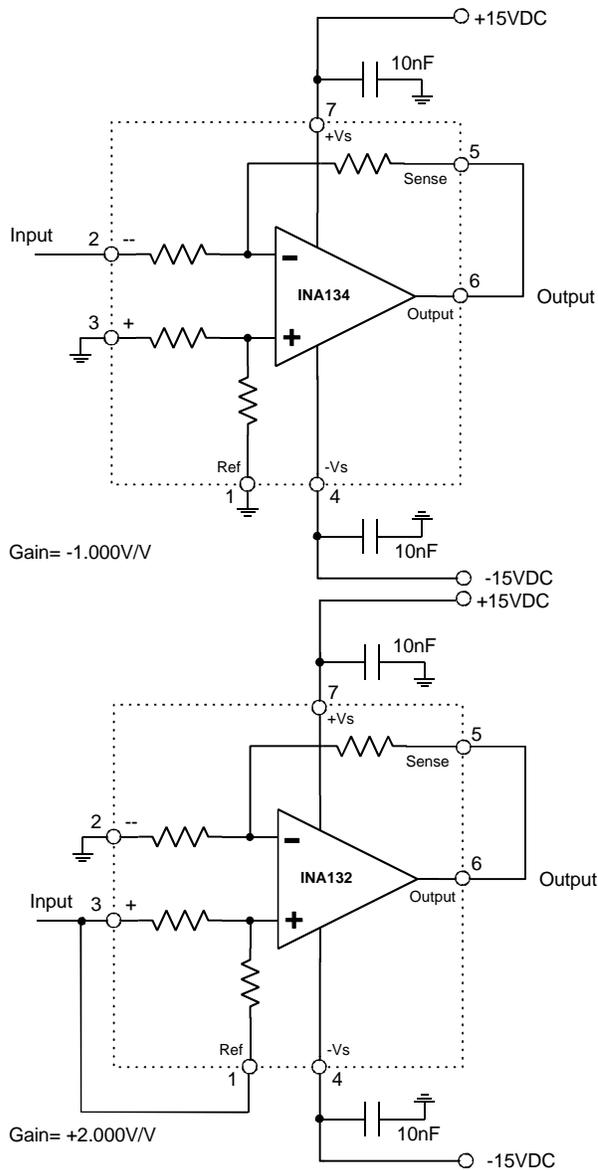
Figure 6- 3.   (a.) Precision Gain Of -1.000V/V Amplifier.   (b.) Precision Gain Of +2.000V/V Amplifier.

Rearranging the difference amplifier connections once again (the circuit of Figure 16. (a.) yields an average value amplifier. Precise resistor ratios now form a voltage divider between two inputs. The op amp serves as a high- impedance unity- gain buffer for this divider.

Connecting the feedback resistor for an op amp gain of 2V/V in Figure 16. (b.) compensates for the voltage divider ratio of ½ and converts the difference amplifier into a 2- input summing amplifier.
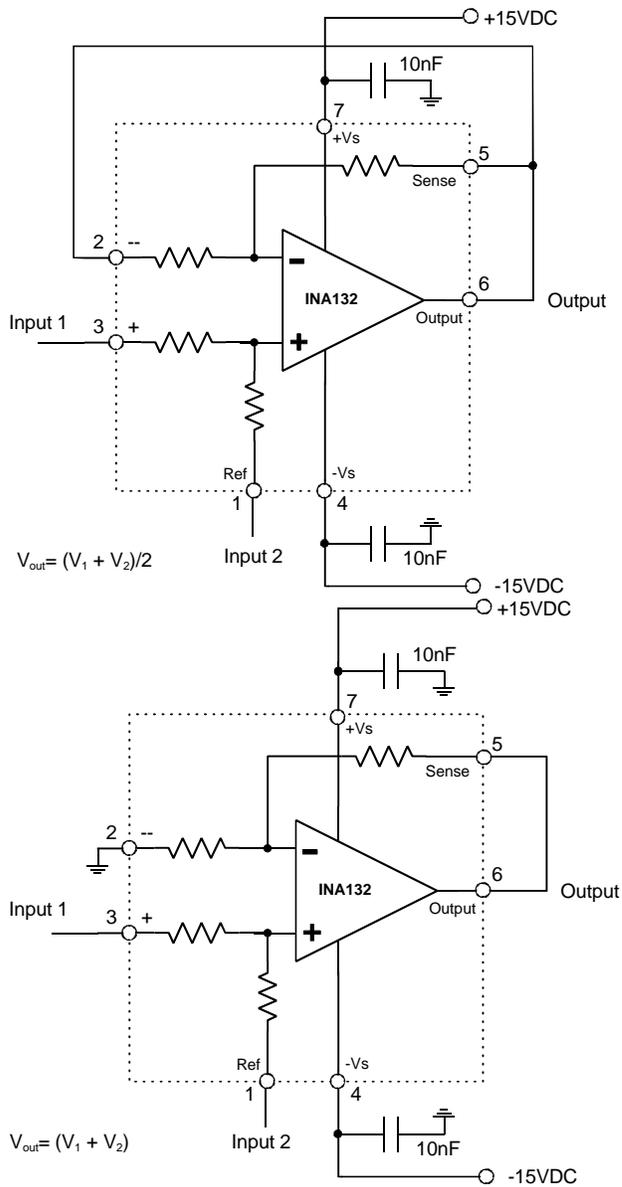
Figure 6-4. A Difference Amplifier Connected As (a.) An Average Value Amplifier.   (b.) A 2-Input Summing Amplifier.

Using the average value circuit of Figure 16 (a.) with one input grounded gives us an amplifier with a precision gain of +0.500V/V. This variant is shown in Figure 17.
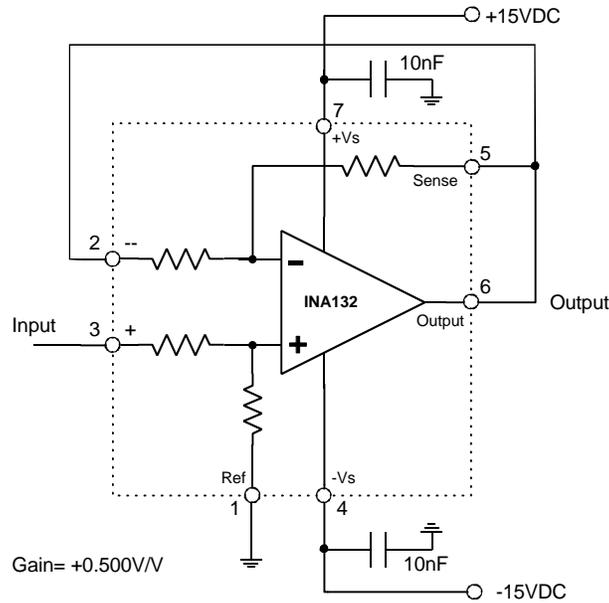
Figure 6-5. A Precision Gain Of +0.500V/V.

By connecting the inputs of two INA134 unity gain difference amplifiers together in an anti-parallel connection, their output signals then appear 180 degrees out of phase. In other words, this is a differential-input to differential-output amplifier (Figure 18.). It can also be used as a single-ended-input to differential-output amplifier if one input is grounded.

In this amplifier, because one output swings positive while the other swings negative, two benefits accrue: we have doubled the output voltage swing to +/-20V and also doubled the amplifiers' slew rate from 14V/us to 28V/us.
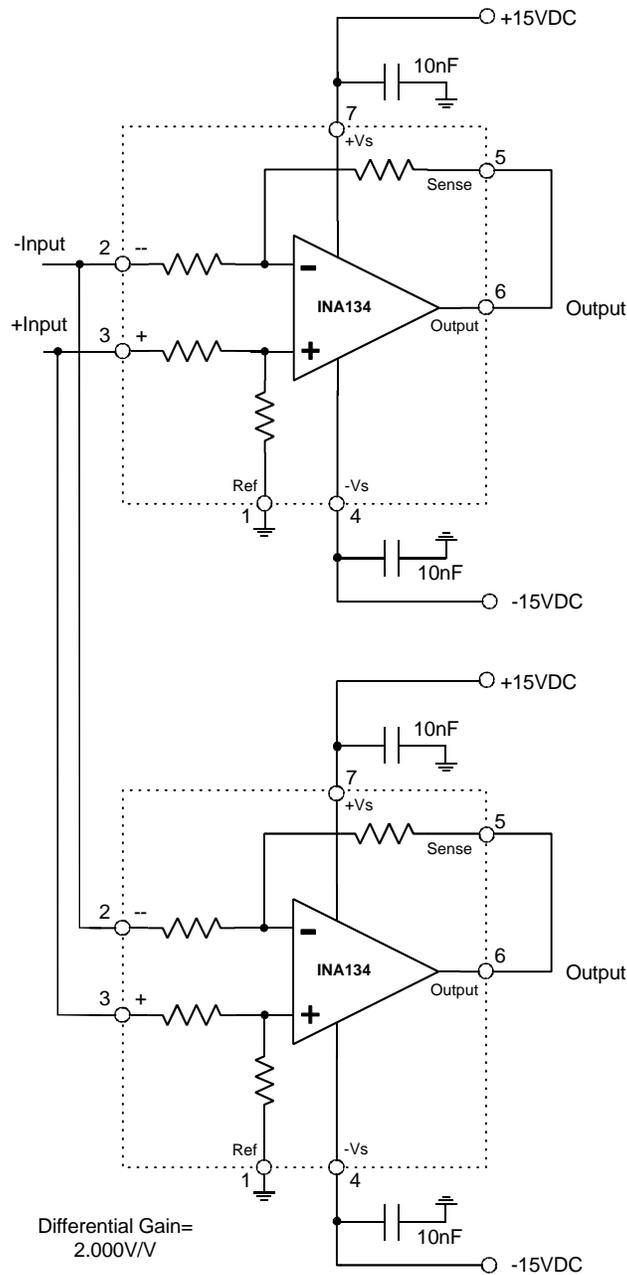
Figure 6-6. A Differential-Input/ Differential-Output Amplifier.

With the addition of a potentiometer, a difference amplifier can be transformed into an amplifier with continuously variable gain (i.e., no switching "steps") from -1.000V/V to +1.000V/V.

Referring to Figure 19, consider the amplifier's gain path with the potentiometer in the fully counter-clockwise (ccw) position. The inverting input is tied to ground and there are equal value resistors in the op amp's negative feedback and input path. This can be recognized as an ordinary inverting amplifier with a gain of -1V/V.

When the potentiometer is rotated to its fully clockwise position (cw), the circuit's operation is not so intuitively obvious. Now the op amp's non-inverting input is connected to the signal instead of ground as in the previous condition. The negative feedback path through the resistors is still in place, but—in this case—there is no current flow through those resistors.

Op amp theory dictates that the op amp's inverting input is driven by feedback to be equal to its non-inverting input. Since there is no voltage drop in the two parallel resistors in the diff amp's + input, the input signal voltage appears on the non-inverting input. Feedback forces the inverting input to also be equal to the input signal voltage. As there is an equal voltage (signal) across the - input resistor, the current flow through it is zero. An ideal op amp has no current flow into its inputs, so the current flow through the negative feedback resistor is also zero. Without any current flowing in this negative feedback circuit, it is just "going along for the ride" and contributing no gain whatsoever. Therefore, this connection becomes a simple voltage follower with a gain of +1V/V.

As we have seen, there is a -1V/V gain connection and a +1V/V gain connection that can be selected by the wiper position of the pot. In fact, if we center the wiper position on the potentiometer, the amplifier's -1V/V gain will exactly cancel its +1V/V gain and the resulting amplifier will have a gain of zero—no output.

Rotating the potentiometer will change the ratio of inverting and non-inverting gain and thus give us a continuously adjustable gain. This circuit can be combined with a precision +10VDC reference such as a REF101 to make a useful general purpose reference circuit that can be adjusted to any output voltage between -10V to +10V; a truly "universal" voltage reference!
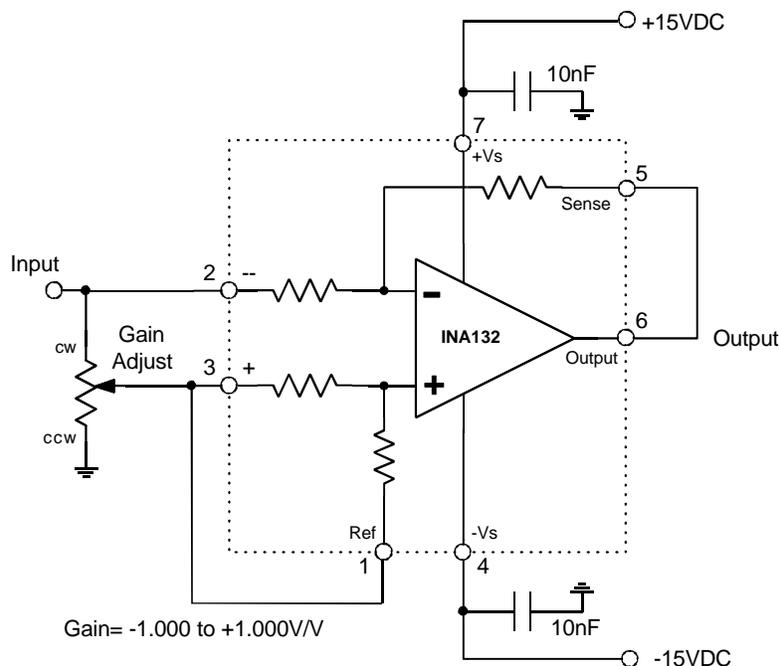
Figure 6-7. An Amplifier With A Continuously Adjustable Gain Range Of -1.000V/V To +1.000V/V

If the potentiometer in Figure 19. is replaced by a switch, we can perform the electronic equivalent of driving the pot from one end to the other. This yields a gain of +/-1V/V amplifier circuit of Figure 20. This amplifier circuit forms a synchronous detector (also known as a phase sensitive detector), which is a powerful tool for signal detection. Extracting weak signals from noise is its forte.
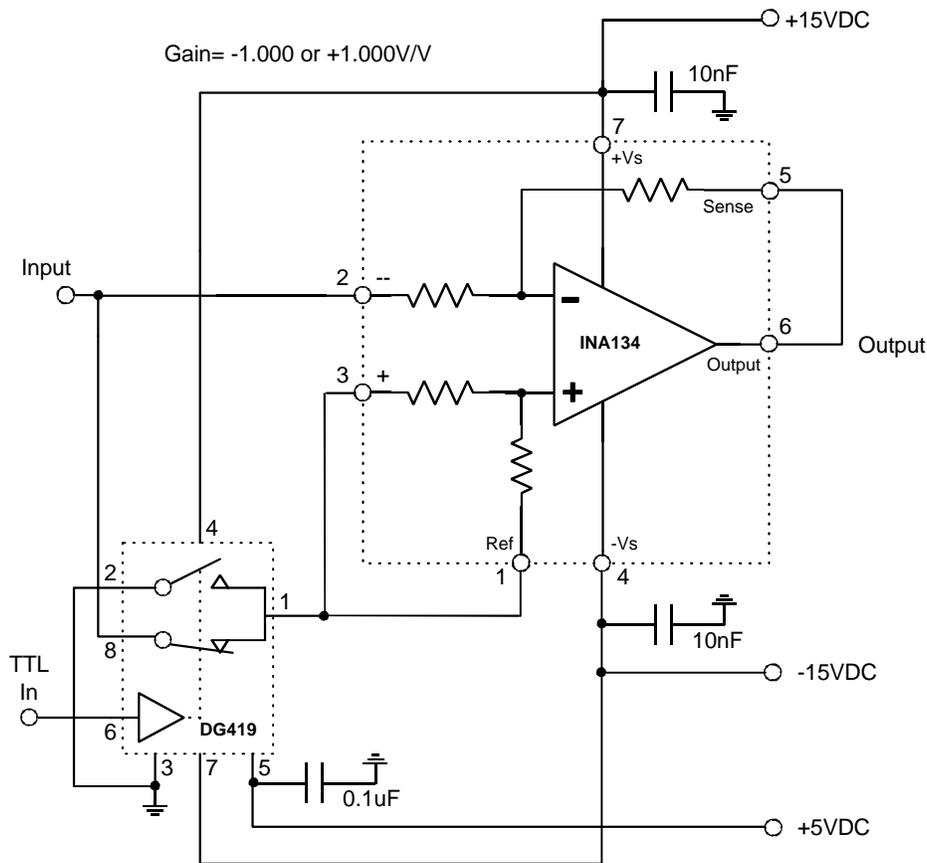
Figure 6-8. Adding A Switch To A Difference Amplifier Creates Turns It Into A Synchronous Detector, a.k.a. Phase Sensitive Detector.

Accurate carrier suppression requires the diff amp input offset voltage to be low, that there be a very good gain match between the positive and negative gains, and that the positive and negative slew rates of the difference amplifier be closely matched. These requirements are met by an INA134.

To preserve good carrier suppression one should also match the duty cycle of the TTL switch drive (usually called the "reference signal") to that of the input signal which we are attempting to recover. Most input signals will use a 50% duty cycle (square wave modulation), so deriving the sync by dividing a free-running oscillator with a D-Q flip-flop will guarantee that both the signal and detector receive an accurate 50% duty cycle reference signal.

Synchronous detector operation can be understood by thinking of it as an RF mixer. The RF input is the "input" and the local oscillator (LO) input is the reference. At the mixer intermediate frequency (IF) output we find a low pass filter (LPF) rather than the usual band pass filter (BPF) that one normally encounters in a radio receiver.

In its operation, the synchronous detector operates exactly like the RF mixer. While the mixer generates sum and difference image frequencies (sidebands) that are centeredaround its carrier (the LO frequency), these are spaced at some spectral distance–which is determined by the IF center frequency– from the LO.

Due to the use of a LPF for the IF stage following a synchronous detector, its sidebands extend symmetrically from the carrier to a frequency that is determined by the LPF corner frequency. The "receiver" thus formed by a synchronous detector, its reference, and LPF can have an incredibly narrow bandwidth.

Fortunately, the input signal frequency is determined by the same reference signal that drives the detector. This keeps the input signal centered in the narrow passband of the synchronous detector. Hence the name "synchronous" detector.

By making the LPF filter corner frequency lower and lower, less and less noise power is passed through the detector system and the signal-to-noise ratio (SNR) is thereby improved. Bandwidths as narrow as 0.01Hz are practical and allow a synchronous detector to easily recover a signal that, on an oscilloscope, appears to be totally buried in white noise. It is remarkable to be able to detect a signal with a SNR of only -30 to -40dB!

Applications Circuits: Audio

Low Noise Applications

INA103 microphone amplifier circuit here.

Low Distortion Applications

INA103, INA134, & INA137 line amplifier circuits here. Include DRV134 in this section.

Difference Amplifier Input Resistance

A thoughtful inspection of a difference amplifier circuit reveals one peculiarity of its inputs—the resistances to ground are different! For signal sources with very low output impedances this is not a problem, but in some instances this "imbalance" is a source of concern.

We can easily analyze a difference amplifier with an analog circuit simulation computer program such as P- Spice. The result clearly reveals the interdependence of the two inputs.

Normally, one defines *input resistance* (Rin) as a change of input voltage divided by the corresponding change in input current. In this case, however, we are referring to an input voltage divided by the corresponding change in input current caused by a voltage applied to the *other* input. There apparently is no industry- recognized word or descriptive phrase that describes this, therefore the author has somewhat arbitrarily adopted the term "source load resistance" (SLR) for this discussion. Apologies may be due to those readers who are photographers and will think of this as referring to a "single lens reflex" camera.

Consider a unity- gain difference amplifier created in P- Spice with 1VDC on the inverting input (INV) and an AC sine wave (4kHz, 1V peak) on the non-inverting input (NI). This circuit is shown in Figure.xx below.
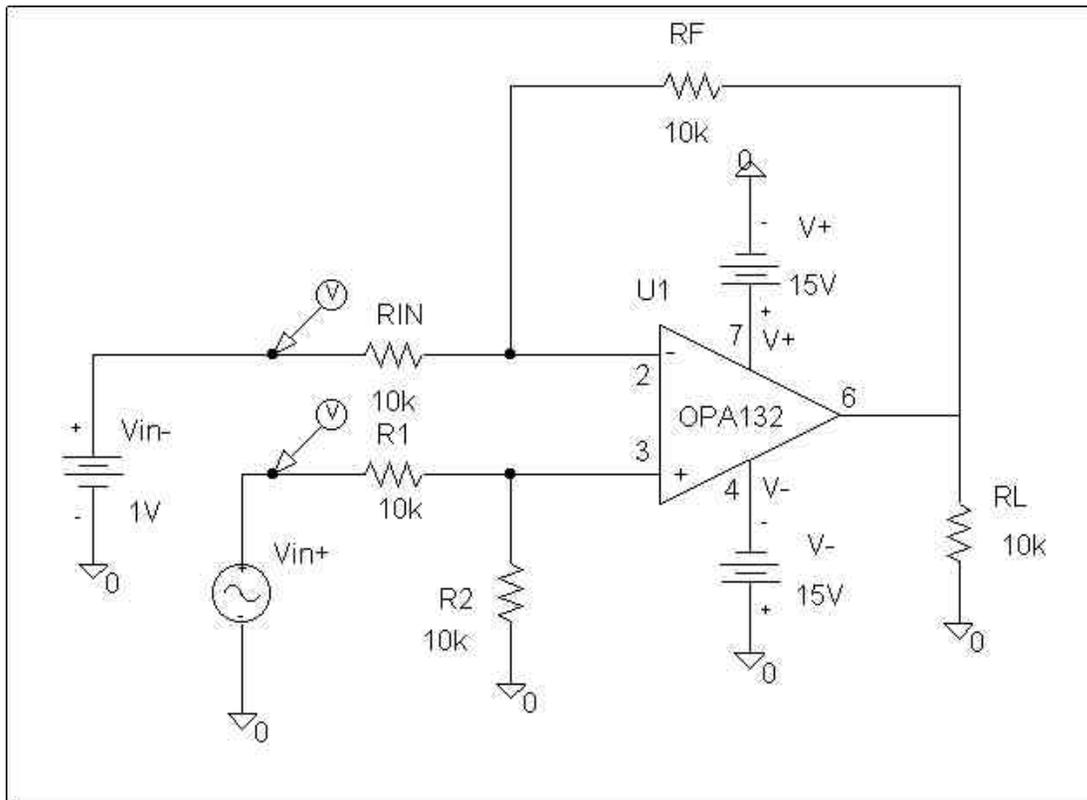
Figure. 6- 9 Schematic Capture Drawing: Simulated Unity- Gain Difference Amplifier With Two Independent Signal Sources.

As expected, due to the op amp's negative feedback or servo action, the AC voltage appearing on the op amp NI (+) input causes the INV (-) input to be equal to it. Since one end of the input resistor RIN is tied to a fixed 1VDC signal source, this AC voltage on the INV input creates a varying voltage drop across that resistor. As the voltage across RIN varies, the current flowing through the input resistor RIN also varies. This input exhibits a varying load on its signal source. Surprise!

Remembering that the current that must be supplied by the signal source is $I_{in} = \Delta V \big/ R_{in}$

and that $\Delta V = (V_{in-}) - \left(V_{in+} \big/ 2 \right)$, we can see that since one voltage is fixed at 1VDC and the other is varying (AC), our input current will not be constant.

When this load "resistance" seen by the source connected to the diff amp's inverting input is calculated using "Ohm's Law", the INV input "resistance" shows a large variation due to the voltage applied to the other input. With the values used in this example, the inverting input of the diff amp presents a load that varies between 20k and about 6k ohms to its source. Input SLR is plotted for both inputs in Figure xx.

In fact, if the voltage on the NI input is twice the voltage on the INV input (assuming equal value resistors R1 and R2 used in this example), there will be no voltage drop across the input resistor. With no current flowing through this resistor, the INV input SLR becomes infinite. An even larger voltage on the NI input will cause current to flow backwards and the calculated SLR becomes negative!
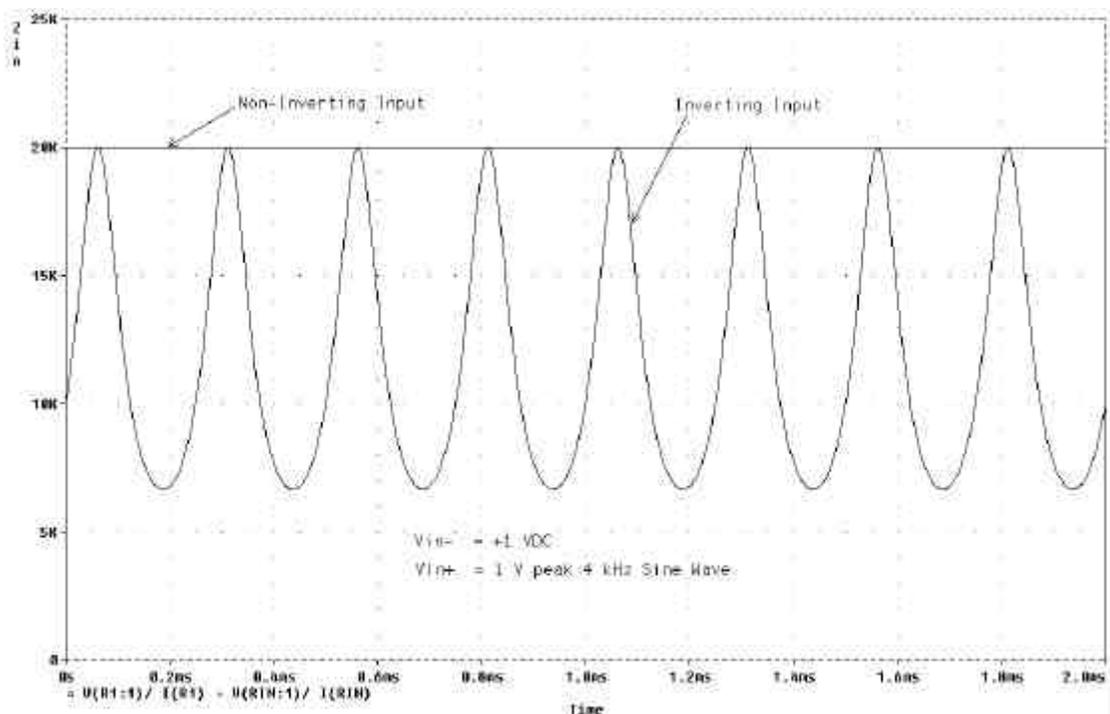
Figure 6-10. Unexpected Behavior? Difference Amplifier Inputs Exhibit "Different" Loading OfTheir Respective Signal Sources.

These considerations are only valid for two independent single- ended signal sources driving the inputs of a difference amplifier. The differential input resistance is another story.

Conclusion: unless the INV signal source has a very low output impedance, its output will be modulated by the NI signal source. The NI input, by comparison, is well- behaved and stays constant– at 20k ohms in this example.
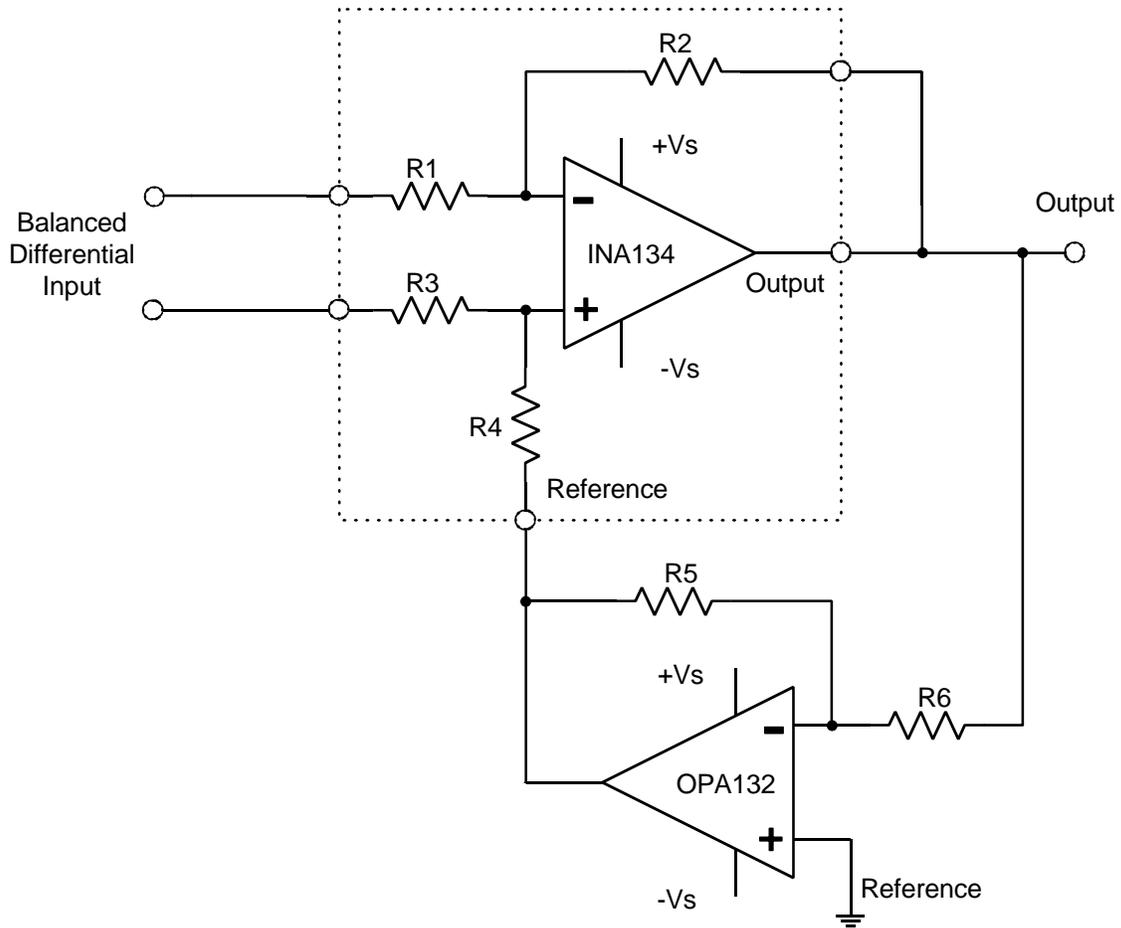
Figure 6- 11. Is The Input Loading of This Circuit Better Than Figure 6- 10?

Applications Circuits: Current Measurement

Current Shunts

High current measurements are made by measuring the voltage drop across a shunt resistor. In order to keep the measurement voltage drop (burden) low, the shunt resistance is kept as low as possible. Typical values lie between a few milliohms and about 10 ohms.

Two conventions are used in current measurement– high side and low side shunts. This refers to whether the shunt is placed in series with the supply output and the load or placed in the supply or load ground return. The current is the same either way, of course , but there is one important consideration in choosing which convention to use. Low side shunts are ground-referenced while high side shunts operate with the full power supply voltage as a common mode voltage on the IA or diff amp.
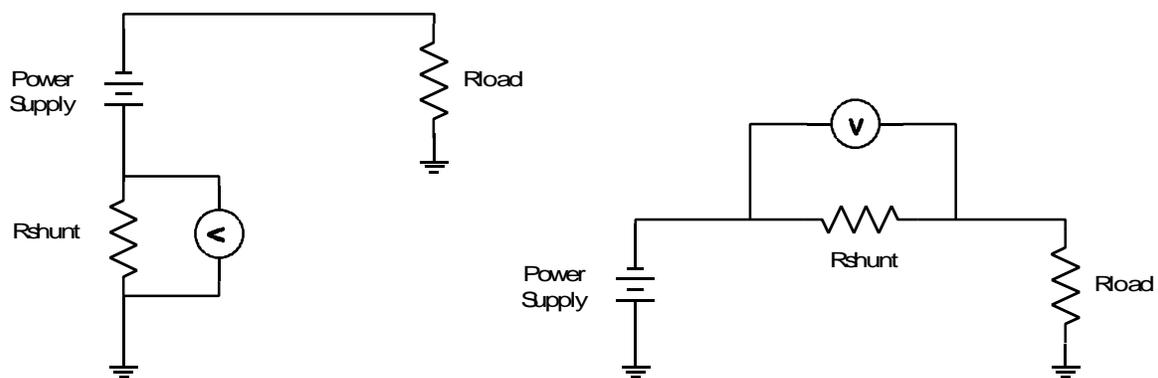
Figure 7- 1   (a.) High Side Current Shunt   (b.) Low Side Current Shunt

Often, there may be shunts on multiple supply outputs so that current can be monitored into each circuit branch. This necessitates a high side shunt convention. A low side shunt placed in the power supply ground return in this example would measure the total supply output current.

In the days of the D'Arsonval (analog panel) meter, an industry standard was developed for shunt resistors which defined their full- scale range to be 50mV. The "ammeter" was actually a sensitive voltmeter; their movements were all 50mV full- scale and full- scale range was determined an appropriate shunt resistor. Meter scales were labeled to correspond to the shunt.

As we shall see, existing 50mV shunts can easily be interfaced to instrumentation amplifiers and difference amplifiers to upgrade older systems. Low accuracy systems can employ an op amp if the shunt is in the low side. While the ground referenced shunt allows this single ended op amp connection, it also creates a serious limitation to its measurement accuracy due to ground loop errors. Small error voltages are created by the PC board traces or wiring in the shunt ground connection that appear in series with the op amp non-inverting input. This error becomes especially serious in high current measurements— here, the shunt resistance may be milliohms and keeping the ground return resistance negligible with respect to such a low resistance becomes impossible.

In addition to wiring and PC board trace resistance errors, there is one more factor to consider. High quality shunt resistors employ Kelvin sensing (a 4- terminal connection) for high current measurements in order to eliminate lead resistance errors within the body of the shunt resistor alone. These types of shunts will require a diff amp or IA, as the Kelvin sensing terminals will probably be at least a few mV above ground— even in a low- side sensing application.

Low-Side Current Sensing

A typical 50mV low-side shunt connection is illustrated in Fig. 22. For high accuracy this common mode voltage— although only a few mV— must be rejected by the amplifier. A common mode error of only 1mV in a 50mV shunt will generate a full- scale error of 2% if a single-ended (op amp) connection is attempted.
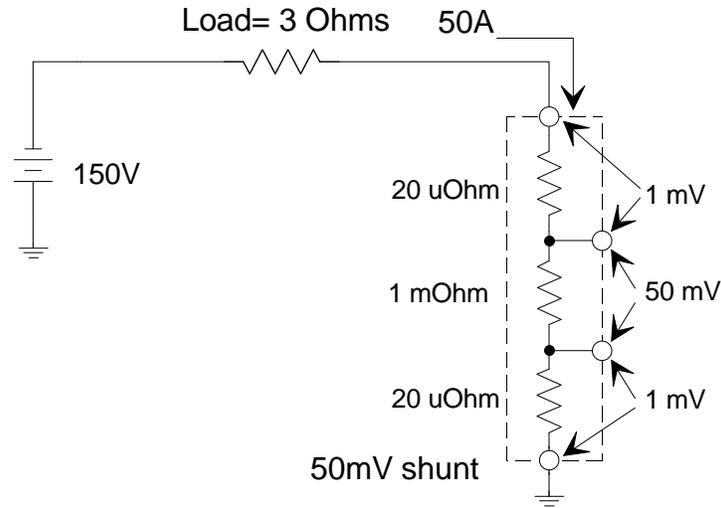
Figure 7-2. Shunt Resistor Kelvin Connection

The solution to ground loop (ground return) errors is to sense the voltage across the shunt with a differential amplifier so that the ground loop error voltage can be rejected. By using a difference amplifier (such as an INA132) as shown in Figure 23 ., the lead connection CMV error is rejected by the CMR of the amplifier.

The measurement error due to the finite input impedance of a difference amplifier will, in most cases, be negligible as the shunt resistor will be far lower than the diff amp input resistance.
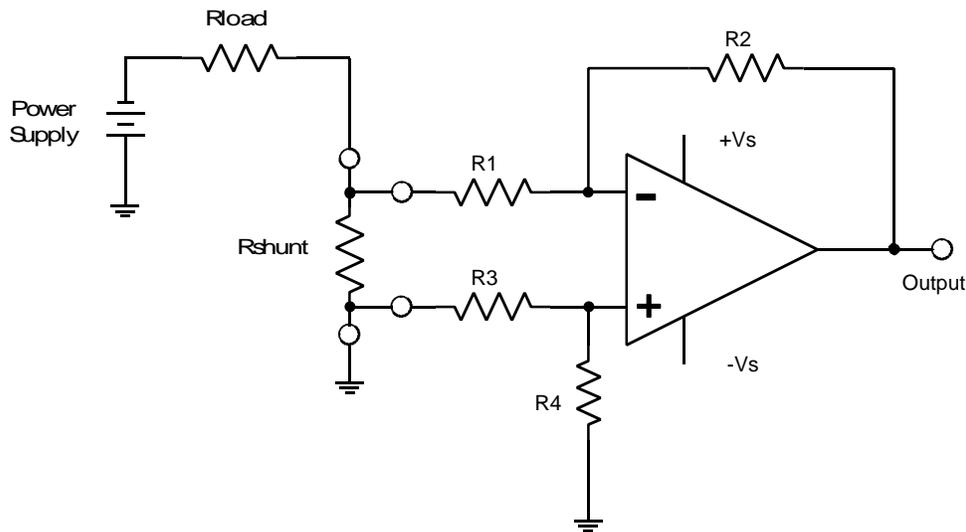
Figure 7-3. Differential Sensing Of The Voltage Drop Across A Low-Side Shunt Resistor Minimizes Ground Loop Errors

In other applications it may be more convenient to have the load tied directly to ground. In this case, the shunt resistor may be inserted into the power supply ground return connection as shown in Figure 24.
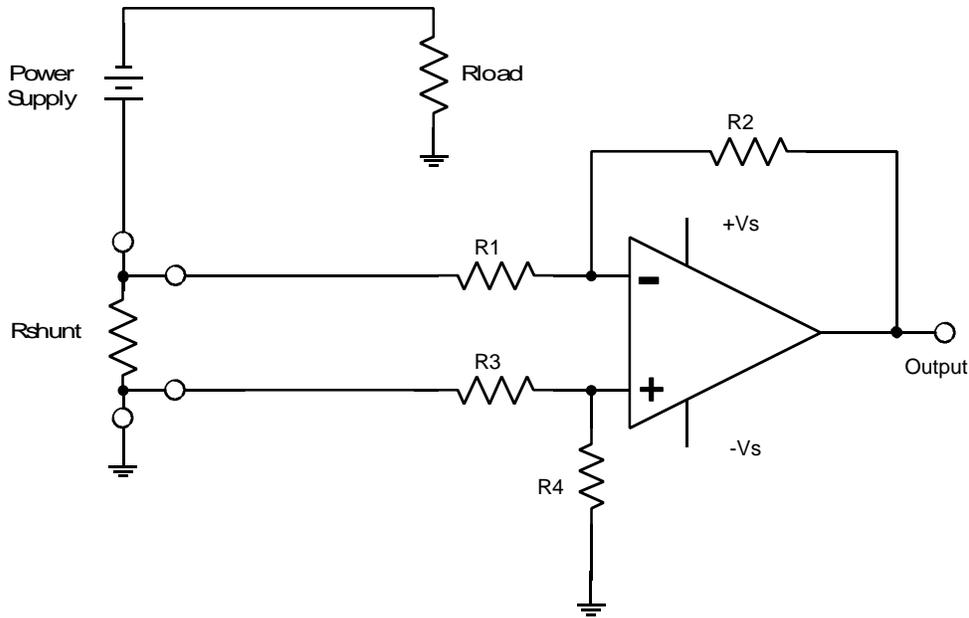
Figure 7-4. Alternative Connection Of Shunt Resistor. Re: Figure 24.

As we have seen, a 50mV shunt produces a very small differential signal that must be amplified to be useful. Small errors such as finite CMR and PSR, input offset voltage, and input offset voltage drift can contribute serious errors. The difference amplifier or instrumentation amplifier must be carefully chosen so that an overall system error budget can be met.

Designing a low-side current sensing circuit that operates on a single supply imposes a requirement that the amplifier be able to handle common mode voltages that go all the way to its negative supply– ground. A rail-to-rail input *and* output two op amp instrumentation amplifier serves this function– in a gain of 10V/V-- in the application circuit of Figure 26. A maximum offset voltage spec of +/-250uV keeps the amplifier errors small in comparison to the 500mV shunt voltage.
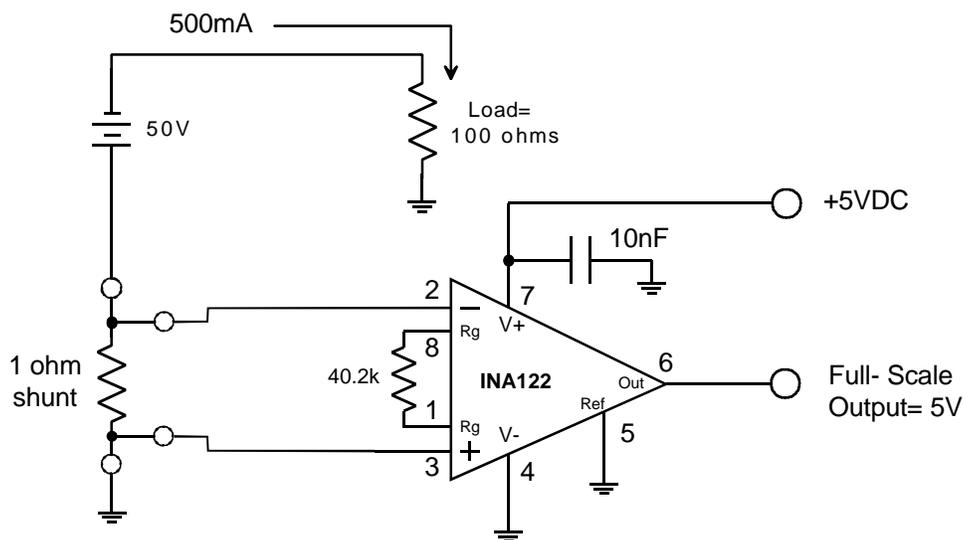
Figure 7.5. Low-Side Shunt Amplifier With Single Supply.

High-Side Current Sensing

High-side sensing is sometimes necessary due to system constraints but it imposes more severe requirements on the differential amplifier's common mode rejection performance. Instead of a few millivolts in the case of low-side current shunt, the amplifier is now subjected to the full output voltage of the power supply. Common mode voltage may be as much as four or five orders of magnitude higher for a high-side current shunt amplifier.

One of the first steps in selecting a difference amplifier or instrumentation amplifier for high-side current sensing applications is to see that the amplifier's common mode voltage range will include the high-side voltage (plus a margin for safety) under all conditions of that amplifier's output swing.

Instrumentation amplifiers offer higher accuracy than simple diff amps and in low current measurements, their extremely high input impedance does not create appreciable error ("shunting the shunt" as it were) even with shunt resistors as high as 1 megohm.

In cost-driven systems, a simple difference amplifier may be a good low cost approach to high-side current measurement. Diff amps also offer higher CMV specifications (up to 200V) than IAs can achieve.

Kilovolt-range common mode voltages are the province of specialized amplifiers with no galvanic connection between their inputs and their output. These are termed "Isolation Amplifiers."

To measure current on up to a 200V common mode voltage requires a special type of difference amplifier that can withstand unusually high CMV without damage– the INA117. A precision thin film resistor network on-chip divides down the input while simultaneously providing gain in the op amp to bring the signal back up to its original level. In general purpose applications this unity-gain amplifier offers a simple low cost alternative to using an isolation amplifier. Accuracy suffers in very low level applications, however, and measuring current with an industry-standard 50mV shunt resistor becomes impractical due to the amplifier's input errors.

A plastic-package surface mount INA117 has an input offset specified as 2mV maximum. This is an initial full-scale error of 4% due to offset alone. Although this offset can be trimmed to zero, the offset drift (40uV max) is a limitation on accuracy. DC common mode rejection is 80dB typ but a quick calculation reveals that a 150V CMV will produce a 30% error (15mV) on a 50mV input signal. Adding both error terms, we see that the worst case error could be as much as 34%. Needless to say, this is not acceptable.

All is not lost, however. If the 50mV signal can be gained up by a preamplifier, the INA117's errors will be reduced by the gain of the preamp. A high accuracy preamp is necessary, of course, so as not to introduce its own errors. There are excellent op amps available on the

market that will meet the accuracy criteria but they are single- ended (ground referenced) and this application requires a "differential amplifier."

By employing an isolated (1kV rms) 1 watt DC-DC converter to power the preamplifier, the op amp is "tricked" into acting as a differential amplifier by floating its output and common. These are then fed to the inputs of the INA117, where the common mode voltage is rejected.

An applications circuit using this approach is shown in Figure 27. The OPA277 op amp preamplifier provides a gain of 200V/V. Thus, the difference amplifier sees a 10V input rather than only the 50mV input that it would see without the preamp. It is important to select a precision op amp as its input errors can be a limitation on the accuracy of the measurement. In the case of the OPA277, its input offset voltage and drift are very low-- +/- 20uV and +/- 0.15uV maximum.

Initial error due to the preamp input becomes 20uV/50mV or only 0.04%. Strictly speaking, the preamp offset voltage will be multiplied by the noise gain of the preamp op amp, not by the circuit gain. Since the difference between a noise gain of 200+1V/V and 200V/V is very small, we can simply use the circuit gain when this gain is high. The diff amp input offset voltage error then is reduced to 2mV/10V or 0.02%.

Errors due to the diff amp's finite common mode rejection must also be considered. In our example, the INA117 inputs see a 150V common mode voltage. Typical INA117 CMR is 80dB or 10^{-4}. Multiplying 150V x 10^{-4} gives a CMR error of 15mV and calculating the percentage of error on a 10V input: 15mV/10V= 0.15%. Worst case error of the two amplifiers is a low 0.21% but realistically a circuit's errors will not be all worst case at the same time. Statistically, one may expect to achieve better than 0.21% error.

Adding a preamp to an INA117 has made quite an improvement in accuracy!
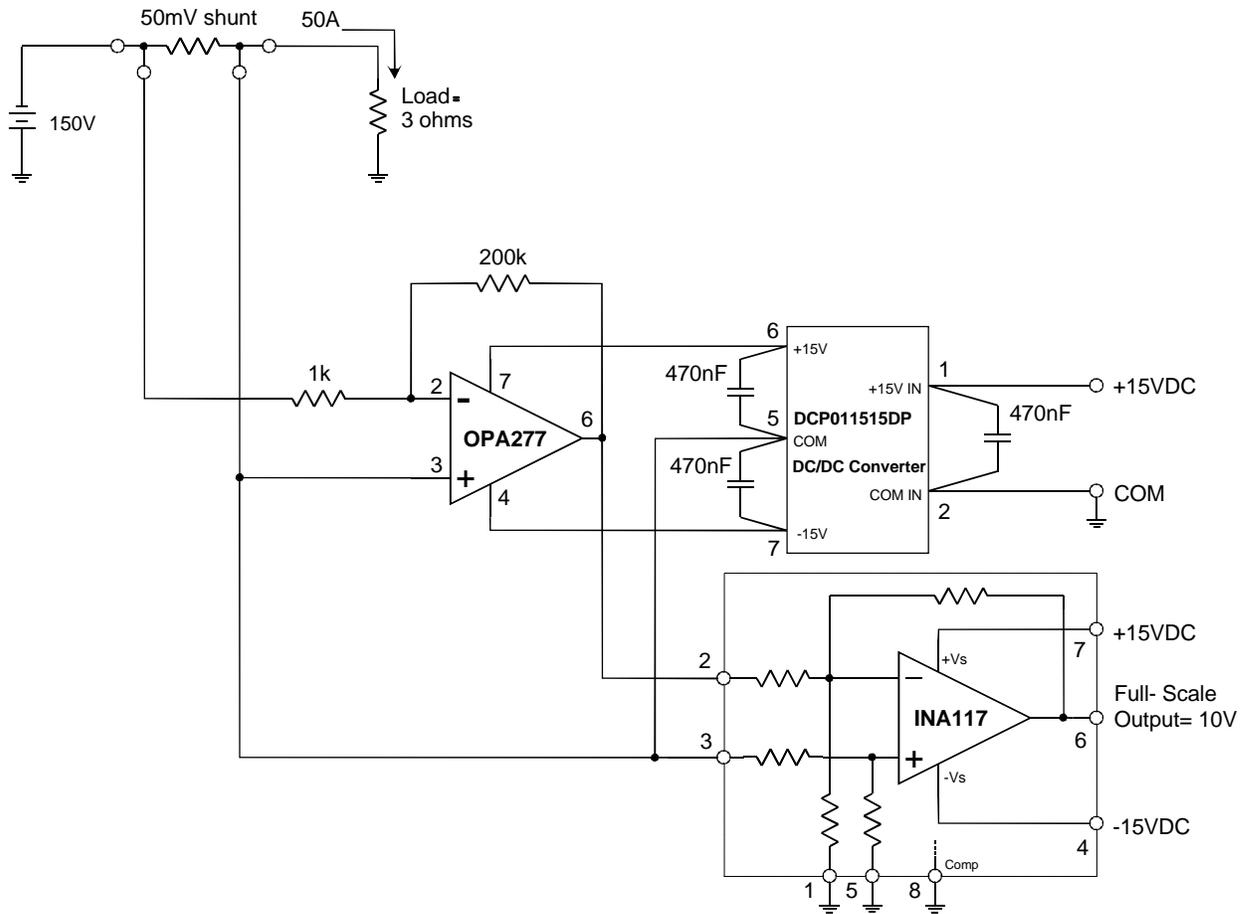
Figure 7- 6. High Accuracy Current Measurement With Up To 200V Common Mode Voltage.

In similar example, designers can use a differential amplifier in a circuit for very small leakage current measurements under high common mode voltage conditions (Figure. 28). When the DUT (device under test) ground return path is not available for low- side sensing, the cir cuit must sense leakage current with a high- side current shunt resistor in series with the DUT's input under test. Although similar in many ways to our previous example, this application requires a very low bias current op amp to minimize error with its 100megohm source resistance.
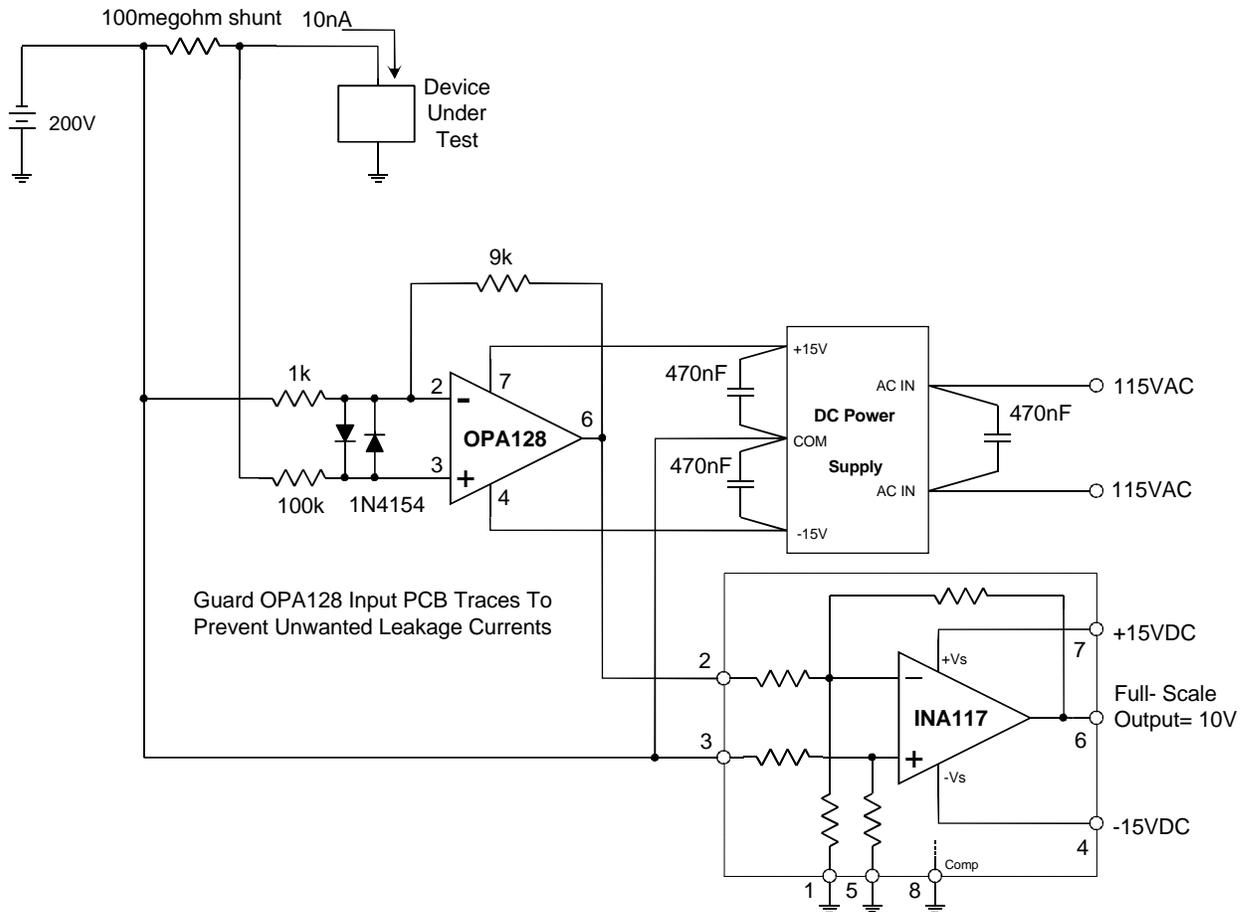
Figure 7-7. Accurate Low Current Measurements With Up To +/-200V Common Mode Voltage.

Because the leakage currents that we are measuring are small, an OPA128 electrometer-grade op amp is chosen to sense a 1V drop caused by a 10nA leakage current flowing through a high (100megohm) shunt resistor. The electrometer op amp's bias current is low-- less than 75fA–compared to the 10nA full-scale measurement in order to preserve accuracy.

Its 1k input and 9k feedback resistors set the preamplifier's non-inverting voltage gain to 10V/V. Two 1N4154 clamp diodes and l00k resistors protect the op amp from DUT short-circuit fault conditions of the +/-200V power source. Since an op amp cannot withstand a 200-V differential input without damage– the result of a DUT short to ground– the input clamps are added for protection.

The preamplifier is powered by an isolated +/-15VDC power supply and thus it "floats" at the +/-200V common mode voltage. This CMV is then rejected by the diff amp and, as a result, it delivers a +/-10V output voltage referenced to the ground that is proportional to the + /-10nA leakage current in the DUT.

Although a small DC/DC converter can be used to power this circuit (as shown previously), the 100 megohm source resistance of the shunt makes this circuit much more sensitive to noise pickup. Linear supplies are much "cleaner" and a small modular +/-15VDC power supply with

at least a 500V isolation voltage rating (200V CMV plus a safety margin) between the output and common and between the output and input is recommended for this preamplifier.

Common-mode rejection ratio (CMR) is the measure of the IA's ability to reject common-mode signals. CMR (expressed in dB) is mathematically equal to Gain (dB) + 20*LOG (dV_d/dV_{cm}). In addition, since CMR is the sum of common-mode gain to differential gain (in dB), this circuit boosts the overall CMR due to the OPA128 preamplifier's gain. The 20-dB gain of the 0PA128 op amp combined with the 86-dB CMR of the INA 117 difference amplifier results in a total minimum CMR of 106dB.

Micropower & Battery - Powered Applications

Single Supply Considerations

Single supply operation imposes important circuit design constraints which may force a designer to take a quite different approach than he would for a conventional split power supply circuit.

Some of these may seem obvious– don't try to swing an output negative, for example. In our previous designs– which always used split power supplies– most of us have become so accustomed to not worrying about negative- going output swings that this error is frequently overlooked in the circuit design stage. This fundamental error will be eventually found when the breadboarded circuit is bench tested-- the red- faced engineer then realizes that he has expected an amplifier output to swing below ground! Hopefully, no one else has noticed his error and he can quietly correct it without further embarrassment.

Remember Albaugh's Law– "two inverting amplifiers in cascade won't work." No matter what polarity of signal is applied to the input, one of the two inverting amplifiers will try to swing negative– ie., below ground. In single supply circuit designs, approach inverting amplifiers with caution.

Biasing the inputs and outputs above ground is the only solution to this negative- going output swing dilemma. Frequently, a reference level of one- half of the positive supply (V+) is chosen for the bias level. An advantage of this approach is that it offers the maximum possible output voltage swing and the input is also relatively well centered in an amplifier's common mode input voltage range. A +2.5V precision voltage reference such as a REF1004-2.5 is recommended for single +5V supply circuits.

If a ratiometric approach can be tolerated, a simple resistive voltage divider can be substituted for a fixed reference voltage. Two equal- value resistors will provide an offset voltage of ½V+ into a high impedance. This is certainly an economical method of generating an offset voltage for referencing a single supply amplifier, but it lacks the absolute accuracy of the fixed voltage reference.

In cases where the circuit supply voltage can vary over a wide range– such as in battery- operated instruments-- the fixed reference voltage can become a liability, however. Consider what happens to a single +5VDC amplifier circuit with a fixed +2.5VDC reference when the battery voltage drops to an end- life voltage of +2.7V: with a +5V supply, everything works quite well as the +2.5V offset voltage is centered within the amplifier's common mode voltage range.

Under the battery end-life operating conditions, the amplifier's supply voltage drops to +2.7V but its offset remains fixed at +2.5V and, unless the amplifier has an adequate input CMV range, the circuit stops working. Even if the amplifier's CMV range is wide enough to accommodate the reference voltage, its output swing will be severely restricted– it is only 200mV from the positive rail.

Fortunately, the resistive divider will always give us a reference voltage that "tracks" ½V+, so under the same conditions as before, the amplifier will see a reference of +1.35V. This is perfectly centered within the amplifier's common mode voltage range and its output is also centered between the +2.7V supply and ground. This allows the maximum possible output swing.

The instrumentation amplifier shown in Figure xx. illustrates the offset method used to allow a single supply circuit to swing its output both negative as well as positive. In fact, this amplifier exhibits outstanding R-R output swing– it can swing to within 10mV (Figure xx.) of either supply rail with a 20k ohm load. This remarkable performance is due to the excellent R-R CMOS op amps used in this IA design. Their 35MHz GBW also contributes to this instrumentation amplifier's good bandwidth and high frequency CMR performance.
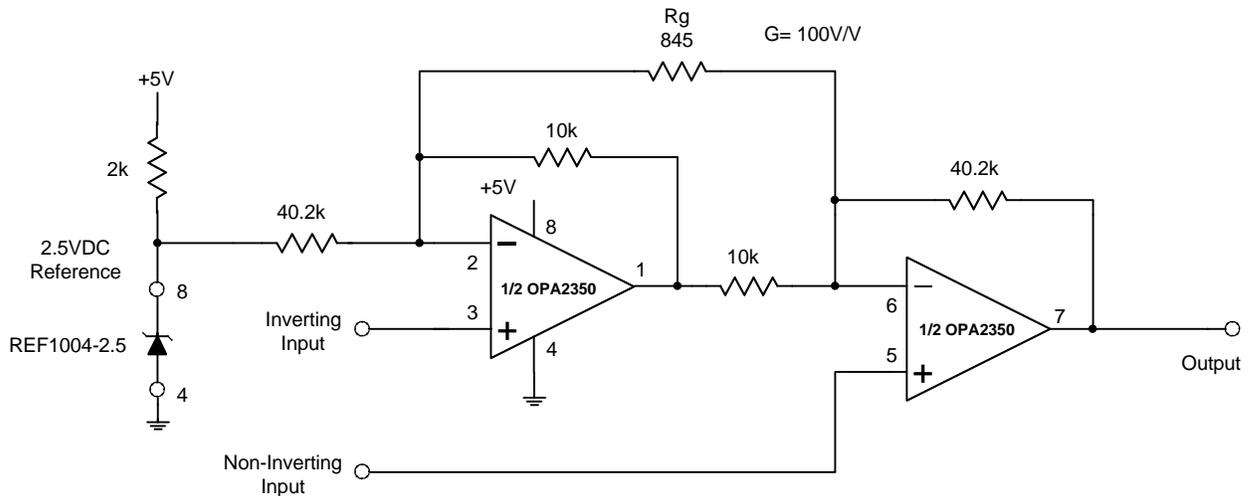
Figure 8- 1. A Fast R-R Single+5V Supply Instrumentation Amplifier With A Gain Of 100V/V.
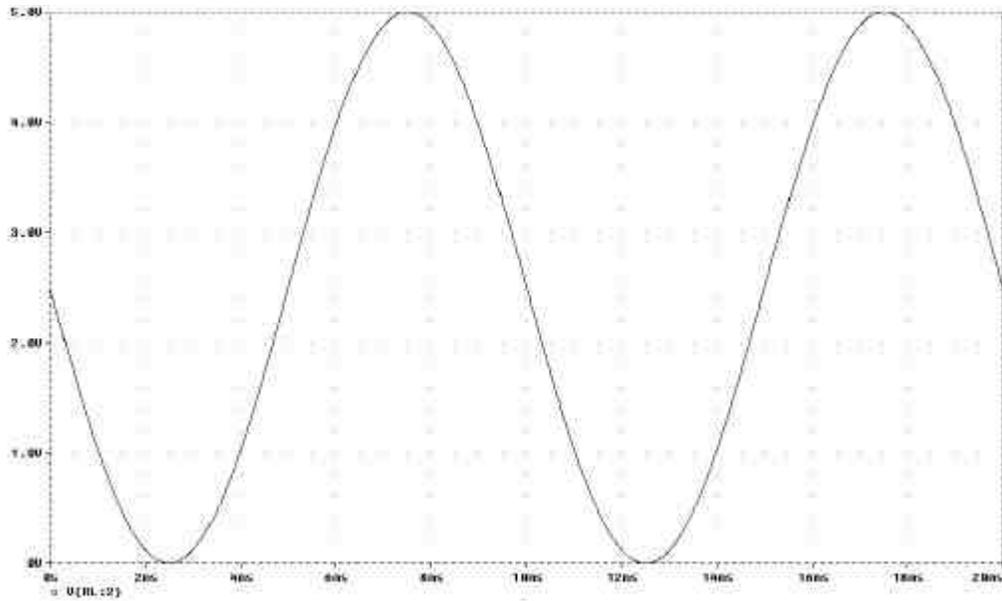
Figure 8- 2. CMOS Instrumentation Amplifier Swings To Within 10mV Of The Supply Rails.

A reference voltage source impedance must be as low as possible to maintain an IA's high CMR. In the circuit shown in Figure xx., the REF1004-2.5 reference has a DC output impedance of only about 0.2 ohms. For higher source impedances– particularly when employing a resistive voltage divider reference– an op amp buffer will be necessary to prevent CMR degradation.

It is particularly important to verify that any amplifier under consideration for a single supply application have an adequate common mode input voltage range as well as an adequate output voltage swing. Even if the amplifier's output voltage swing is within the manufacturer's specification, it is important to recognize that swinging close to the rail does not come without a penalty.

As an op amp output voltage approaches the supply rail, its open- loop gain is seriously reduced. This can compromise accuracy and, in some cases, result in a low- level oscillation that is caused by reduced phase margin. What can be done about this? Actually, the end user has little recourse other than to minimize the load resistance and to select an amplifier with an output stage topology that is less susceptible to this effect. The best approach, of course, is to not push the output close to the rail. This advice may sound similar to the old vaudeville routine (Patient: "Doctor, it hurts when I do this.. ." , Doctor: "Don't do that!"), but it is the best advice you will get.

The trend in industry is to power both analog and digital circuits from a single unipolar supply. Due to most digital logic family requirements of a +5VDC supply, this is also what the analog circuit designer is forced to use as well. As TTL logic families have improved their speed- power product and as battery operated electronics have become more widely used, supply voltages of +3.3VDC are now frequently encountered. Some circuits are even powered by 2.7VDC supplies or lower.

Linear IC manufacturers have followed suit with op amps and IAs that can operate on these very low supply voltages but there is one compromise that simply cannot be overcome— reduced dynamic range. Signal amplitude (voltage swing) within a circuit is limited by its supply voltage but the amplifier's input noise can only be pushed to an irreducible minimum limited by the Laws of Physics. As the amplifiers' supply voltage is reduced, the ratio of their output signal swing to their input noise is also reduced. Thus SNR is lost and cannot be recovered.

Audio engineers recognized this long ago; some audio op amps were designed to operate on +/-24V to increase their output voltage capability and thereby improve the system's dynamic range.

For the linear designer, the supply voltage trend is in exactly the *wrong* direction. A strong case can be made for employing small charge pumps or dc/dc converters (such as the 1 watt DCP01 family) to boost the linear circuit's supply voltage back up to a more reasonable level or to generate a bipolar (split) supply.

Minimizing Supply Current

Portable battery-operated instruments place severe restrictions on a circuit designer's choice of devices and circuit topologies. His primary concern is that his circuit must work over a wide range of supply voltage while drawing as little current as possible in order to maximize battery life. This battery is usually only a single unipolar supply. The battery voltage that a designer must cope with may vary from a high— experienced while the battery is being charged— to a low that is experienced at the end of the battery's useful life (end-life voltage).

Battery voltage depends on the type of battery (actually on its internal chemistry) that has been chosen to power the instrument and the number of cells that are connected in series within the battery. Common cell types in primary batteries (non-rechargeable) and secondary batteries (rechargeable) that are used in portable instruments are listed in Table xx.

This table lists the nominal operating voltage for cells of various types. In operation, the open-circuit or very light load cell voltage will be higher than its nominal voltage and, as the battery is discharged, its terminal voltage will decline to a much lower voltage. The battery's cut-off voltage (also known as its end-life voltage) may be determined by a monitor circuit that drops the load when the battery is partially discharged in order to prevent reverse current from flowing in the weakest (a completely discharged) cell.

Secondary batteries are charged at a much higher voltage than their open-circuit voltage. The charging voltage is determined by the charging method used. Some 1.2V nominal operating voltage cells are charged to as high as 1.8V to 1.9V at the end of their charge cycle by "smart" battery chargers. Sophisticated monitoring of battery parameters such as current, voltage, temperature, and time is employed to allow the battery to be charged quickly and safely.

Choosing an instrumentation amplifier for a battery-operated instrument application is straightforward. It will require the features already mentioned in the Single Supply Considerations as well as requiring low quiescent current (Iq) and a capability of operating on low supply voltages.

All things being equal, a two op amp IA is preferred over a three op amp IA for two reasons:

1. One less op amp in its topology gives it the edge in Iq.

2. The CMV range is superior- especially in low supply voltage applications.

Don't overlook the fact that load currents must be supplied by each amplifier in the circuit. Selecting a very low Iq amplifier and then putting a 1k ohm load resistor on its output will negate all of your efforts. After all, with a 1k ohm load, a 2V output will require a 2mA output current from the amplifier. This may be *decades* larger than the *amplifier's* quiescent current! Remember, too, that current flows in op amp feedback resistors as well.

A simple rule of thumb:

1. Keep resistor values as high as possible when operating current is at a premium.

Type Of Cell	Nominal Voltage Per Cell (VDC)	Rechargeable?
Laclanche (ordinary dry cell)	1.5	No
Alkaline	1.50	No
Lithium (Li/SOCl$_2$)	3.5	No
Lithium (Li/MnO$_2$)	3.0	No
Mercury (Hg)	1.35	No
Silver Oxide (Ag$_2$O)	1.50	No
Zinc- Air	1.35	No
Nickel- Cadmium (Ni-Cd)	1.20	Yes
Lead- Acid	2.0	Yes
Nickel Metal Hydride (Ni-MH)	1.2	Yes

Table xx. Cell Data For Batteries Commonly Used In Portable Electronic Instruments

One final reminder– do not overlook the necessity of bypassing the supply voltage! This is critically important in battery- operated circuits because of the battery's internal impedance. Power supply impedance in battery- operated instruments is higher than one normally encounters in voltage- regulated line- operated circuits. As a battery is discharged, its internal impedance increases significantly and amplifier stage-to-stage coupling through their common supply impedance can cause instability or oscillation.

The high internal resistance of partially- discharged batteries is also a source of noise. Adding a large electrolytic or tantalum capacitor across the battery will restore its low source impedance above a few Hz and cure the battery's internal impedance problems.

A few devices recommended for low power single supply applications or for battery operated portable equipment are listed in Table xx. These are by no means the only devices that are suitable for these applications– consult the latest Burr- Brown catalog, individual data sheet, CD-ROM, or website (http://www.burr-brown.com) for the latest information on these and other new products.

Device Family Part Number	Device Function	Single Supply Operating Range (VDC)	Quiescent Current Per Device (μA)
INA122	2 Op Amp Inst. Amp	2.2 to 36	60
INA126	2 Op Amp Inst. Amp	2.7 to 36	175
INA132	Difference Amplifier	2.7 to 36	160
OPA234	Op Amp- Bipolar	2.7 to 36	250
OPA237	Op Amp- Bipolar	2.7 to 36	350
OPA241	Op Amp- Bipolar	2.7 to 36	24
OPA277	Op Amp- Bipolar	4 to 36	800
OPA336	Op Amp- CMOS	2.3 to 5.5	20
OPA340	Op Amp- CMOS	2.5 to 5.5	750
OPA2244	Op Amp- Bipolar	2.2 to 36	40
OPA2337	Op Amp- CMOS	2.5 to 5.5	525
OPT101	Photodiode + Transimpedance Amp	2.7 to 36	120

Table xx. A Few Devices That Are Recommended For Single Supply Or Battery Operated Instrument Applications.

Applications Circuits: Unusual

Extending Common Mode Range To 1kV

For some applications the common mode voltage requirements may exceed anything that is currently available from IC vendors. If the circuit can tolerate an input impedance of 10M, we can add a pair of 100:1 voltage dividers to the inputs of an instrumentation amplifier and extend its common mode voltage range to 1,000V. The key to success with this approach is the use of very well matched dividers.

Any error in the divider ratio matching will degrade the CMR of the amplifier, so well-matched components must be employed. The matching must be extremely accurate and they must have a low (and matching) temperature coefficient of resistance (TCR). In addition, since we are dealing with a fairly high voltage across the input resistors, they must not only be able to safely withstand this voltage but they must also exhibit a low voltage coefficient of resistance (VCR).

Constructing this amplifier with garden- variety 1% metal film resistors is not advisable– unless many RN55 resistors are series- connected, they cannot withstand the required voltage (their continuous rating is only 250V) and their temperature matching will probably not be very good in most PCB layouts. With care, it could be made to work but there is a better solution.

By using two commercially available precision high voltage dividers, we can establish a definite error budget from their specifications. One vendor for the two 10Megohm 100:1 dividers is Ohmcraft. Their Type HVD thick film dividers are made with voltage ratings of 4kV to 40kV and feature TCRs of +/-25 to 200ppm/C and ratio tolerances of +/-0.1 to 5%. These networks also exhibit a lower VCR than one usually finds in high value thick film resistors.

A 1kV CMV differential amplifier is shown in Figure 29. An INA118 instrumentation amplifier in a gain of 100V/V makes up for the 100:1 input voltage divider and the resulting differential gain is unity (1V/V).
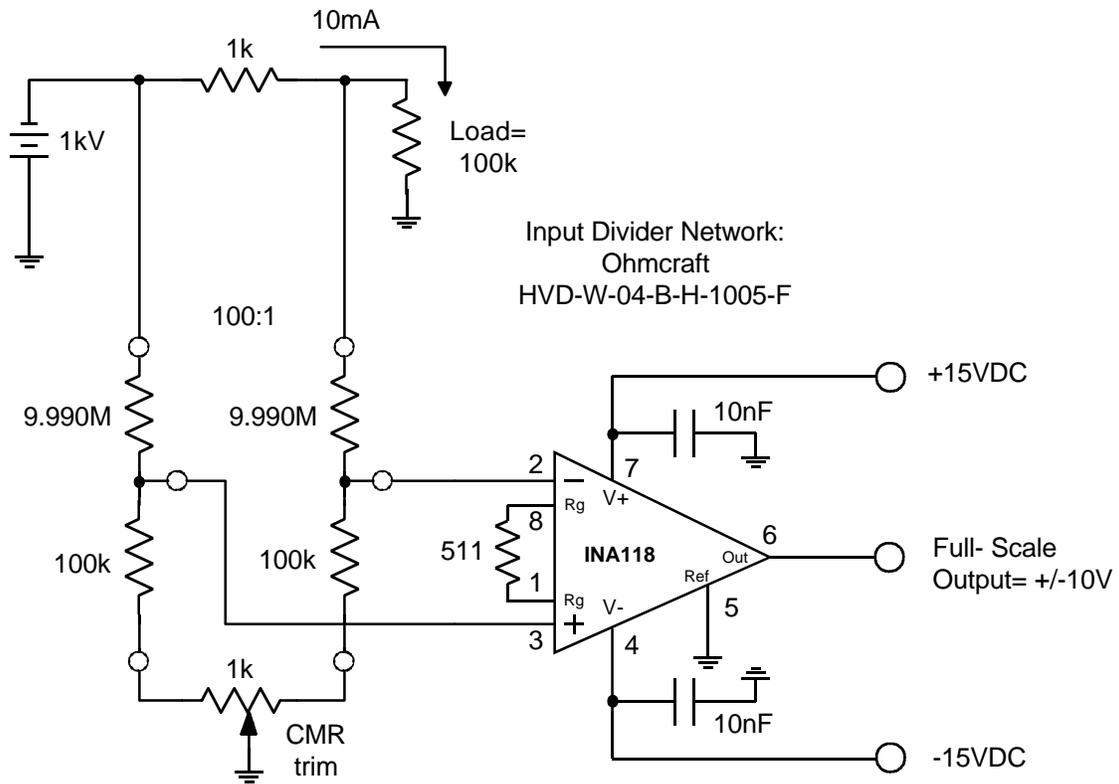
Figure 8-3. A 1kV CMV Differential Amplifier Made With A Precision 100:1 Voltage Divider Added To An IA.
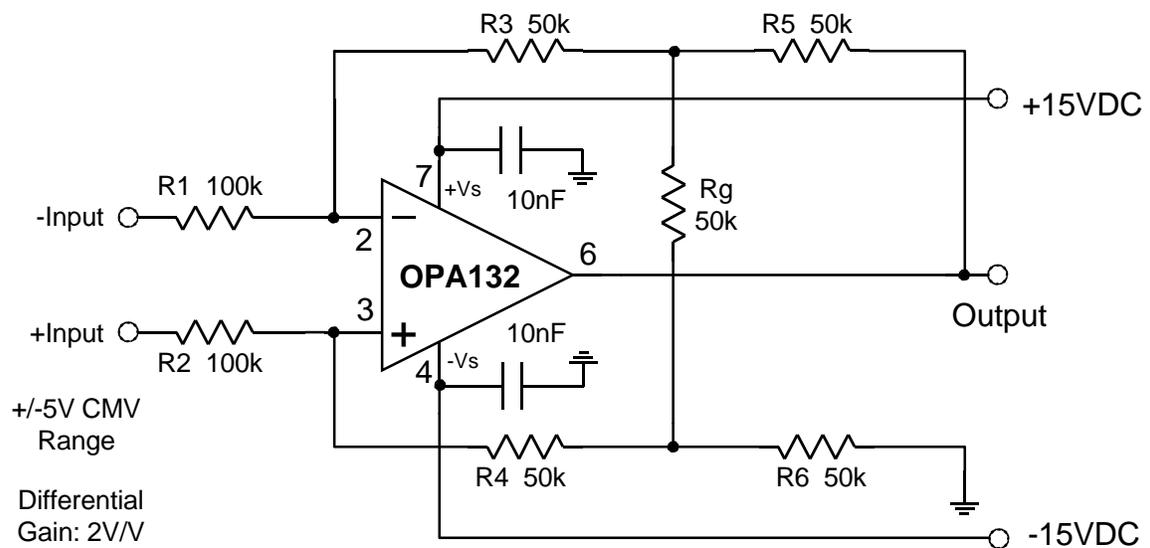
An Adjustable Gain Difference Amplifier

Figure 8-4. A "Different" Resistor Network Plus An Op Amp Yields An Adjustable-Gain Difference Amplifier

Difference amplifiers are commonly encountered in test and measurement circuits and the majority of them are unity gain amplifiers. Most commercially available diff amps are also unity gain amplifiers but some offer different gains[1]. One thing these difference amplifiers have in common is that they all have fixed gain. In many applications it would be useful to "tweak" the amplifier's gain. But how can this be accomplished without juggling two resistors at once?

A recently rediscovered[1] circuit technique for adjusting the gain of a difference amplifier is shown in Figure 30. By adding a couple of extra resistors and a pot (or another fixed resistor), the amplifier's differential gain can be adjusted without compromising its CMR. Its gain equation is:

$$e_o = 2(1 + \frac{1}{K})\frac{R_2}{R_1}[e_2 - e_1] \qquad\qquad\qquad (xx)$$

By inspection, it is found that there are two feedback paths in this circuit– a negative feedback path through R3 and R5, and a positive feedback path through Rg and R4. The two paths are interconnected by Rg which allows a gain adjustment to be made by varying the ratio of negative and positive feedback.

Note that the gain that is achieved by this technique comes at the expense of reduced input CMV range. A gain of 2V/V amplifier is limited to a +/-5V CMV range.

Although not addressed in the original reference, this circuit is not limited to equal value resistors. By using larger resistors in the difference amplifier's input, a high common mode voltage range can be obtained and the overall circuit's gain can be set to unity by employing the differential technique. Nothing is free, however. The drawback is that the op amp's offset voltage, drift, and noise are all multiplied by the gain used to overcome the input resistors' division ratio.

By using an excellent FET op amp such as the OPA132, both its multiplied input offset errors and bias current errors are kept low enough to be practical for most applications. The unity gain difference amplifier shown in Figure 31 achieves a CMV range of +/-50V on +/-15VDC supplies.

1. For example, the INA106 has a differential gain of 10V/V and the INA137 is capable of either a gain of 2V/V or 0.5V/V.

2. This circuit is found in "Operational Amplifiers- Design And Applications", by Graeme, Tobey, & Huelsman, McGraw- Hill Book Company 1971, ISBN 07-064917-0, pp. 203-204.
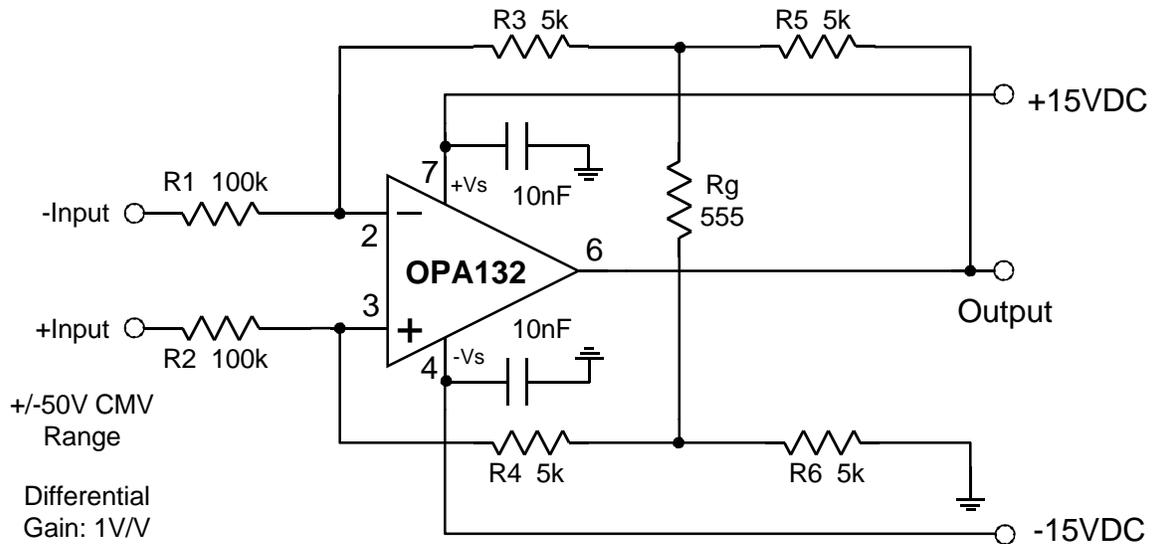
Figure 8-5. Adding Gain To Compensate For An Input Voltage Divider Yields A Unity Gain Difference Amplifier With A 50V Common Mode Voltage Range.

Advantages of Asymmetrical Power Supplies

Discussions of CMV range skew by non-standard PS voltages here.

VLF & LF Loop Antenna Amplifiers

Discussion of the advantages of a low noise INA103 loop antenna preamp here.

Applications Circuits: Optoelectronics

Differential Photodetectors ("Edge Detectors")

Discussion of using a diff amp to generate an edge detector output here.

X/Y Position (Quadrant) Detectors

Discussion of using six diff amps to generate a quadrant detector output here.

CW Laser Current Driver

Discussion of how to generate a low noise, stable & accurate constant current output here.

Selecting Your Instrumentation Amplifier

Sensor Source Impedance Considerations

The output impedance of a signal source will determine which type of instrumentation amplifier will be best suited to that particular application. Bias current effects will be the primary concern, but there are also considerations such as noise (voltage noise, current noise, or both) as well input impedance. Designers have a wide choice of IAs to choose from-- bipolar, FET, CMOS, and chopper types all have their place in appropriate circuits.

High source impedances preclude the use of a difference amplifier due to its low input impedance. Loading of the signal source by the low impedance diff amp input generates serious gain errors, so they must be dropped from consideration when we are facing a high impedance sensor.

Two and three op amp instrumentation amplifiers have input impedances on the order of 1000 megohms or more, so these types are well suited to virtually any source impedance. In addition, the IA offers higher accuracy and gain than a difference amplifier.

Selection of an IA is then made on the basis of bias current and noise as well as the usual requirements of input offset, drift, gain accuracy, bandwidth, CMV range, etc.

Add Table of recommended IAs for low, medium, and high impedancesources here.

Very High Impedance Sensor

An example of a very high impedance sensor is a pH measurement electrode. Depending of the type of probe, a source impedance may be as high as a few gigohms. Clearly, this application requires a very special type of amplifier-- one with very low bias current and high input impedance.

An excellent amplifier for this application would be the INA116 dielectrically isolated FET-input instrumentation amplifier. Featuring a typical bias current of only 3fA ($3x10^{-15}$A typical, $2.5x10^{-14}$A max) and a 10^{15} ohm input impedance, this three op amp instrumentation amplifier could actually be called a "Differential Electrometer."

The pH probe amplifier circuit diagram is shown in Figure 9-1 below.
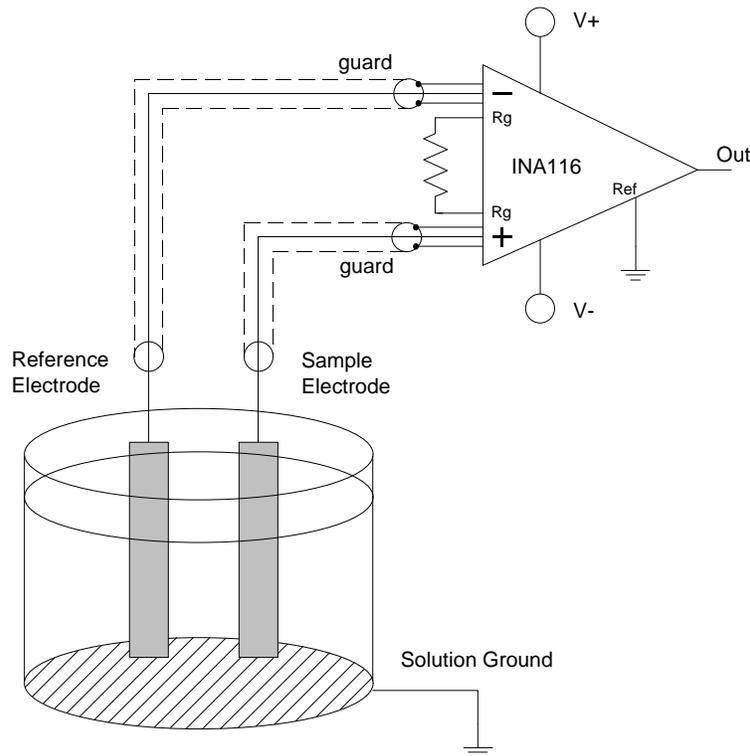
Figure 9-1. Amplifier For pH Measurement

Using the guard driver outputs of the INA116 allows both electrode connections leads to be shielded from external noise pickup without incurring a leakage current or input capacitance penalty. In extremely high impedance circuits such as this, Faraday shielding is absolutely essential.

Without shielding on the electrometer inputs, electrostatic fields such as 60 Hz power line noise will render the circuit useless and even the motion of the human body nearby will displace enough charge into the inputs to interfere with accurate measurements.

Sharp-eyed readers will notice the absence of bias current return resistors in this circuit. They are actually there, but they are "hidden." In this application, the pH solution itself provides a path for IA input bias current conduction to ground. Contact to the conductive container is labeled "Solution Ground." In cases where a nonconductive solution container is used, a third electrode may be introduced into the solution to provide a bias current return path.

One might be tempted to use a resistor from each IA input to ground as a substitute for the "Solution Ground" bias current return, but to maintain the amplifier's high input impedance, resistors with an extremely high value are required. Finding teraohm resistors is not an easy task; besides, they are expensive so when possible, they should be omitted.

Very Low Impedance Sensors

Thermocouples are very low impedance temperature sensors with output voltages in the millivolt range. An instrumentation amplifier for thermocouple use must offer high gain with low offset voltage, low drift, low noise, and—since the thermocouple is frequently in an electrically noisy environment—very high CRR. Bipolar input IAs with their input NPN transistors running high collector currents offer very low input noise [1] but pay a penalty by having higher input bias current and current noise. Fortunately, bias current and current noise are of little concern when interfacing an IA to a low impedance source.

Power Supply Constraints

Wafer processes place limits on the operating voltages of the integrated circuits built on those processes. Breakdown voltages are only one of the parameter tradeoffs that are made when an IC process is optimized for a particular purpose. Oxide thickness, silicon resistivity, and feature size are some of the parameters that determine the IC's operating voltage.

Devices built on standard linear bipolar and 15 micron CMOS processes usually handle a wide range of operating voltages while high speed linear and much higher density CMOS devices are limited to a much lower operating voltage. Table 9-1. lists a few recommended instrumentation amplifiers and difference amplifiers and their supply voltage range.

1. The Burr-Brown INA103 Instrumentation Amplifier achieves an input noise spectral density of 1nV/rt Hz @ 1kHz and only 2nV/rt Hz @ 10Hz.

Device Family Part Number	Device Function	Bipolar Supply Operating Range (±VDC)	Device Description
INA103	3 Op Amp Inst. Amp	9 to 25	Very Low Noise & Distortion
INA106	Difference Amplifier	5 to 18	G= 10V/V
INA111	3 Op Amp Inst. Amp	6 to 18	High Speed FET
INA116	3 Op Amp Inst. Amp	4.5 to 18	Electrometer Difet®
INA117	Difference Amplifier	5 to 18	±200V Common Mode Range
INA118	3 Op Amp Inst. Amp	1.35 to 18	Precision, Low Power
INA121	3 Op Amp Inst. Amp	1.35 to 18	Low Power FET
INA122	2 Op Amp Inst. Amp	1.1 to 18	Single Supply, Micropower
INA125	3 Op Amp Inst. Amp	1.35 to 18	On-chip Voltage Reference
INA126	2 Op Amp Inst. Amp	1.35 to 18	Low Cost, Micropower
INA132	Difference Amplifier	1.35 to 18	General Purpose
INA141	3 Op Amp Inst. Amp	2.25 to 18	Precision, G= 10 & 100V/V

Table 9-1. A Selection of Instrumentation Amplifiers and Difference Amplifiers.

Common Mode Voltage Range Requirements

Discussion & Table of recommended IAs for different CMV requirements here.

Improving Common Mode Rejection

Monolithic difference amplifiers are made with metal film resistors deposited on-chip. To achieve good accuracy the resistor networks are laser trimmed to match as accurately as possible. Inevitably, there is some shift in the trimmed resistors' value in the process of attaching the silicon die to its metal leadframe and in the plastic molding or ceramic package sealing process. Thus the final product does not have quite as high CMR as it was originally trimmed. This final CMR is fully specified in manufacturers' data sheets and it is fully adequate for most applications.

In critical applications where it is necessary to squeeze the last bit of CMR out of a difference amplifier or instrumentation amplifier, we can add a bit of circuitry to make some improvement.

Difference amplifiers can be trimmed by adding a fixed resistor into one input and then balancing the sum of that external fixed resistor and its series internal resistor with the sum of another external variable resistor and its series internal resistor. Thus a resistor and a potentiometer is all that is required. To maintain the internal network's tempco matching and to have a high resolution CMR adjustment, the external resistors should be small compared to the values of the internal resistors.

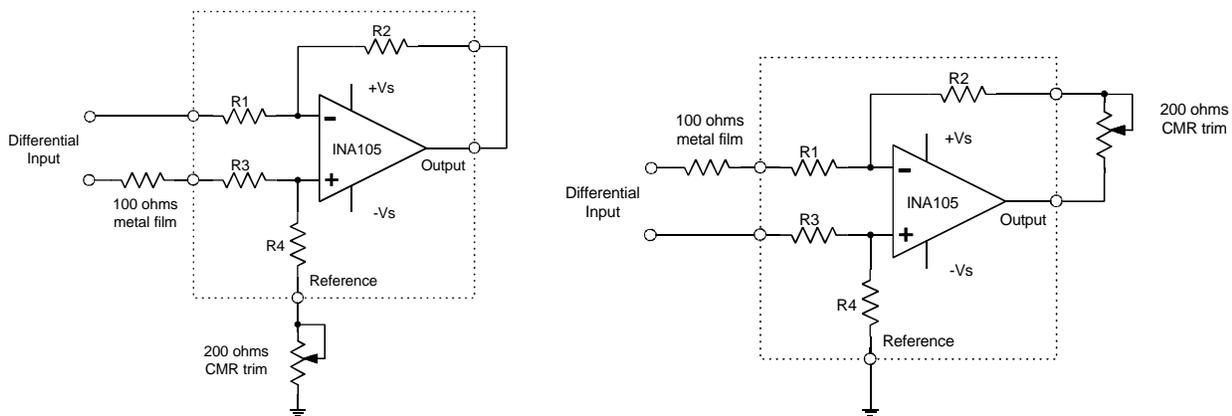Two similar diff amp trim circuits are shown in Figure 9-2 (a.) and (b.).

Figure 9-2. (a.) Difference Amplifier CMR Trim   (b.) Difference Amplifier CMR Trim-- Alternate

Two and three op amp instrumentation amplifiers must use a different approach as there is no resistor network on their inputs. Classical three op amp instrumentation amplifiers do have a diff amp (the third op amp) buried within their circuitry but there is no access to one or more resistors in the diff amp network. Fortunately, there is a Reference Pin that is accessible on all IAs which can be used to trim the amplifier's CMR as well as to trim the output offset of the IA.

Simply adding a variable resistor between the Reference Pin and ground (or common) may work if the resistor network has a too-low resistance in that leg. Statistically, one would need to subtract resistance from that leg in 50% of the amplifiers (in a large sample), but Murphy's Law says it will be Wrong Most Of The Time. In any case, how do we *subtract* resistance? Enter the "Negative Impedance Converter (NIC)." This useful little circuit transforms impedance by feedback. Varying the ratio of positive to negative feedback in an op amp, its input can appear to be either a positive (normal) resistor or a *negative* resistor. That's just what we need! See Figure 34.
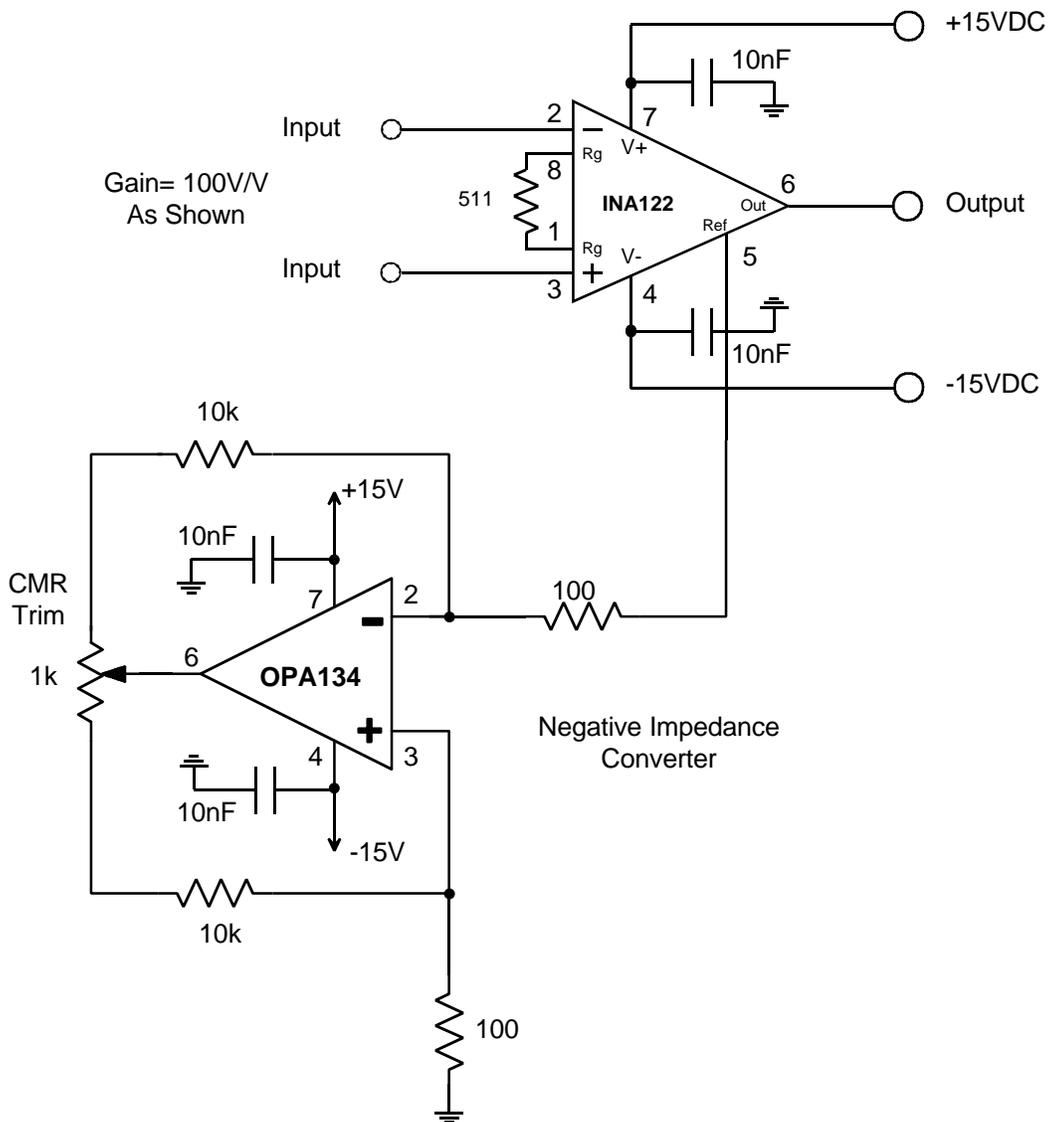
Figure 9-3. CMR Trimming a Classical Instrumentation Amplifier With A Negative Impedance Converter.

Two op amp instrumentation amplifiers offer many advantages– a large CMV range and low Iq, for example– but they do suffer from one disadvantage: lower high frequency common mode rejection. This weakness is inherent in the two op amp topology. Analysis of this architecture will show that one op amp has a higher noise gain than the other. Therefore, identical gain bandwidth op amps will display different bandwidth and phase characteristics. Common mode rejection depends on maintaining a precise match between the signals through both op amps. When the gain and phase mismatch becomes significant, the AC CMR degrades accordingly.

Since this AC mismatch is predictable and always in the same direction, we can take advantage of the "mirror image" characteristics of the circuit shown in Figure 35. The inputs to each INA126 two op amp instrumentation amplifier are connected out of phase (anti-parallel) and each output is connected to an INA134 difference amplifier.

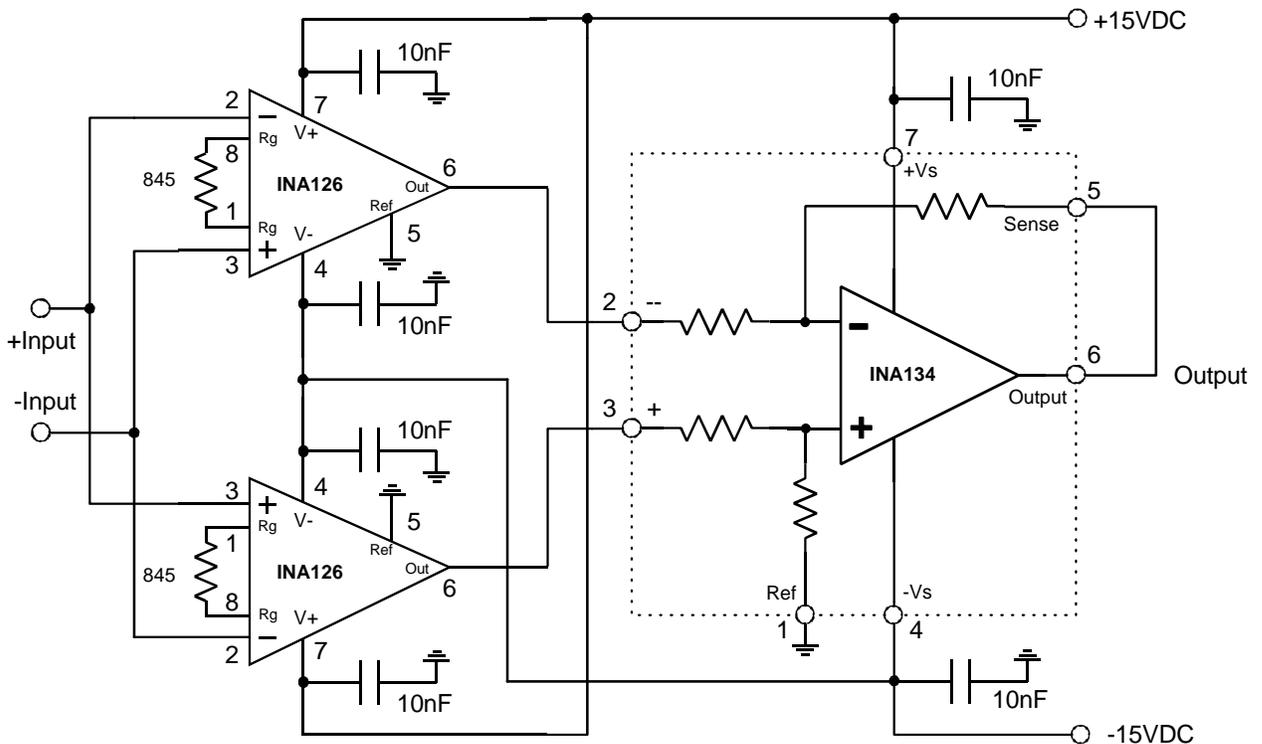Figure 9- 4. "Mirror Image" Amplifier Extends High Frequency CMR Of Two Op Amp Instrumentation Amplifiers.

Input signals are amplified 180 degrees out of phase by the two IAs and therefore add together in the diff amp. Each INA126 in our example is shown in a gain of 100V/V so our overall composite gain is 200V/V. AC CMR errors appear in- phase (0 degrees) at the IA outputs and therefore they are canceled by the diff amp. Although the input amplifiers' CMR degrades at high frequency, the CMR of the diff amp maintains the mirror image amplifier's overall CMR.

Ideally, we have the CMR of the input IA added to the CMR of the diff amp. In practice, this circuit doesn't quite achieve the composite DC CMR of 94dB + 90dB because the overall amplifier is unable to significantly improve on the diff amp's CMR alone.

A small DC gain mismatch in the input amplifiers may cancel residual errors in the diff amp's resistor network and make a slight improvement on 90dB, but it may also make it worse. At high frequency, things work to our advantage. At 10kHz, we see in the INA126 data sheet that its typical CMR is only about 48dB in a gain of 100V/V. The mirror image amplifier achieves far better CMR due to the INA134's 90dB CMR at 10kHz.

Measurements on random pairs of INA126 in the input amplifier stage of a mirror image show consistently almost 98dB CMR @ 10kHz for the INA126 & INA134 combination– an improvement of 50 dB! See Figure 36. for measured data on unadjusted CMR of both a single INA126 in a voltage gain of 100V/V and the mirror- image INA126 & INA134 combination in a voltage gain of 200V/V. In both cases, a CMV test signal of 1V peak- to- peak was used. Ordinary 1% metal film resistors were used to set the INA126 gain.
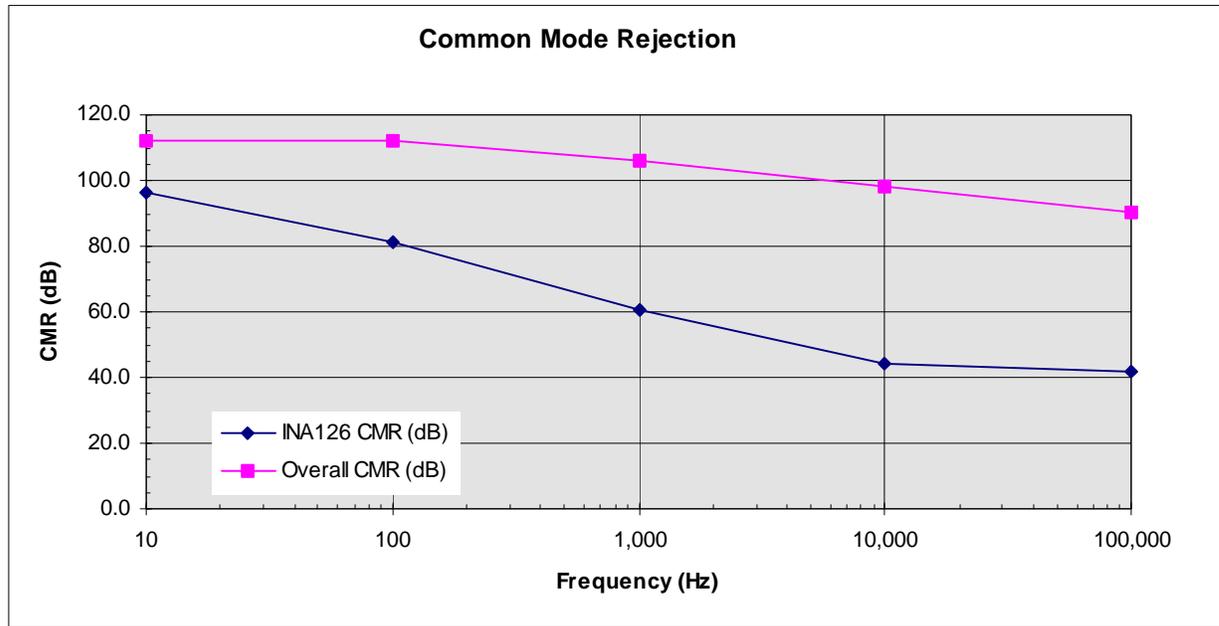
Common Mode Rejection

Figure 9-5. Measured CMR Of A "Mirror-Image" Instrumentation Amplifier vs. Single INA126.

In addition, the IA input noise is summed vectorially so the overall noise is improved by $\sqrt{2}$. If we had a large number of input amplifiers-- instead of only two-- we could also expect their input offset voltages, drift, and PSR errors to sum to zero [1]. An infinite number of amplifiers has no error.

While some other high quality three op amp instrumentation amplifiers can exhibit excellent high frequency CMR, this unique "mirror-image" circuit preserves the two op amp IA's very wide common mode range.

One may be tempted to extend this "mirror-image" technique to IAs other than the two op amp topology described above. Unfortunately, other topologies do not have the predictable CMR error-- always in the same direction-- and the results will be disappointing. To test this hypothesis, a similar mirror-image amplifier was built with very low noise INA103s at the input [2]. Instead of improving the CMR, the circuit was actually about 3-6dB worse than a single INA103!

A little thought and analysis will reveal why this is. This IA topology has a CMR error that is almost completely random-- it is determined by the match of internal thin-film resistors. In production, these resistors are laser-trimmed as accurately as possible and any residual error is uncorrelated between random INA103s.

1. See "Lunatic Fringe Amplifier", Burr-Brown OPA111 data sheet.
2. The Burr-Brown INA103 achieves an input noise spectral density of 1nV/sq rt Hz @ 1kHz and a CMR of about 93dB @10kHz in a gain of 100V/V.

Depending on phase, the CMR error of the two INA103s may either add or subtract in the output difference amplifier. If it subtracts CMR error and improves the overall CMR, it's just luck– some three op amp IA. pairs will improve CMR but an equal number will degrade CMR.

Input voltage noise does improve somewhat, however, as it is uncorrellated and adds vectorially while the signal, being 100% correlated, adds arithmetically. Thus the mirror- image amplifier's SNR is improved. Bandwidth and slew rate are doubled by using two mirror- image IAs instead of one. The output difference amplifier must also be capable of this level of performance if it is to be realized, however.

Bandwidth And Settling Time

Discussion & Table of recommended IAs for different BW & settling time requirements here.

Noise and Distortion

Audio applications require active components with especially low noise, low distortion, and wide bandwidth. These requirements are not necessarily restricted to audio applications, however, as some analytical applications also benefit from these same parameters.

Source resistance is the most important consideration in choosing a low noise amplifier. Simply selecting the lowest noise amplifier one can find may, in fact, be a very bad choice. When manufacturers speak of "low noise" they are almost always referring to low *voltage* noise rather than low *current* noise. It is important to choose the correct "low noise" amplifier based on the input signal's source resistance or impedance.

A simple rule of thumb:

1. For a <u>low impedance</u> signal source application, choose an amplifier with low <u>voltage noise</u> (e_n), i.e., a bipolar transistor input stage.

2. For a <u>high impedance</u> signal source application, choose an amplifier with low <u>current noise</u> (i_n), i.e., a JFET input stage.

3. For a <u>medium impedance</u> signal source application, choose an amplifier with a <u>balance</u> of both low voltage noise (e_n) and low current noise (i_n), i.e., a low power bipolar transistor input stage.

An amplifier's input voltage noise spectral density (e_n) is usually found in a vendor's data sheet, but sometimes its current noise spec is missing. It can be easily calculated if the bias current is known.

Input bias current spectral noise density (i_n) is defined as the rms noise current in a bandwidth of 1Hz that is centered on a specified frequency, such as 1kHz. It is related to input bias current by:

$$i_n = \sqrt{2qI} \qquad \text{where I= bias current and q= electron charge} = 1.602 \times 10^{-19} \text{ coulombs.}$$

When selecting a low input bias current bipolar amplifier, don't be fooled by thinking that an amplifier with low I_b must also exhibit low i_n. While this is usually the case, there are exceptions such as bias-current canceled amplifiers.

In the case of bias-current canceled amplifiers, we are seeing only a small portion of the actual input bias current of the amplifier. Only a remainder of the imperfect current cancellation technique appears at the device inputs– the actual bias current flowing inside the device may be

two orders of magnitude higher. Current noise is determined by the total internal bias current, not the tiny uncancelled portion that appears at the amplifier's inputs.

JFET input amplifiers offer very low bias current and, consequently, very low current noise. Although a JFET's voltage noise is higher than that of a bipolar transistor (especially when operating at high collector current), they can offer a quite acceptable e_n for medium to high impedance applications.

If one is concerned with only achieving very low input bias current, there is another choice—CMOS. Although a insulated- gate CMOS input device can achieve extremely low bias current, it exhibits a noise voltage characteristic that makes it a less attractive choice than a JFET input device. Although they are capable of good noise performance at 10kHz and above, the CMOS amplifier is handicapped by a high 1 /f noise corner frequency and so, for audio applications, are less popular.

Much progress has been made in the last few years on linear CMOS design techniques and on wafer processing to achieve low noise, stable CMOS devices. Where it was once unthinkable to consider using a CMOS amplifier, they are now used routinely. All things being equal (currents, device geometry, second-stage contribution, etc.), the JFET is still a better choice for lowest noise in the 20Hz to 20kHz audio range.

Best FET- input noise performance (e_n) is achieved by large- geometry, high transconductance, N- channel JFETs. The more commonly encountered P- channel device is easier to integrate in an IC process, but it has a higher noise voltage than the higher carrier mobility N- channel.

Instrumentation amplifiers and difference amplifiers that are recommended for audio applications are shown in Table 9-2.

Model Number	Inst. Amp	Diff.Amp	en@ 1kHz (nV√Hz)	in@ 1kHz (pA√Hz)	THD 1kHz (%)	Bandwidth (kHz)	Gain Range (V/V)	CMR 60Hz (dB)	Slew Rate (V/us)
INA103	X		1	2	0.0009 (G=100V/V)	800 (G=100V/V)	1 to 1000	125	15
INA111	X		10	0.0008	0.002 (G=10V/V)	2000 (G=10V/V)	1 to 10000	110	17
INA134		X	52*	*	0.0005	3100	1	90	14
INA137		X	26*	*	0.0005	4000	0.5 or 2	90	14

* Difference amplifier noise is typically specified as the total output noise in a bandwidth of 20Hz to 20kHz, and includes both op amp voltage and current noise and resistor noise. Specifications for the INA134 and INA137 output voltage noise are 7μV rms and 3.5μ rms (G= 0.5V/V) respectively.

Table 9- 2. Instrumentation Amplifiers and Difference Amplifiers Recommended for Audio Applications.

A data sheet is available for each model and it should be consulted for detailed technical information. Check the BB web site often for new models that may not be listed in the table.

"Rail-To-Rail" Input & Output Swing

Discussion & Table of recommended IAs for R-R input & R-R output swing requirements here. Mention OPA340 design for 2 op amp IA.

First, a caveat: there are no industry-accepted definitions of what constitutes rail-to-rail (R-R) swing– input or output. With that in mind, we will plunge onward with this rather arbitrary R-R definition: a capability of swinging within a few hundred millivolts of the rail while also maintaining the device's important specified parameters, such as linearity.

This last requirement is important to note. An instrumentation amplifier's output swing is *not* simply how close you can drive it to the supplies (or to ground in single supply applications). The IA linearity will be seriously degraded well before its output stage saturates because the internal op amps have run out of gain when driven into the supply rails.

Requirements for input swings to the negative rail were discussed earlier in the current shunt section of this Handbook. Input swing to the positive rail allows the designer considerable flexibility in the choice of devices for his application.

As an illustration of how this can be advantageous, consider the problem of measuring current with a high-side shunt in a 12VDC battery-operated system. A further requirement is that under a load dump condition, the line voltage can rise to 28V. Under these conditions, the difference or instrumentation amplifier must be capable of operating with a 12 to 28VDC common mode voltage. If only digital system +/-5V power supplies are available, we can use the -5V supply for the amplifier's negative supply and the DC line voltage for our amplifier's positive supply as well as to connect to one side of the current shunt resistor. See Figure xx.
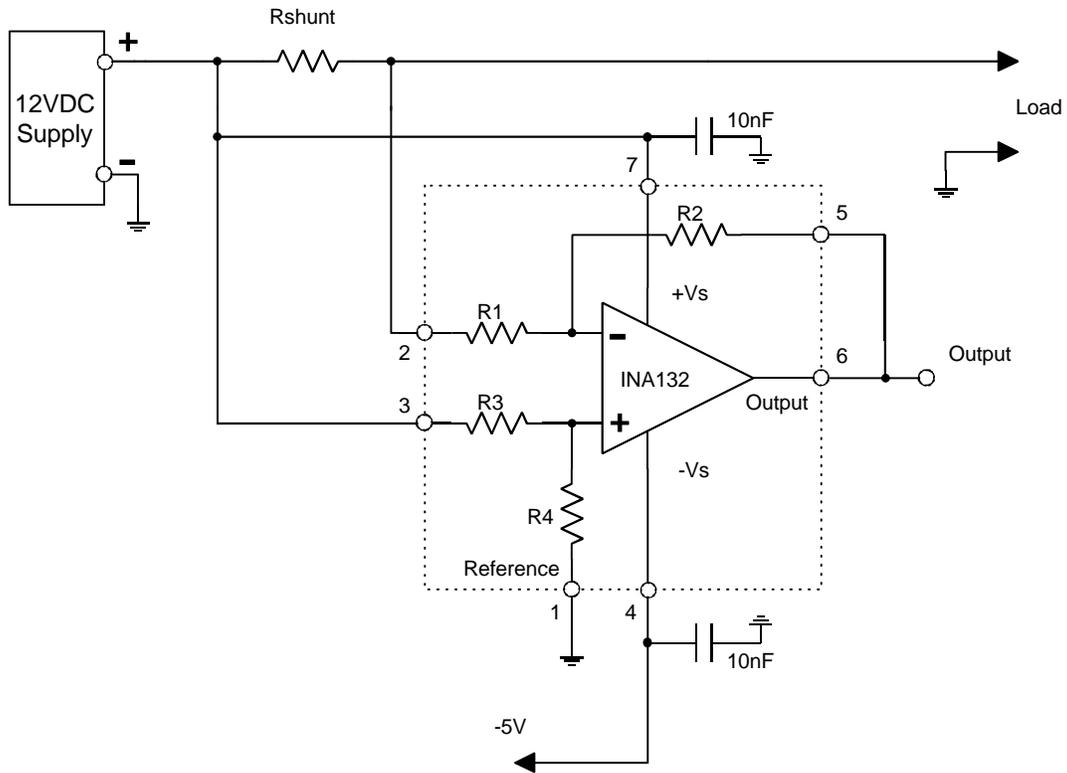
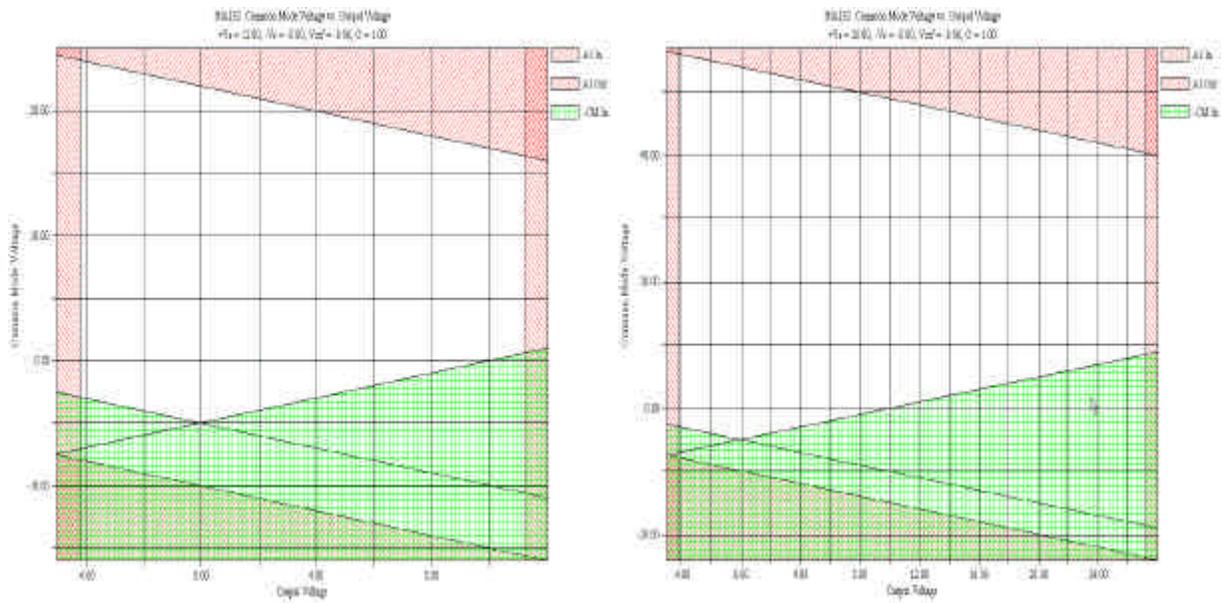Figure 9-6. Unity Gain Difference Amplifier Common Mode Range Automatically Follows High-Side Shunt Voltage.

Figure 9-7. (a.) INA132 Common Mode Range With +12/-5VDC Supply   (b.) INA132 Common Mode Range With +28/-5VDC Supply

As the large white areas in the CMV plots of Figure xx. illustrate, our allowable common mode input voltage "tracks" the INA132 positive supply. The difference amplifier output voltage can swing from zero to +10V full scale with an input CMV range that is well above its input and positive supply.

This is unlikely in this case, as we are using a unity- gain difference amplifier to sense the voltage across the shunt resistor and, to minimize the voltage drop across the shunt , its resistance will usually be low. Small signal unity- gain operation is feasible if the INA132 input offset voltage (75μV typ, 250μV max) and drift (1μV/C typ, 5μV max) are within an acceptable measurement error budget.

Using a new- generation precision CMOS op amp such as an OPA340 or the dual version, OPA2340, and a few precision resistors, it is possible to easily build a two op amp instrumentation amplifier with outstanding R-R input and R-R output swing. The circuit diagram of this gain- of- 10 instrumentation amplifier is shown in Figure 9- 8.

The ability of the OPA340 and OPA2340 to have its input swung 300mV *beyond* ground or its positive supply together with its output capability of swinging to within 10mV (R $_L$= 10k) of either ground or its positive supply produces the best R-R single supply IA currently available!

If desired, wider bandwidth and improved high frequency CMR can be obtained by simply substituting faster OPA350s (or a dual OPA2350) for the OPA340s (or dual OPA2340) op amps. The OPA350 features a much higher gain bandwidth product than the OPA340 -- 35MHz vs. 5.5MHz GBW.
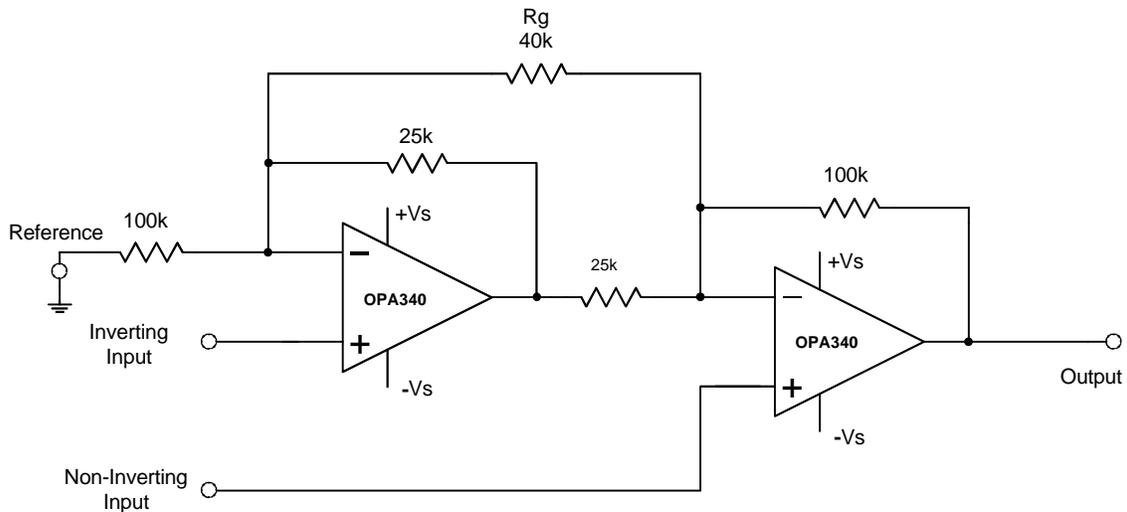

Figure 9- 8. A R-R Input And R-R Output Instrumentation Amplifier With a Gain Of 10V/V.
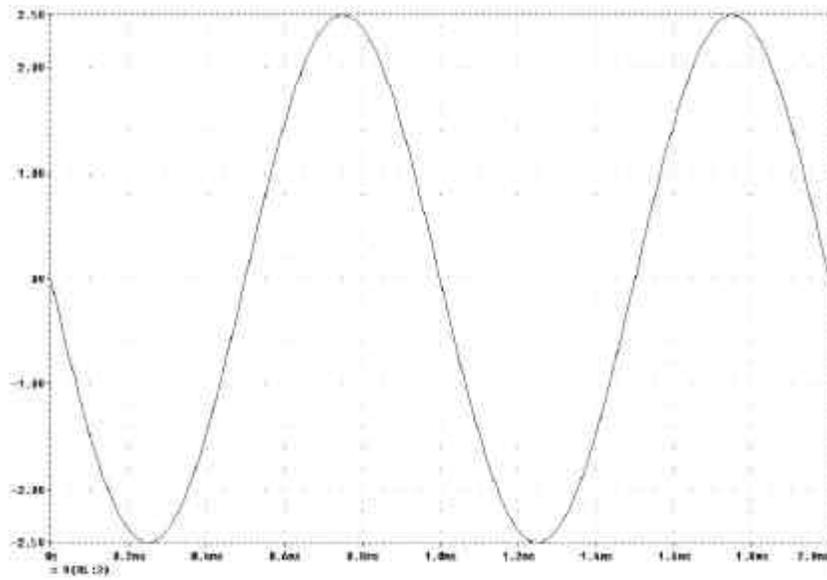
Figure 9- 9. OPA340 Instrumentation Amplifier R-R Output Swing Into A 10k Load With Vs = ±2.5VDC And Vin= 250mV p-p.

RFI Problems

Input Rectification—the Most Common Problem

In days gone by—before computers— kids were frequently introduced to the world of electronics by building a "crystal set". These simple radio receivers consisted of a long wire antenna, a tuned circuit (the coil was usually wound on a Quaker Oats box!) a good ground connection, and a simple detector[1]. Early experimenters used a Galena crystal and "cat whisker" as a detector and spent endless hours searching for the most sensitive spot on the crystal.

With a good antenna and earth ground connection, a nearby AM radio station would give an audio signal adequate to drive a pair of high impedance headphones. In the author's experience, given a very strong AM signal, the headphones could even be heard across a room. Of course, it helped to have AFN Frankfurt—running 150 kW on 873kHz—only 20 kilometers away!

The introduction of commercial semiconductor diodes made life easier for crystal set builders— the germanium 1N34 eliminated all the tweaking with the cat whisker but it also eliminated a lot of the challenge! Silicon diodes such as the 1N914, in turn, eventually replaced the germanium ones. Interestingly, as the crystal set detectors evolved, they became less sensitive.

Looking at the detector diode's I-V curve (Figure xx.) shows why. Since the detectors in these simple receivers operated at "zero bias", a large RF voltage was required to reach the diode's conduction "knee" and cause current to flow into the headphones. As the semiconductor material evolved from galena[2] to germanium to silicon, the diodes' conduction 'knee' voltage increased due to higher bandgap energy and a larger and larger RF input voltage was required— i.e., less detector sensitivity.

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1. The old-fashioned galena detector is a rather interesting device. Galena is a naturally- occurring lead sulfide ore in crystalline form.   When touched lightly by a sharp pointed wire, a quantum- tunneling junction can be formed if the experimenter is very careful. Little did we kids know how "high tech" we were back in those halcyon days of yesteryear.

[2] Galena and other materials, such as iron pyrite ("fools" gold") and silicon carbide (Carborundum) were used as RF detectors as far back as 1901.
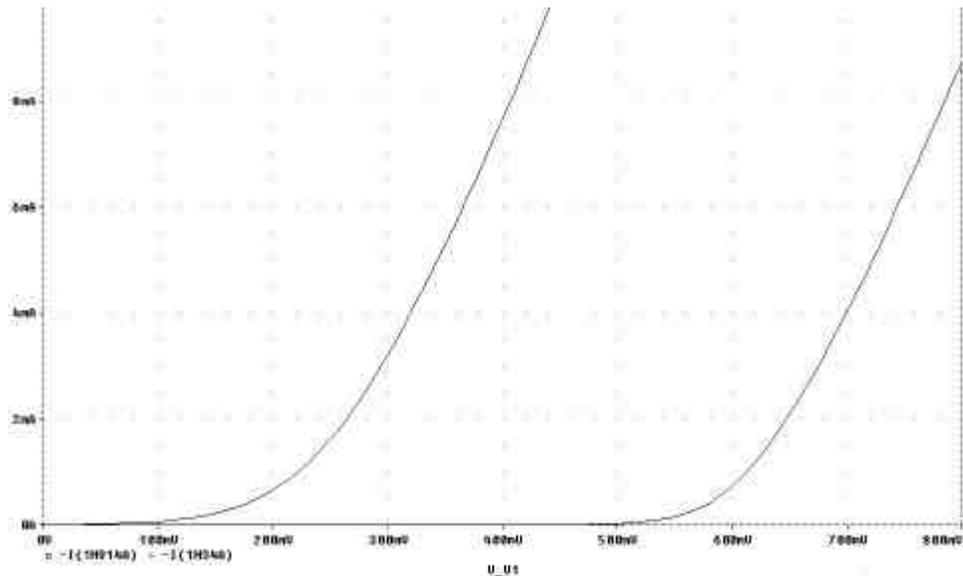
Figure 10-1. I-V Curves For Germanium and Silicon Diodes.

Forward biasing a silicon (highbandgap) detector diode greatly improves its RF sensitivity. In fact, we find a sensitive RF detector in the base-emitter junctions of the input transistors of an op amp. Any significant RF on these junctions will be rectified, causing a net DC shift in its operating point. This appears as an input offset voltage change. If the RF signal is an amplitude modulated carrier (AM), its modulation envelope will appear and a corresponding AC signal can be seen with an oscilloscope on the amplifier's output. Feeding the AC signal into an audio amplifier and loudspeaker can sometimes reveal the source of the RF interference. AM broadcast stations are frequently heard but FM stations generate only a DC offset.

Other sources can sometimes be identified by listening to a demodulated RFI carrier—a loud buzz can be caused by SCR AC line controllers or the video carrier of a TV broadcast station, a regular buzzing that comes and goes periodically can be caused by pulse radar, etc. A good spectrum analyzer is still the best method of identifying the RFI source, though.

Typical Instrumentation Amplifier Swept-Power RFI Tests

To evaluate RFI sensitivity and test the effectiveness of input filtering, a series of RF tests were run on a Burr-Brown INA129 instrumentation amplifier. In each test an RF generator was connected to both IA inputs to apply a common mode RF signal. The generator input was AC coupled to eliminate possible generator DC offset voltage and it was properly terminated with a 50 ohm (actually, it was 49.9 ohms) load. The resistor R3 provides a bias current return path to ground.

 The generator was swept over a range of 100kHz to 1GHz in steps of 1x, 2x, 5x, and 10x. Power levels were varied over a range of –100dBm to 0dBm and the amplifier's DC offset was

recorded. After subtracting the baseline DC offset at –100dBm, the INA129's delta Vos was plotted vs. frequency.

The test results of the **_unfiltered_** INA129 RFI response test circuit (Figure xx) is shown in Figure xx. Equivalent input offset can be found by simply dividing the output offset voltage by the test circuit's amplifier gain of 200V/V.

As one might expect, this test confirms that high RFI levels can create serious DC offset problems.
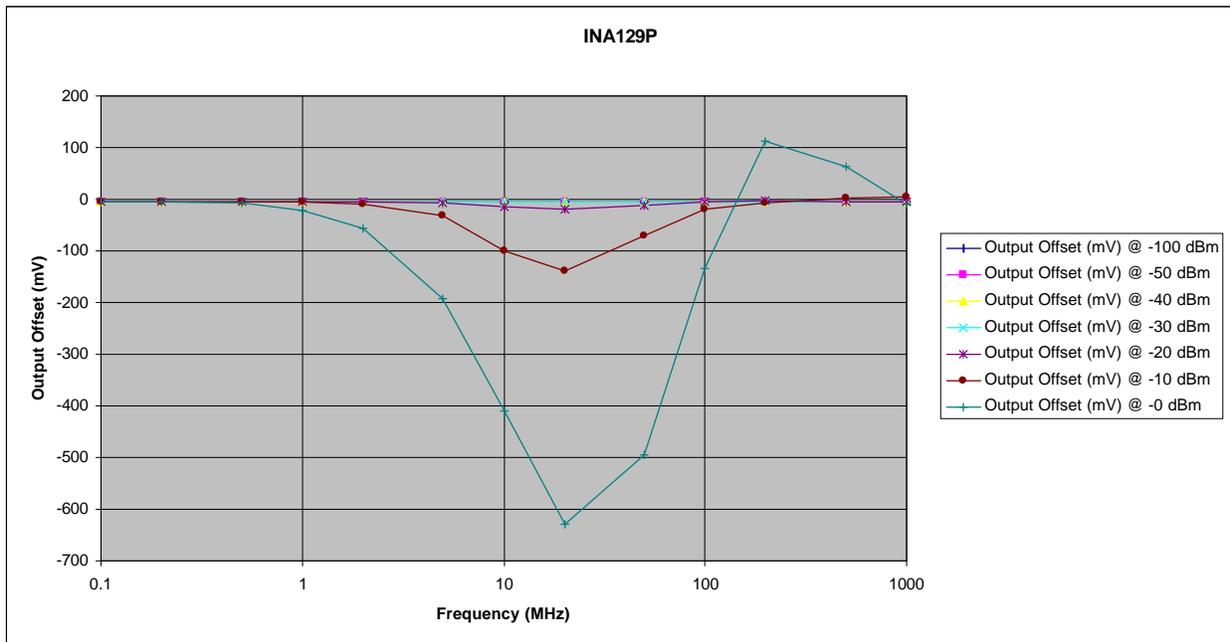

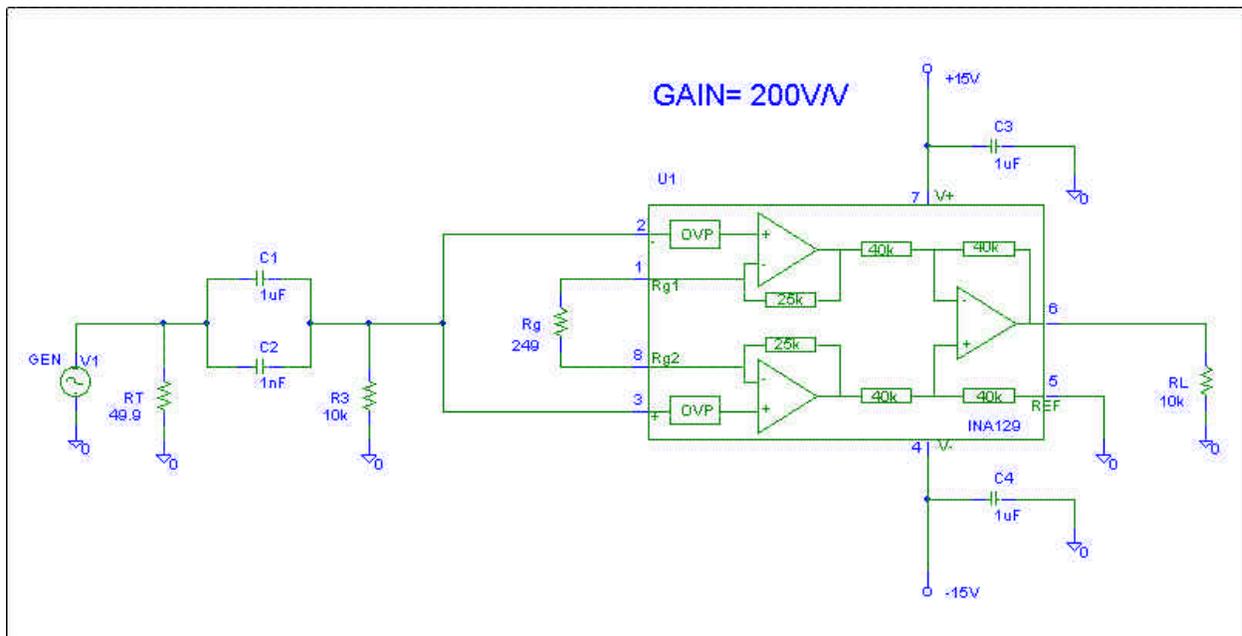Figure 10-2. **Unfiltered** INA129 Input Offset Shift: Note +200, -700 mV Vertical Scale.

Figure 10-3. **Unfiltered** Input INA129 RFI Test Circuit.

Input RFI Filtering

As can be seen in the previous test, reducing the INA129 RFI input power level to below -30 dBm gives dramatically lower RFI- induced input offset shift. This can be accomplished by adding a simple low- pass filter (LPF) to the instrumentation amplifier's inputs. But how effective are filters?

Many LPF configurations are possible but we will limit out investigation here to only the simplest implementations. More complex and effective filters are certainly possible and these may be required for extremely severe RFI environments.

Analog Devices in their AD620 data sheet (Figure xx.) recommends a simple RFI filter circuit that makes a worthwhile improvement in RFI performance. In essence, this is a simple one- pole RC low pass filter. Although AD recommends that C1 & C2 be no larger than 150pF, a low frequency filter cut off can be obtained by using suitably high values of R1 & R2. As resistor values are limited by IA input bias current errors, however, very low cut off frequencies cannot be obtained.

One important point to remember from our previous discussion of input filters is that $R1 \cdot C1$ should be as closely matched to $R2 \cdot C2$ as possible to avoid converting common mode voltages to a differential voltage.
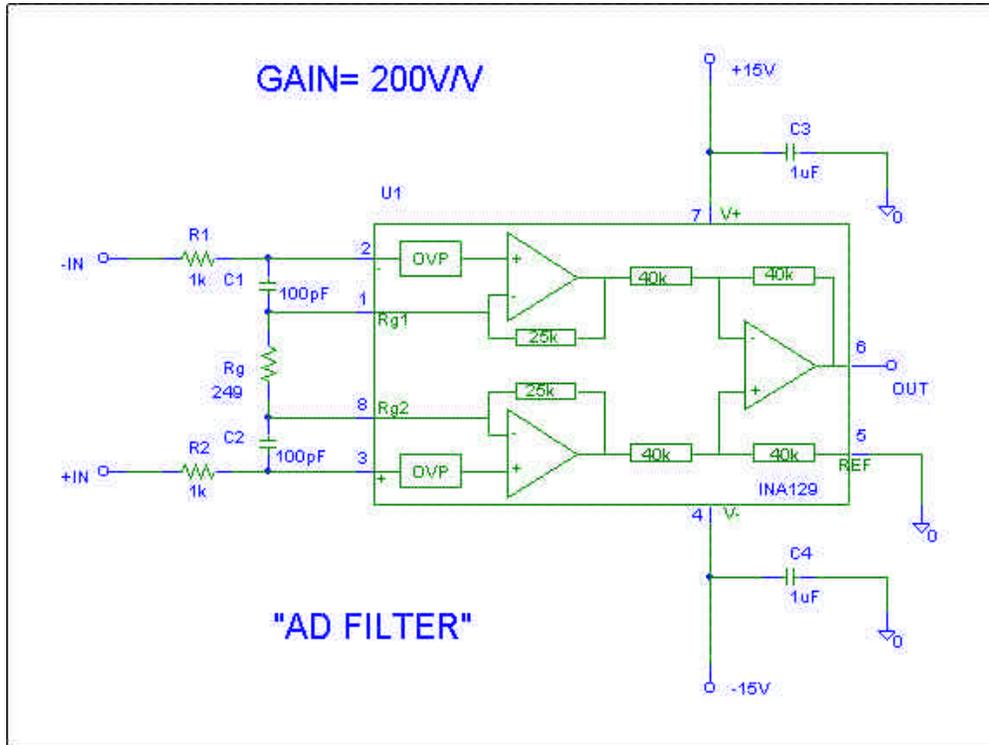
Figure 10- 4. One Approach To Input RFI Filtering–The "AD Filter."

Test results for a Burr- Brown INA129– similar to an AD620–with an "AD Filter" added are shown in Figure 10- 5. A dramatic improvement in the amplifier's response to input RFI is clearly seen.
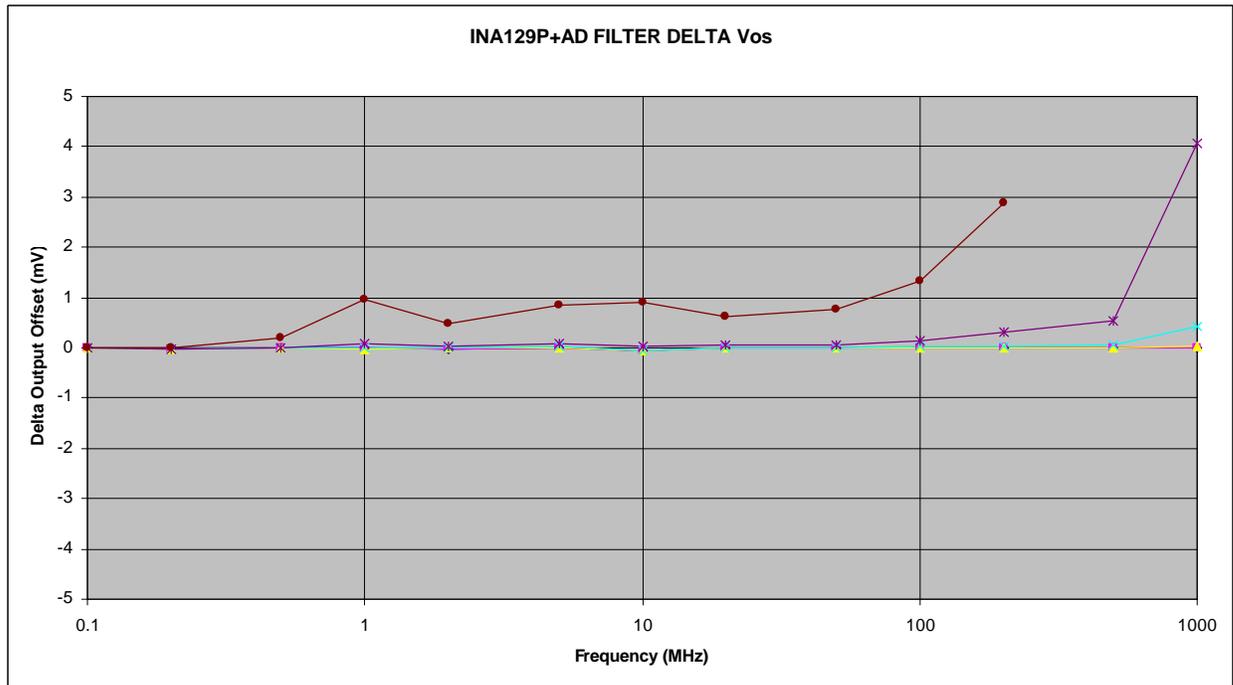

Figure 10- 5. Input Offset Shift With Analog Devices' Filter: Note ± 5mV Vertical Scale.

Both the INA129 and AD620 are current-feedback IAs and a connection of the filter capacitor between the IA input and its gain setting resistor is effective, in this case, because of the wide bandwidth of the current-feedback input stage. In a conventional IA, this approach is not nearly as effective, however.

An improved performance LPF can be realized by taking a different approach. This is an all-passive approach that requires one more capacitor than in the "AD Filter". In addition to better RFI performance, this type of filter (Figure xx.) is effective on either current- or voltage-feedback instrumentation amplifiers and it is less sensitive to RC matching.

To distinguish the two filter types, the author has arbitrarily dubbed this approach the "BB Filter." This type of filter has been used in industry for years and certainly no claims of invention are made here.
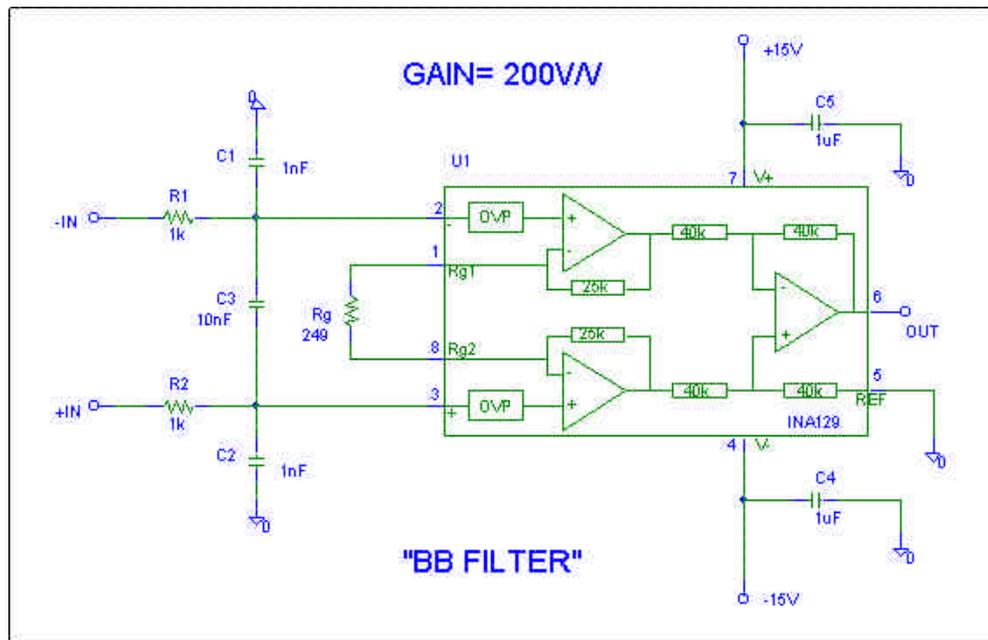

Figure 10-6. An Improved All Passive Input Filter–Dubbed the "BB Filter."

Test data (Figure 10-7.) for the "BB Filter" illustrates how much of an improvement we see with a "BB Filter" filter. Note that the graph scale is only± 2mV for this test! This type of filter– with appropriately scaled resistors– is highly recommended for any instrumentation amplifier.

Good layout techniques and a solid ground plane are a must if the filter is to remain effective into the VHF/UHF range. Short capacitor leads will minimize inductance and their push self-resonant frequency to as high as possible. Surface mount components have an advantage; due to their small size, lead inductance is less of a consideration. NPO or COG ceramic capacitors are particularly effective in RFI filters but they are practical only up to about 1000pF.
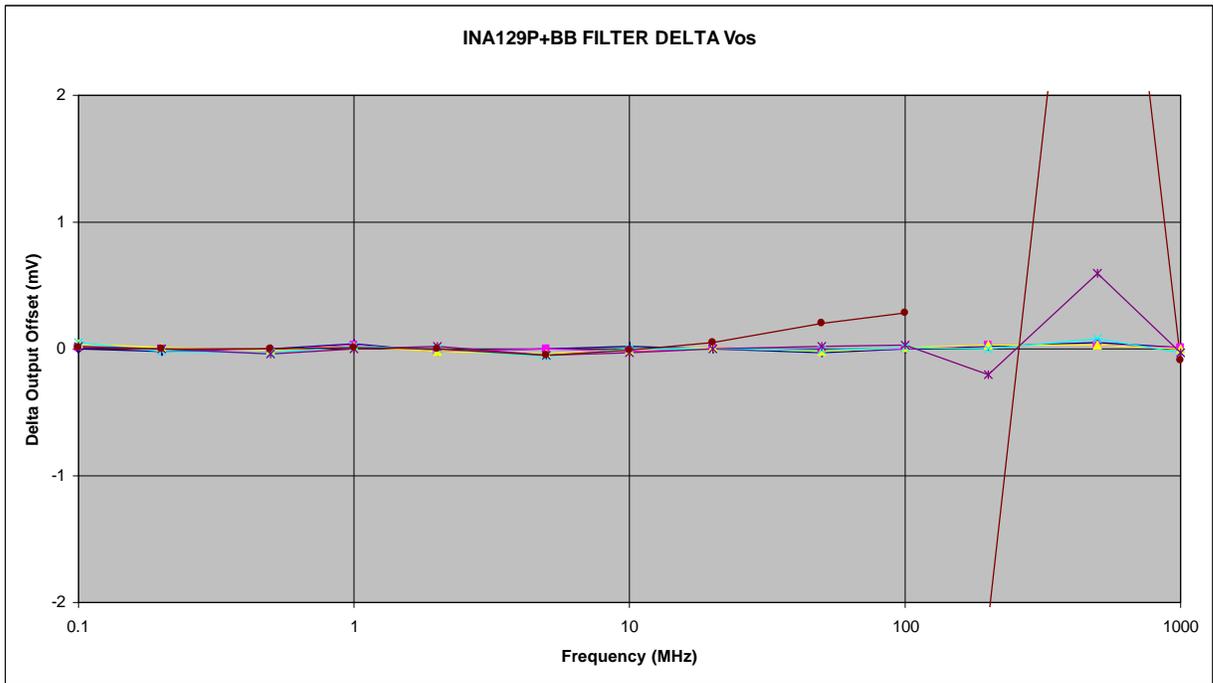
Figure 10-7. Input Offset Shift With Improved "BB Filter": Note ± 2mV Vertical Scale.

Circuit diagrams of both types of instrumentation amplifier filter test circuits are included in Figure 10-8 for reference.

Note: The term "dBm" is a logarithmic measure of power (referred to 1mW) that is commonly used in RF. For example, 0dBm is equal to one milliwatt and is equivalent to 223.6 mV rms across 50 ohms. Here is a quick reference table showing the conversion:

RF Power (dBm)	Voltage Across 50 Ohms (mVrms)
-50	0.707
-40	2.24
-30	7.07
-20	22.36
-10	70.7
0	223.6

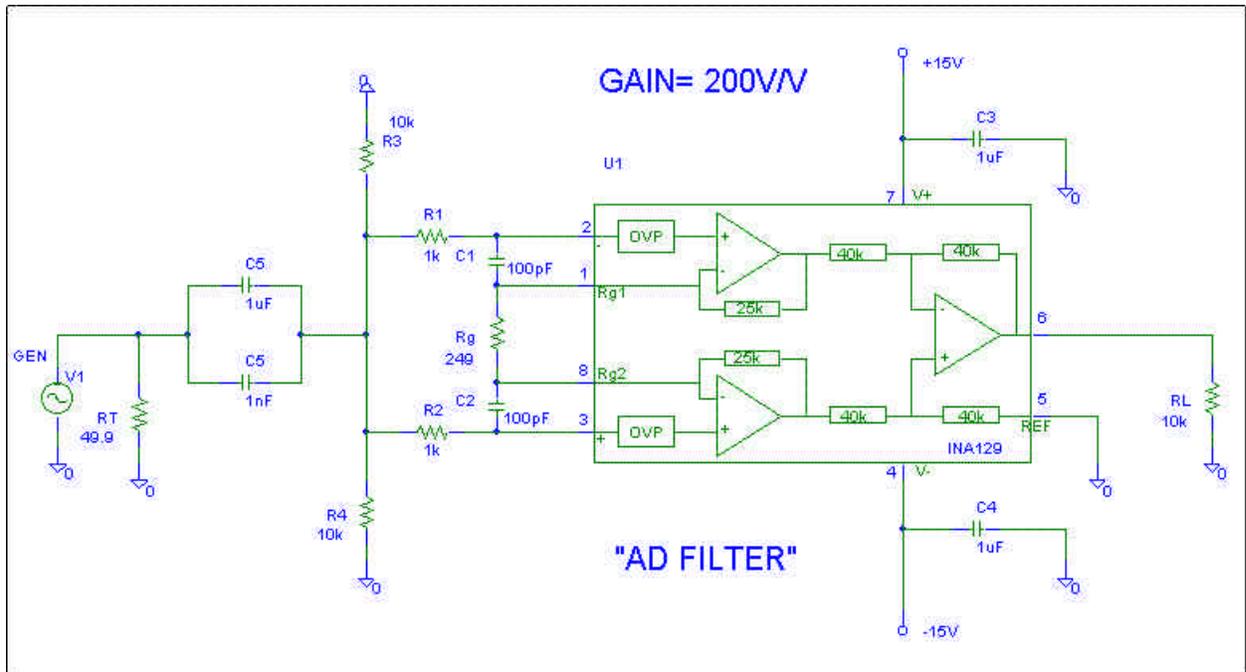Table 10-1. RF Input Power To RF Input Voltage Conversion Table.

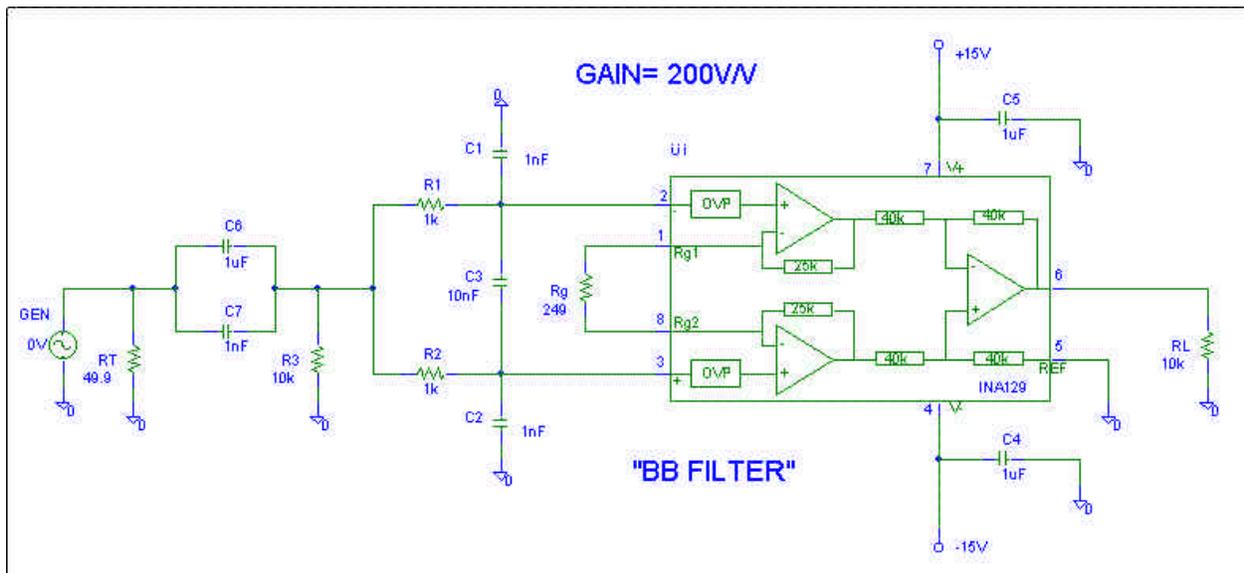Figure 10-8. RFI Test Circuits For Measuring Offset Shift Of INA129 With "AD" LPF.

Figure 10-9. RFI Test Circuits For Measuring Offset Shift Of INA129 With "BB" LPF.

Other RFI Considerations

While input filtering will solve most conducted RFI problems, other wires are capable of conducting interference into a circuit as well. Power supply lines should be filtered by including a small inductor of about 10uH to 1000uH in series with each power lead. Don't ignore the outputs, either. An inductor in the output can also be helpful in some cases.

Murata-Erie makes a very nice little RFI noise filter that has excellent broadband attenuation and can handle 10 amps. Their BNX002-01 achieves a minimum of 40dB attenuation over a range of 1MHz to 1GHz and it also provides ground return filtering to minimize noise due to ground loops. One of these on each power lead or even on the output of a two-wire 4-20mA current transmitter can do wonders.

In short, **filter __everything__** that goes in and out of your housing or PCB.

Complimenting the practice of RFI filtering is the practice of shielding. Enclosing your circuit in a conductive housing will reduce RFI problems by making your filtering more effective. Creating a "Faraday cage" around your circuit will eliminate problems from radiated RFI as it presents an effective barrier to RF fields.

Radiated RFI consists of both an E-field (electric) and an H-field (magnetic). Shielding effectiveness will depend on what type of shielding material is chosen and which field (E or H) is the culprit.

Only a thin conductive shield is needed to be highly effective against E-fields. Aluminum or copper foil offers high attenuation, as does virtually any other type of sheet metal. Sprayed-on conductive coatings containing graphite or nickel on a plastic housing can also be acceptable.

H-fields are more difficult to shield effectively with non-ferrous material. Fortunately, steel is readily available and inexpensive. Low frequency shielding can be accomplished very effectively with a high permeability material such as Mu-metal. Commonly used materials such as aluminum or copper foil offer almost no attenuation for low frequency H-fields.

This is due to the fact that non-ferrous metals such as copper, aluminum, or brass attenuate H-fields by induced eddy currents and at low frequency their required thickness becomes prohibitive.

Miscellaneous Applications Circuits

Absolute-Value Amplifier

Occasionally you may need an amplifier that gives you an output that is proportional to the *absolute value* of its input signal. That is, its output is always positive (or negative)– irrespective of the polarity of the input signal! This isn't as difficult as may sound. One type of absolute value circuit is shown in Figure 11-1.
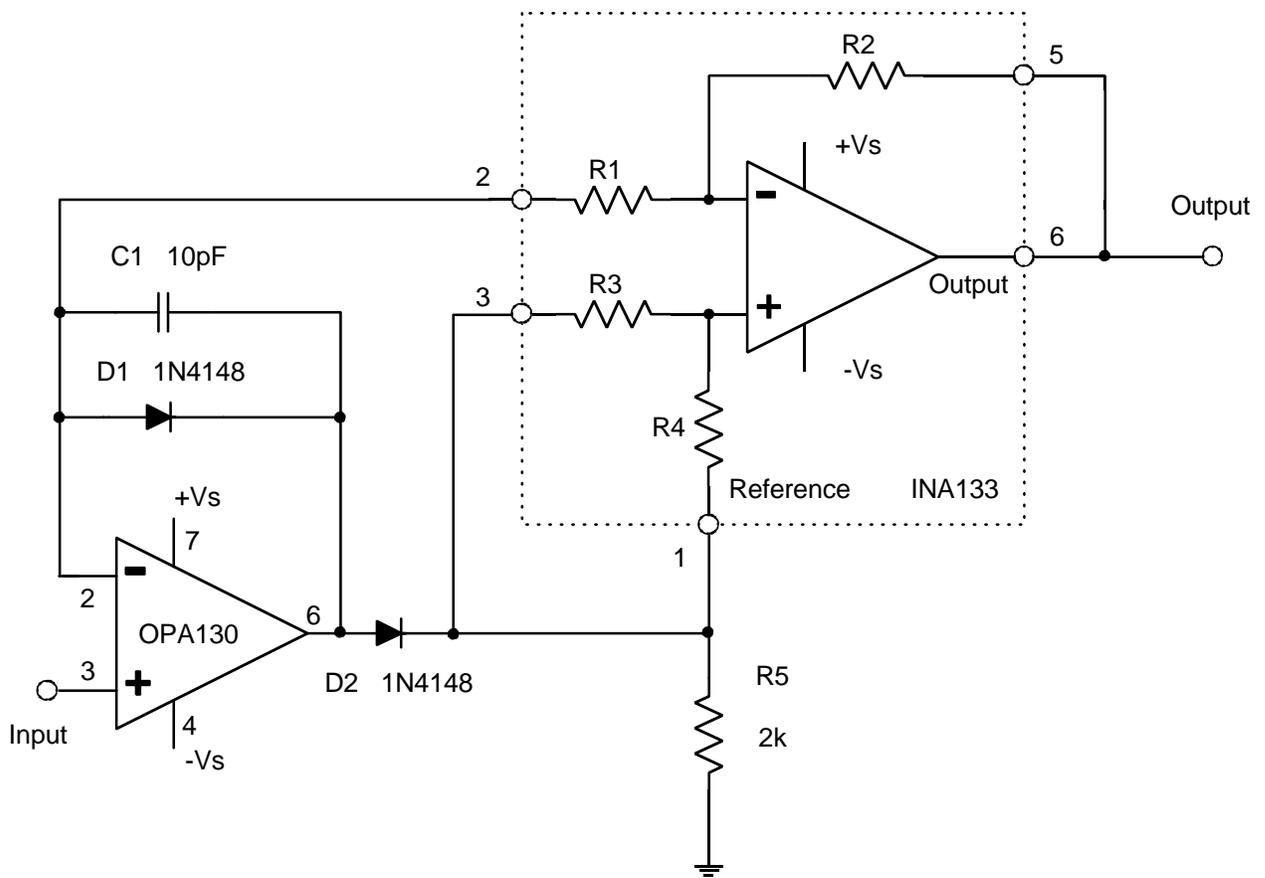

Figure 11-1. Unity Gain Absolute Value Circuit– Positive Output.

A FET op amp is used to minimize bias current errors and small-signal silicon diodes "steer" the op amp's output to the appropriate difference amplifier input This way, the INA133 output remains positive for any input voltage polarity. The output can be made negative by simply reversing D1 and D2.

A small capacitor, C1, provides high frequency feedback to maintain loop stability in the input
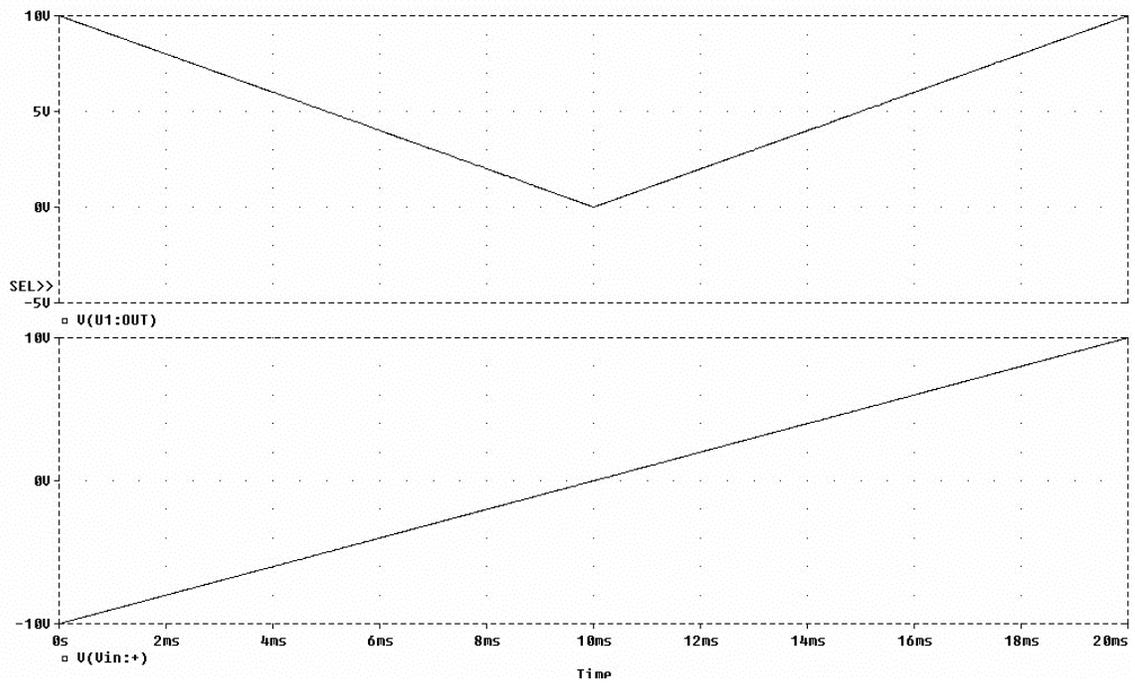
op amp.

Figure 11‑2. Absolute-Value Circuit Input (Bottom Curve) vs. Output (Top Curve) Transfer Function. The Amplifier's Output Is Always Positive.

Avoiding Difference Amplifier Pitfalls

Adding External Resistors—Don't !!

Looking at the circuit diagram of a difference amplifier, it appears easy to simply add external resistors to each input and thereby achieve an increased common mode voltage range. Adding a couple of 1% metal film resistors to the inputs (shown in Figure 12-1.) and measuring the resulting CMR performance reveals a precipitous drop in the circuit's common mode rejection.

Thinking the 1-% resistor tolerance may be causing the problem, you find a pair of precision 0.01% matched resistors and try again. The CMR is still terrible– *what's going on*?

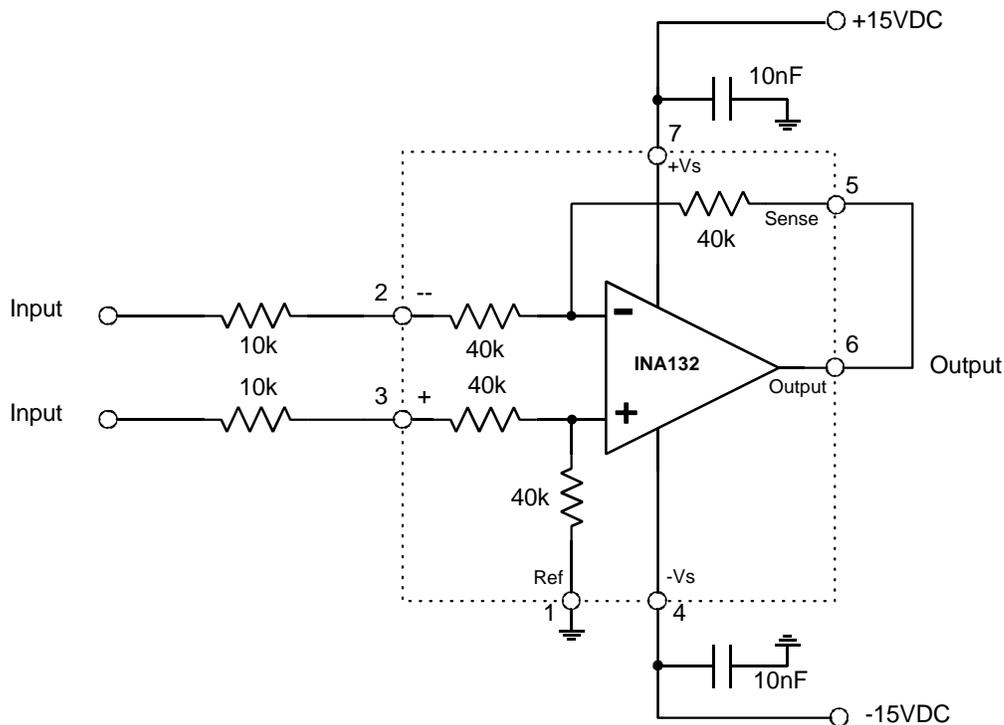There is a good explanation, but it isn't obvious from Figure 12-1.

Figure 12-1. How To Ruin Your Difference Amplifier's Common Mode Rejection– Add External Resistors.

The problem with this circuit is that the "40k resistors" shown in the INA132 block diagram are only their nominal values; *not* their exact resistor values. In actuality, these on-chip metal film resistors are laser trimmed to a very high precision <u>ratio match</u> in order to achieve the amplifier's CMR and gain specification, but not to a high accuracy absolute value. Nominal values may, in fact, differ from the actual trimmed value by as much as 5%.

Calculating the effect of adding 40k resistors to a difference amplifier with resistor network values of 38k to 42k will clearly show the folly of adding external resistors. CMR drops to a very low value!

To investigate these effects, we can perform a circuit simulation on two difference amplifier circuits, each consisting of a high precision OPA227 op amp and a matched-ratio resistor network. In one case, we add precisely matched external 40k resistors to the diff amp inputs while in the other case it remains a conventional input connection (Figure 12-2.).
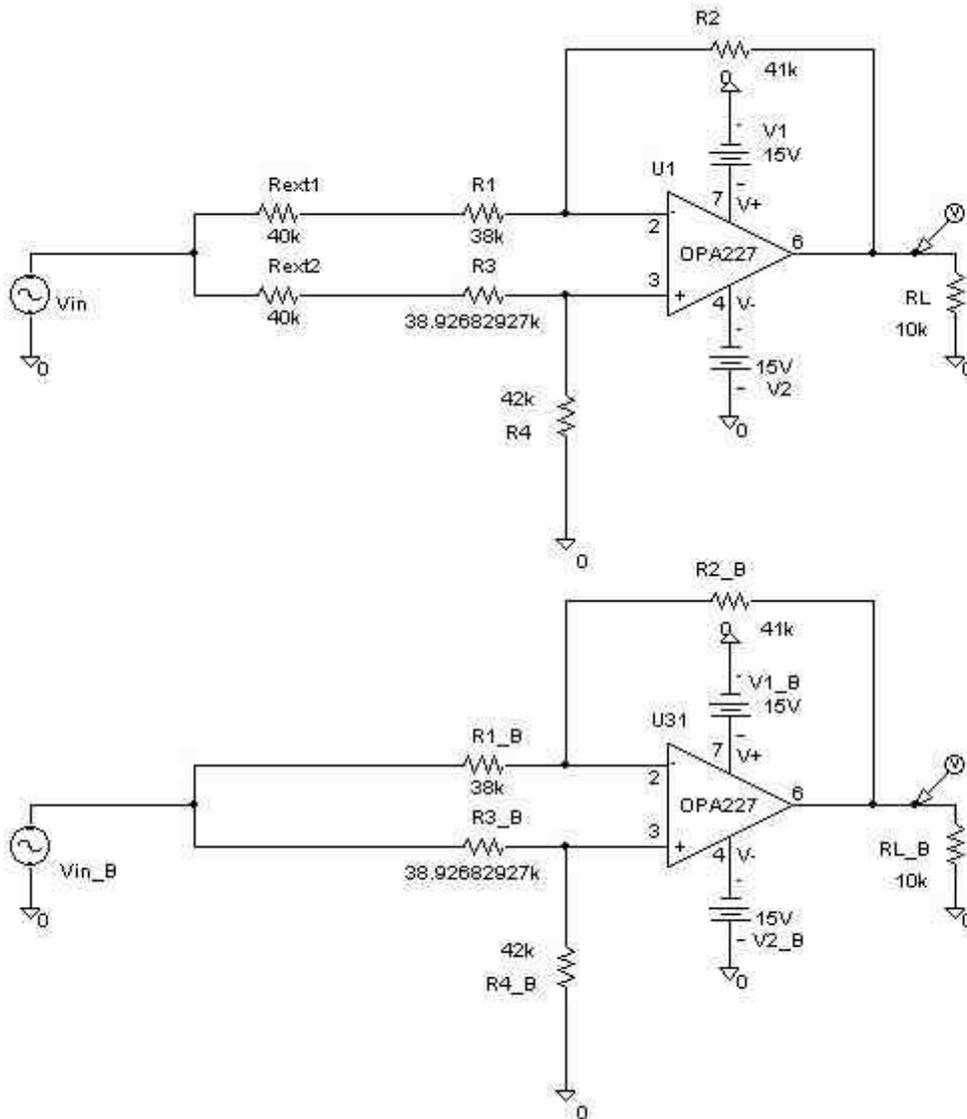

Figure 12-2. Circuit Diagram To Investigate Effects Of Adding External Resistors To A Ratio-Trimmed Difference Amplifier.

 A 1VAC signal source is connected to each of the diff amp's inputs as a common mode signal and we look at the output error voltage– common mode rejection.

Figure xx. reveals how much CMR is sacrificed by adding even perfectly matched resistors to a ratio-trimmed difference amplifier. Remember that we are doing a simulation here; we can have perfectly trimmed resistors and no parasitics so the theoretical CMR of the OPA227 diff amp is very, very high. In practice, we might achieve 120 dB with great care, super precision discrete resistors, a good layout, and lots of luck.

Even so, adding external 40k resistors and dropping the CMR to less than 60 dB is a disaster.
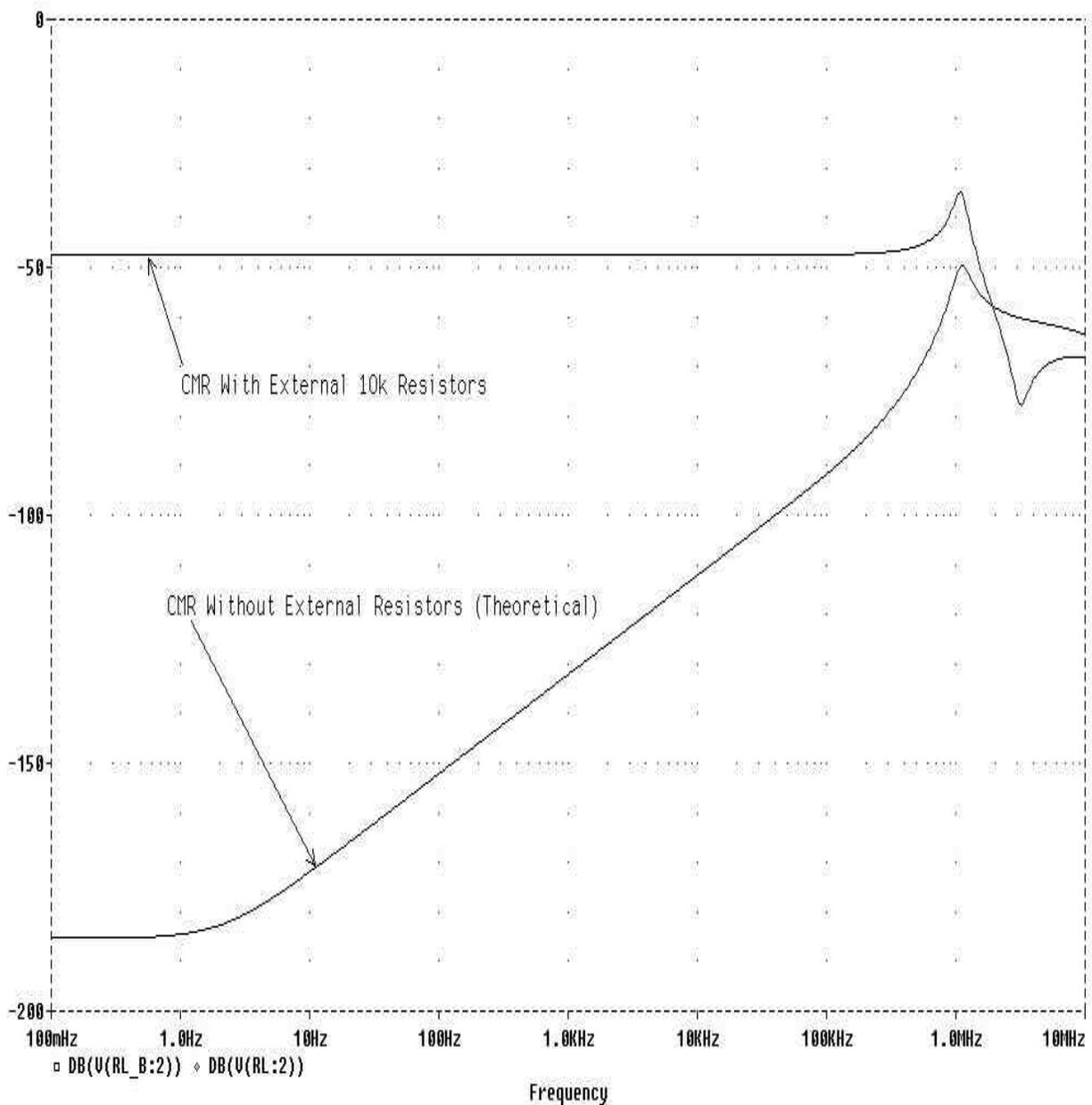
CMR With External 10k Resistors

CMR Without External Resistors (Theoretical)

□ DB(V(RL_B:2)) ◇ DB(V(RL:2))

Frequency

Figure 12- 3. Theoretical CMR With And Without External 40k Resistors– A Big Difference !.

If a difference amplifier's resistor network were trimmed to exact values, adding external resistors to increase its CMV range would be a viable option, so one might ask, "Why not trim the resistor network to absolute values?" The answer is "Cost." Laser trimming adds cost to the final product and, to keep the price low, only a ratio trim is performed.

If you are the type who can't give up this idea easily, here are a few things to consider:

1. You might be extremely fortunate and have the external resistor tolerance compensate the internal network's absolute value mismatch. Don't count on it.

2. Interchanging the position of the external resistors may improve the CMR; but it's just as likely that it will get worse.

3. Adding a trimpot in series with one of the external resistors can trim CMR to an acceptable value, but to restore gain accuracy, a second trimpot must also be added in series with the resistor network.

4. An INA117 or INA148 can achieve an input range of ±200V when it is operated on ±15VDC supplies. This is the easy way out of a CMV limitation predicament. Of course, adding external resistors to an INA117 or INA148 will cause problems, too, but with such a high CMV range they wouldn't be necessary.

5. Of course, there is still the option of building your own difference amplifier with a precision op amp and matching precision resistors. Remember that to achieve CMR of higher than 80dB, you must use resistors of better than 0.01%. And don't forget bias current errors.

Adding External Resistors—Sometimes?

After warning you about the pitfalls of adding external resistors to the inputs of a difference amplifier, an exception to the rule must be pointed out-- there is one case where adding external resistance is actually beneficial and recommended! In this case, we are only adding resistance to compensate for an unavoidable external resistance that is mandated by its application– the current shunt resistor.