

A PRACTICAL REVIEW OF Common Mode and Instrumentation Amplifiers

Addressed in
this tutorial are common mode voltage range
and common mode rejection as they pertain to instrumentation
amplifiers, their architectures, and their applications.

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Instrumentation amplifiers (in-amps) amplify the difference between two signals. These differential signals typically emanate from sensors such as resistive bridges or thermocouples. Figure 1 shows a typical in-amp application where the differential voltage from a resistive bridge is amplified by the AD620, a low-power, low-cost, integrated in-amp. In thermocouple and bridge applications, the differential voltage is generally fairly small (a few millivolts to tens of millivolts). However, the two voltages from the bridge are equal to about 2.5 V when each is referred to ground. This voltage, which is *common* to both inputs, is called the common mode voltage of the dif-

ferential signal. This voltage contains no useful information about the measurement. So ideally, the in-amp should amplify only the *difference* between the signals at its two inputs. Any common mode component should be ignored by the in-amp. Indeed, removing the common mode component is often the sole reason for using an in-amp. In practice, common mode signals will never be completely rejected by the in-amp; some remnant of the signal always appears at the output.

The specification of common mode rejection ratio (CMRR) is a measure of the extent to which common mode signals are rejected by an amplifier. CMRR is defined by:

$$\text{CMRR (dB)} = 20 \cdot \log \left(\frac{\text{GAIN} \cdot V_{\text{CM}}}{V_{\text{OUT}}} \right) \quad (1)$$

where:

gain = (differential) gain of amplifier
 V_{CM} = common mode voltage present at the input
 V_{OUT} = output voltage resulting from the presence of common mode voltage at the input

We can rewrite this equation to allow calculation of the output voltage that results from a particular common mode voltage:

$$V_{\text{OUT}} = \frac{\text{GAIN} \cdot V_{\text{CM}}}{\log^{-1} \left(\frac{\text{CMRR}}{20} \right)} \quad (2)$$

Now let's insert some numbers into this equation. The CMRR of an integrated in-amp such as the AD620 is 100 dB for a programmed gain of 10. In Figure 1, the common mode voltage is 2.5 V. This results in a voltage at the output of the in-amp of 250 μV . To put this into context, we should note that the output voltage that results from the combination of the input and output offset errors of the AD620 is ~ 1.5 mV. This suggests that as an error source, CMRR is less important than offset voltage. Up to now,

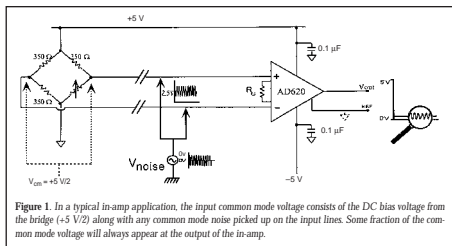


Figure 1. In a typical in-amp application, the input common mode voltage consists of the DC bias voltage from the bridge (+5 V/2) along with any common mode noise picked up on the input lines. Some fraction of the common mode voltage will always appear at the output of the in-amp.

however, we have been talking only about common mode rejection of DC signals.

AC and DC Common Mode Rejection

As shown in Figure 1, common mode signals can be steady-state DC voltages (such as the 2.5 V from the bridge) or they can be AC signals such as external interference. In industrial applications, the most common cause of external interference is pickup from 50/60 Hz mains sources (e.g., lights, motors, or any equipment running on the mains). In differential measurement applications, interference tends to be induced equally onto both in-amp inputs. The interfering signal therefore appears as a common mode signal to the in-amp. This signal will be superimposed on the DC common mode input voltage (from the bridge). At the output of the in-amp, we will see an attenuated version of the overall input common mode signal.

While a DC offset can be easily removed by trimming or calibration, AC errors appearing at the output are much more troublesome. If, for example, the input circuit picks up 50 Hz or 60 Hz interference from the mains, the AC voltage that appears at the output will reduce the resolution of the application. Filtering out this interference can be expensive and is feasible only in very slow applications. Obviously, high common mode rejection over frequency will help minimize the effects of external common mode interference.

We can conclude that specifying CMRR over frequency is in practice more important than specifying it at DC. Data sheets for integrated in-amps should ideally list the CMRR at 50/60 Hz on the specifications page and include a plot of CMRR vs. frequency in the plots section of the data sheet.

Figure 2 shows the change in CMRR over frequency for the AD623, a low-cost integrated in-amp. The CMRR remains flat up to 100 Hz and then begins to decrease. In this example, interference from 50/60 Hz mains interference will be well suppressed by the AD623. However, we also need to be conscious of interference caused by harmonics of the mains frequency. In industrial environments, harmonics of the mains frequency can be significant up to the seventh harmonic (350/420 Hz). In the case of the AD623, we see that the CMRR has dropped to -90 dB at a gain of 10 at these frequencies. This results in a common mode gain of -70 dB, still enough to suppress most common mode interference.

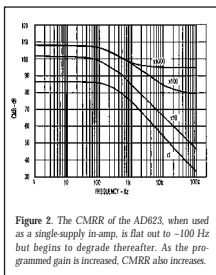


Figure 2. The CMRR of the AD623, when used as a single-supply in-amp, is flat out to ~100 Hz but begins to degrade thereafter. As the programmed gain is increased, CMRR also increases.

We will now look at different in-amp architectures. It will become clear that choice of architecture and the precision of passive components affects both the AC and DC CMRR.

The 2 Op-Amp In-Amp

Figure 3 is a circuit diagram for a basic 2 op-amp in-amp. The differential gain is given by [1]:

$$V_{OUT} = (V_{IN+} - V_{IN-}) \left(1 + \frac{R1}{R2} \right) \quad (3)$$

where:

$$R1 = R4 \text{ and } R2 = R3$$

With R1 equal to 10 kΩ, and R2 equal to 1 kΩ, the differential gain is equal to 11. We can see from Equation 3 that a programmed gain of 1 is fundamentally not achievable.

Common Mode Gain. The output volt-

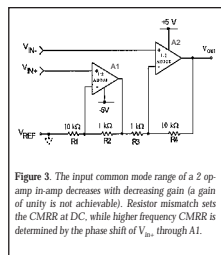


Figure 3. The input common mode range of a 2 op-amp in-amp decreases with decreasing gain (a gain of unity is not achievable). Resistor mismatch sets the CMRR at DC, while higher frequency CMRR is determined by the phase shift of V_{in} through A1.

age that results from the presence of DC common mode voltage is given by:

$$V_{OUT} = V_{cm} \left(1 - \frac{R2R4}{R1R3} \right) \quad (4)$$

Using Equation 1, the formula for the CMRR of the circuit comes out to be:

$$CMRR = 20 \cdot \log \left(\frac{GAIN}{1 - \frac{R2R4}{R1R3}} \right) \quad (5)$$

Because the resistor ratio in the denominator is always close to 1, regardless of the in-amp's gain, we can conclude that the CMRR or a 2 op-amp in-amp increases with gain.

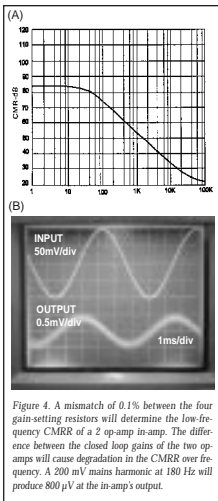
It is common to specify the accuracy of resistor networks in terms of resistor-to-resistor percentage mismatch. We can rewrite Equation 5 to reflect this:

$$CMRR = 20 \cdot \log \left(\frac{GAIN \cdot 100}{\%MISMATCH} \right) \quad (6)$$

Any mismatch between the four gain setting resistors will have a direct impact on the CMRR. Precision resistor networks are typically trimmed for maximum accuracy at ambient temperature. Any mismatch in the temperature drift of the resistors will further degrade the CMRR. Clearly, the key to high common mode rejection is a network of resistors that are well matched from the perspective of both resistive ratio and relative drift. It should be noted that the absolute values of the resistors and their absolute drifts are of no consequence. Matching is the key.

Integrated instrumentation amplifiers are particularly well suited to meeting the combined needs of ratio matching and temperature tracking of the gain-setting resistors. While thin film resistors fabricated on silicon have an initial tolerance of up to $\pm 20\%$, laser trimming during production allows the ratio error between the resistors to be cost effectively reduced to 0.01% (100 ppm). Furthermore, the tracking between the temperature coefficients of the thin film resistors is inherently low and is typically $< 3 \text{ ppm}/^\circ\text{C}$ (0.0003%/°C).

The plot in Figure 4A exhibits the practical results, at ambient temperature, of resistor mismatch. The CMRR of the circuit in Figure 3 (gain = 11) was measured using four resistors that had a mismatch of almost exactly 0.1% ($R1 = 9999.5 \Omega$, $R2 = 999.76 \Omega$, $R3 = 1000.2 \Omega$, $R4 = 9997.7 \Omega$). As expected, the



CMRR at DC was measured at -84 dB (the theoretical value is 85 dB). As the frequency increases, however, the CMRR quickly degrades. Figure 4B is an oscilloscope photo of the output voltage that results from mains interference. A 200 mV peak-to-peak harmonic of the mains frequency at 180 Hz results in an output voltage of ~ 800 μ V. To put this into context, a 12-bit data acquisition system with an input range of 0 V to 2.5 V has a least significant bit (LSB) weighting of 610 μ V.

The degradation in CMRR over frequency is caused by the phase shift of the V_{IN+} signal through A1. This sets up a vector error between V_{IN+} and the output of A1. For infinite CMRR, these signals would have to have identical phase and amplitude from a common mode perspective. This would only be possible if there were no delay through A1. Choosing a well-matched high-speed dual op-amp will extend the frequency over which the CMRR stays flat, but high-speed op-amps also have a tendency to pick up high-frequency external interference. An alternative solution to this problem is to use an AC trim capacitor between

the inverting input of A1 and ground [1]. This trim, however, would need to be done by hand for every in-amp produced.

So the plot of CMRR vs. frequency in Figure 4 is influenced by two distinct parameters. The CMRR at low frequencies is related directly to the mismatch between the gain-setting resistors, while the degradation in CMRR at higher frequencies is caused by the different closed loop gains of the op-amps.

Common Mode Range. The input common mode range of the 2 op-amp in-amp is affected by the programmed gain. In Figure 3, we can see that A1 is operating at a closed loop gain of 1.1. Any common mode voltage present at the input will be amplified by this amount by A1 (i.e., $1.1 \times$ the common mode voltage appears at the output of A1).

Now consider a case where the in-amp has a programmed gain of 1.1 ($R_1 = 1$ k Ω , $R_2 = 10$ k Ω , $R_3 = 10$ k Ω , $R_4 = 1$ k Ω). Now A1 is operating at a closed loop gain of 11. Because the common mode voltage is being amplified by A1, the input common mode range is severely restricted by the output swing of A1. The problem is especially acute in applications where low voltage supplies are mandatory. The use of rail-to-rail amplifiers will improve matters somewhat by adding some more headroom.

The 3 Op-Amp In-Amp

The 3 op-amp in-amp architecture (see Figure 5) is a popular choice for both discrete and integrated in-amps. The overall gain transfer function is quite complicated, but if $R_1 = R_2 = R_3 = R_4$, the transfer function simplifies to [1]:

$$V_{OUT} = (V_{IN+} - V_{IN-}) \left(1 + \frac{R_5 + R_6}{R_C} \right) \quad (7)$$

R_5 and R_6 are typically set to the same value, usually somewhere between 10 k Ω and 50 k Ω . The circuit's overall gain can be adjusted from unity to an arbitrarily high value simply by changing the value of R_C .

Common Mode Gain. As we would expect, the common mode gain of the in-amp should ideally be equal to zero. To work out the common mode gain, let's imagine that there is only a common mode voltage of V_{CM} present at the inputs (i.e., $V_{IN+} = V_{IN-} = V_{CM}$). As there is no voltage drop across R_C , the voltage on the outputs of each of the amplifiers, A1 and A2, is also equal to V_{CM} . So to a first approximation (i.e., assuming A1 and A2 are ideally

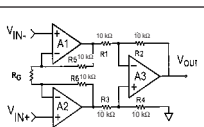


Figure 5. In a 3 op-amp in-amp, a mismatch of 0.1% among resistors R_1 , R_2 , R_3 , and R_4 will result in a worst case CMRR (at a gain of 1) of 60 dB. Drift mismatch between resistors will further reduce the CMRR. Monolithic in-amps offer greater accuracy at lower cost.

matched) the common mode gain of the first stage is equal to unity and is independent of the programmed gain.

Assuming that op-amp A3 is ideal, the common mode gain of the second stage is given by:

$$\frac{V_{OUT}}{V_{CM}} = \left(\frac{R_2 + R_1}{R_1} \cdot \frac{R_4}{R_3 + R_4} - \frac{R_2}{R_1} \right) \quad (8)$$

Plugging this into Equation 1, the equation for the common mode rejection ratio becomes:

$$\text{CMRR} = 20 \cdot \log \left(\frac{\text{GAIN}}{\left(\frac{R_2 + R_1}{R_1} \cdot \frac{R_4}{R_3 + R_4} - \frac{R_2}{R_1} \right)} \right) \quad (9)$$

The denominator of this equation is more complicated than it is for the 2 op-amp in-amp. Just as in Equation 6, however, the denominator can be replaced by the percentage mismatch between the resistors:

$$\text{CMRR} = 20 \cdot \log \left(\frac{\text{GAIN} \cdot 100}{\% \text{MISMATCH}} \right) \quad (10)$$

Now, if all four resistors in Equation 9 are equal (or even if $R_1 = R_3$ and $R_2 = R_4$), the denominator will reduce to zero. But any mismatch between the four resistors will cause a portion of the common mode voltage to appear at the output. Similar to the case of the 2 op-amp in-amp, any mismatch between the temperature drift of the resistors will further degrade the CMRR as the temperature changes.

AC CMRR. If A1 and A2 are well matched (i.e., have similar closed loop bandwidths), the CMRR will not tend to degrade so quickly as it does with the 2 op-amp in-amp. Again referring to Figures 2 and 4, we see that the CMRR of the 3 op-amp in-amp remains relatively flat out to

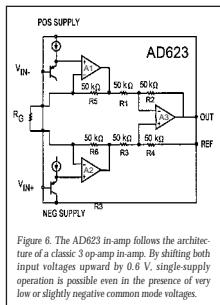


Figure 6. The AD623 in-amp follows the architecture of a classic 3 op-amp in-amp. By shifting both input voltages upward by 0.6 V, single-supply operation is possible even in the presence of very low or slightly negative common mode voltages.

100 Hz while the CMRR of a 2 op-amp in-amp begins to degrade at -10 Hz.

Common Mode Range. As we have previously noted, the common mode gain of the first stage of a 3 op-amp in-amp is unity, with the result that the common mode voltage appears at the output of A1 and A2 in Figure 5. The differential input voltage, V_{DIFF} , however, appears across the gain resistor. The resulting current that must flow through R5 and R6 means that the voltage on A1 will rise above V_{cm} and the voltage on A2 will drop below V_{cm} as the differential input voltage increases. Therefore, as the gain and/or input signal increases, so does this "spreading" of the voltages on A1 and A2, ultimately to be limited by the supply rails.

We can conclude that the achievable ranges on the common mode voltage, the differential input voltage, and the gain are interrelated. For example, increasing the gain reduces both common mode range and input voltage range. By the same token, increasing the common mode voltage tends to limit the differential input range and the maximum achievable gain. If the output swings of the input stage op-amps are known, the relationship governing input range, common mode range, and gain can be well defined for a particular 3 op-amp in-amp [2].

As the industry moves to lower supply voltages, this issue becomes more critical with less and less headroom being available. As in the case of the 2 op-amp in-amp, the use of rail-to-rail op-amps maximizes available headroom. A rail-to-rail output stage (A3) is of little use, though, if the output voltages of the input stage, A1 and A2, are being clipped because of excessive input voltage, common mode voltage, or gain.

Single-Supply In-Amp for Low Common Mode Applications

The AD623, a low-cost single-supply rail-to-rail in-amp (see Figure 6), follows the classic 3 op-amp in-amp architecture. But before being applied to the input stage op-amps, both the inverting and noninverting input voltages are shifted upward by 0.6 V (i.e., a diode drop) as they each pass through a pnp transistor.

To understand the consequences of this level shifting, we should consider the conditions under which the in-amp is usually operated. In Figure 7, it is shown amplifying the signal from a J-type thermocouple. The in-amp, along with the A/D converter into which it feeds, is powered by a single supply of $+5$ V. The temperature to be measured ranges from -200°C to 200°C , which corresponds to a thermocouple voltage of -7.890 mV to $+10.777$ mV.

As is normal practice, one side of the thermocouple is grounded to allow the necessary bias currents to flow into the in-amp. As a result, the common mode voltage, which is halfway between the inverting and noninverting input voltages, is very close to ground. Indeed, as the voltage from the thermocouple becomes negative, the effective common mode voltage also goes negative.

In a conventional 3 op-amp in-amp, the voltage-spreading effect of the input stage would cause the output voltage on one of the input op-amps to run into the ground rail as soon as the thermocouple voltage gets above 0 V. The level shifting architecture in Figure 6 gets around this problem by effectively adding 0.6 V to the common mode voltage. This creates more headroom to ground and allows the output voltages of A1 and A2, which are rail-to-rail, to stay in a

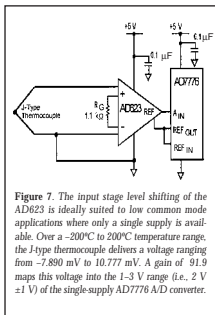


Figure 7. The input stage level shifting of the AD623 is ideally suited to low common mode applications where only a single supply is available. Over a -200°C to 200°C temperature range, the J-type thermocouple delivers a voltage ranging from -7.890 mV to 10.777 mV. A gain of 91.9 maps this voltage into the 1–3 V range (i.e., 2 V \pm 1 V) of the single-supply AD7776 A/D converter.

linear region, even when the input voltage and the common mode voltage go below ground. The input voltage can go negative by as much as 150 mV, depending on the programmed gain and the common mode voltage [2].

In this example, the programmed gain on the in-amp is 91.9 ($R_G = 1.1$ k Ω). The voltage on the in-amp's REF pin has been set to 2 V. So as the thermocouple voltage varies from -7.890 mV to $+10.777$ mV, the in-amp's output voltage ranges from 1.274 V to 2.990 V (relative to ground). This voltage swing fits comfortably into the input range of the A/D converter, which is 2 V \pm 1 V. ■

References

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2. AD623 Single Supply, Rail-to-Rail, Low Cost Instrumentation Amplifier. Oct. 1997. Analog Devices: 15. Available at <http://www.analog.com/AD623/>

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