Reducing Offset in Wideband Amplifiers

High-precision op-amps, such as auto-zero op-amps (AZA), possess an internal mechanism that continuously reduces their input offset to a few micro-volts (μV) or less. This feature makes them well suited in high-gain signal conditioners of sensor applications as output offset voltage and drift are drastically reduced.

Other applications of AZAs are the use as servo amplifiers in the offset cancellation schemes of wideband amplifiers (WBA) with high input offsets.

This application note discusses the design of such a circuit, also known as an auto-zeroed wideband amplifier (Figure 1). It explains the offset cancellation principle and answers frequently asked questions related to the design.
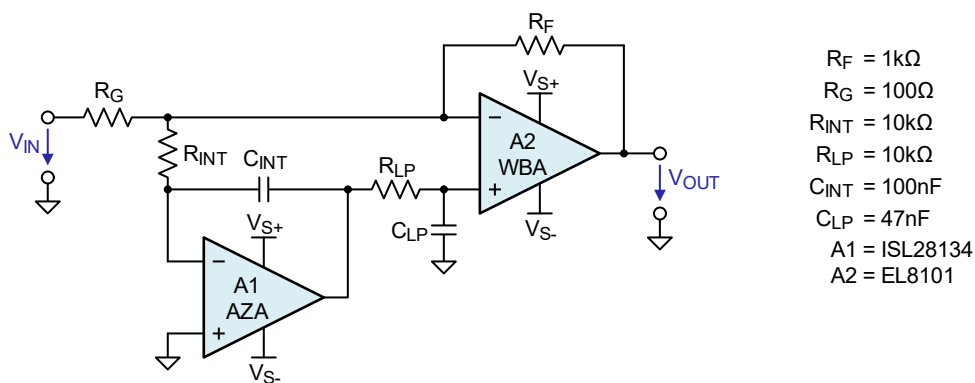

R_F = 1kΩ
R_G = 100Ω
R_{INT} = 10kΩ
R_{LP} = 10kΩ
C_{INT} = 100nF
C_{LP} = 47nF
A1 = ISL28134
A2 = EL8101

Figure 1. DC-Coupled Auto-Zeroed Wideband Amplifier

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1. Auto-Zeroed Wideband Amplifier

Figure 2 shows the schematic of an auto-zeroed wideband amplifier operating in the inverting amplifier configuration with the closed-loop signal gain: $V_{OUT}/V_{IN} = -R_F/R_G$.

A high-precision auto-zero op-amp (A1), configured as an inverting integrator, measures the input offset of the wideband amplifier (A2) at the inverting input. The integrator converts this offset into an output voltage of the same magnitude but opposite polarity and applies it to the non-inverting input of A2.

The passive low-pass filter at the integrator output filters the internal noise of A1 and prevents it from being amplified by A2.

This approach not only eliminates the input offset voltage of the wideband amplifier, but also its offset drift due to the continuous offset correction of the auto-zero amplifier.
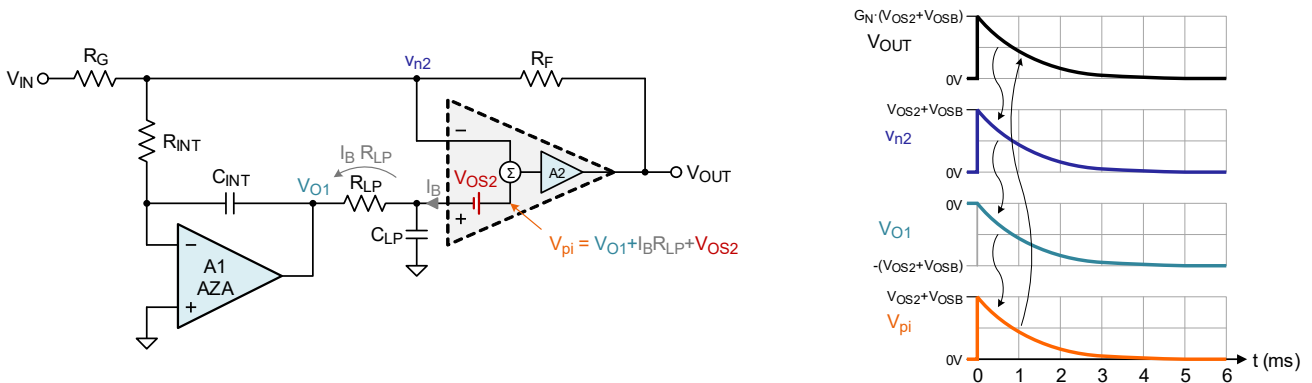

Figure 2. Operational Principle of the Wideband Amplifier and its Offset Cancellation Scheme

1.1 Offset Cancellation in Detail

Consulting Figure 2, the voltage at the internal summing point of the wideband amplifier, V_{pi}, is the sum of the input offset, the offset due to bias current, and the integrator output voltage:

(EQ. 1) $\quad V_{pi} = V_{OS2} + I_B \times R_{LP} + V_{O1}$

It is this voltage that is amplified by the amplifier noise gain, $G_n = 1 + R_F/R_G$ to make $V_{OUT} = V_{pi} \times G_n$. Assuming that at t = 0, the integrator output, $V_{O1} = 0V$, the wideband amplifier output voltage is $V_{OUT} = G_n (V_{OS2} + I_B R_{LP})$. Negative feedback attenuates V_{OUT} by the feedback factor $\beta = R_G/(R_F + R_G)$ to make the voltage at the inverting input $v_{n2} = V_{OS2} + I_B R_{LP}$. This voltage forces V_{O1} to move towards negative voltage potential; therefore, reducing V_{pi}. Completing the loop, a decrease in V_{pi} causes V_{OUT} and v_{n2} to decrease, which slows the discharge rate of the integrator until V_{pi}, V_{OUT}, and v_{n2}, are near 0V. **Note:** In this steady state, the integrator output voltage has the same magnitude but opposite polarity of the sum of the input offset and the offset due to bias current:

(EQ. 2) $\quad V_{O1} = -(V_{OS2} + I_B \times R_{LP})$

2. Frequently Asked Questions

2.1 Why select the inverting amplifier configuration?

Many wideband amplifiers possess low common-mode rejection (70dB to 80dB) and a limited input common-mode voltage range. Because the output error due to common-mode input change is $\varepsilon_{cm(out)} = A_{CL} \times \Delta V_{CM}/CMRR$, a non-inverting amplifier produces a common-mode error at the output because the input voltage is also the change in common-mode voltage, $\Delta V_{CM} = V_{IN}$ (Figure 3).

However, in the inverting configuration, the non-inverting input is connected to a reference potential that ties the inverting input to the same fixed potential because of negative feedback. Therefore, $\Delta V_{CM} = 0V$, which eliminates the common-mode output error.

Also, in the non-inverting configuration, the input voltage must not exceed the upper and lower limits of the input common-mode range, which often is a few volts below and above the upper and lower supply rails, respectively. In the inverting amplifier configuration, these limits do not apply, allowing the input voltage to exceed the supply rails.
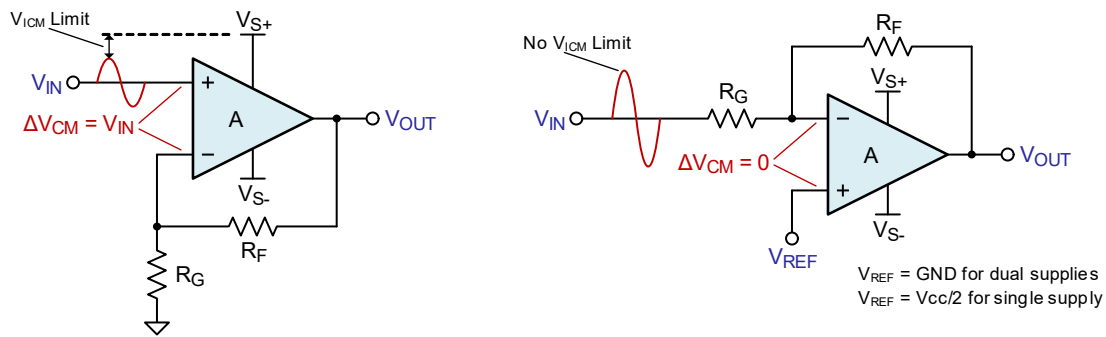

Figure 3. Benefits of Inverting over Non-Inverting Amplifier

2.2   Why sense V_{OS} at the inverting input?

Because the inverting input allows direct access to the op-amp internal offset voltage without any interference from the input signal, which becomes clear when consulting the inverting amplifier in Figure 4. Here, v_{n2} is the sum of a fraction of the input voltage and a fraction of the output voltage, which is expressed through Equation 3:

(EQ. 3) $v_{n2} = V_{IN}\dfrac{R_F}{R_F + R_G} + V_{OUT}\dfrac{R_G}{R_F + R_G}$

Because the output voltage of the inverting amplifier is:

(EQ. 4) $V_{OUT} = V_{OS}\left(1 + \dfrac{R_F}{R_G}\right) - V_{IN}\dfrac{R_F}{R_G}$

replacing the V_{OUT} term in Equation 3 with Equation 4 eliminates V_{IN} and results in Equation 5:
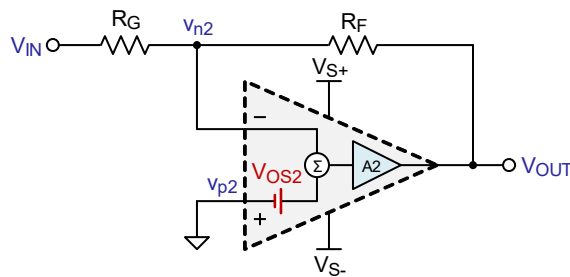
(EQ. 5) $v_n = V_{OS}$

Figure 4. V_{OS} at v_{n2} is independent of V_{IN}

2.3 How much output offset voltage remains after cancellation?

You can determine the remaining output offset by drawing the auto-zeroed wideband amp in the DC condition, which removes the capacitors and makes A1 operate at its open-loop gain.
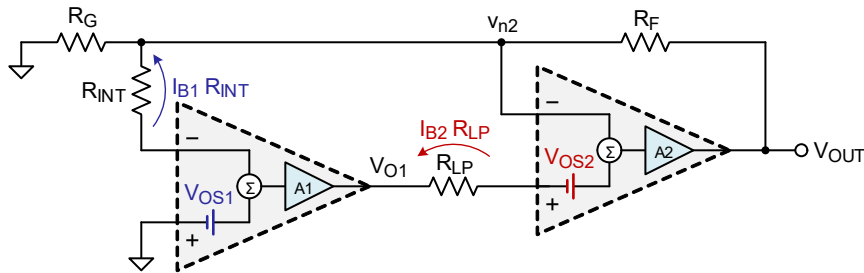

Figure 5. Auto-Zeroed Wideband Amplifier at DC

Establishing the equations for both op-amp outputs gives Equation 6 and Equation 7.

(EQ. 6) $\quad V_{OUT} = A_2(V_{OS2} + I_B R_{LP} + V_{O1} - v_{n2})$

(EQ. 7) $\quad V_{O1} = A_1(V_{OS1} - I_{B1} R_{INT} - v_{n2})$

Substituting the v_{n2} term in both equations with $v_{n2} = V_{OUT} R_G/(R_F + R_G)$ and replacing the V_{O1} term in Equation 6 with Equation 7 provides the remaining output offset voltage:

(EQ. 8) $\quad V_{OUT} = \left(1 + \dfrac{R_F}{R_G}\right) \times \left(\dfrac{V_{OS2} + I_{B2} R_{LP}}{A_1} + V_{OS1} - I_{B1} R_{INT}\right)$

Equation 8 shows that the high input offset of the wideband amplifier, A_2, in addition to the offset voltage caused by its bias current, I_{B2}, are attenuated by the high DC open-loop gain of the precision amplifier, A_1. Therefore, the effective offset components being amplified by the noise gain $1 + R_F/R_G$ are V_{OS1} and the offset caused by the bias current, I_{B1} of A1. The following numerical example emphasizes the difference in magnitude between the individual offset components.

Note: As offset voltages and bias currents can have positive or negative values, the highest offset occurs when adding all offset components.

Example:

The circuit in Figure 5 has the following resistors values and op-amp parameters:

R_F = 10kΩ, R_G = 100Ω, R_{INT} = 100kΩ, R_{LP} = 100kΩ, V_{OS2} = 6mV, I_{B2} = 2µA, V_{OS1} = 4µV, I_{B1} = 550pA, A_1 = 174dB or 501.2×10^{6}V/V

$$V_{OUT} = 11 \times \left(\frac{6mV + 2\mu A \cdot 100k\Omega}{501.2 \cdot 10^6 \text{ V/V}} + 4\mu V + 550pA \cdot 100k\Omega\right)$$

$$\Downarrow \qquad\qquad \Downarrow \qquad\qquad \Downarrow$$

$$V_{OUT} = 11 \times (\quad 411pV \quad + \quad 4\mu V \quad + \quad 55\mu V \quad) = 649\mu V$$

Although the 649µV represents a remarkable offset reduction from the 66mV without offset cancellation, it is still possible to reduce the output offset by another 0.5mV when lowering the value of R_{INT} to 10kΩ.

$$V_{OUT} = 11 \times (\quad 40pV \quad + \quad 4\mu V \quad + \quad 5.5\mu V \quad) = 105\mu V$$

This example shows that R_{INT} becomes the major offset contributor for values larger than 10kΩ. On the other hand, R_{INT} should be large enough to prevent the loading of the signal path:

(EQ. 9) $\quad R_{INT} \geq 100 R_G$

Fortunately, wideband amplifiers favor low-impedance feedback and gain resistors, allowing R_{INT} to have a relatively low value.

2.4 Why is a low-pass filter required?

What is not immediately obvious from the circuit schematic in Figure 2 is the potential noise contribution of the integrator to the non-inverting input of the wideband amplifier, A_2.

While the signal gain of the integrator, its inverting gain, results in a slope of -20dB/dec, the noise gain of the integrator, its non-inverting gain, becomes one (0dB) at the zero-frequency $f_{INT} = 1/(2\pi R_{INT} C_{INT})$.

This noise gain of 1V/V continues across the remaining bandwidth of the auto-zero amplifier A_1 (Figure 6). It is this bandwidth across which the internal noise of the op-amp can integrate to a substantial RMS noise voltage, which would be amplified by the noise gain of the wideband amplifier.

Therefore, to minimize integrator noise, a low-pass filter (LPF) is added to its output that continues reducing the noise gain across the bandwidth of A_1.

However, the low-pass filter adds a pole into the offset cancellation path, leading to a gain peak in the overall noise gain. To minimize this peak, make the filter cutoff frequency twice that of the integrator zero:

(EQ. 10) $\quad f_{LP} = 2f_{INT}$

To measure the various noise gains, the integrator must operate in the non-inverting configuration (Figure 6). Therefore, V_{IN} is now applied to the non-inverting input of A_1 so that V_{O1}/V_{IN} represents the integrator noise gain and v_{p2}/V_{IN}, the noise gain with a low-pass filter.
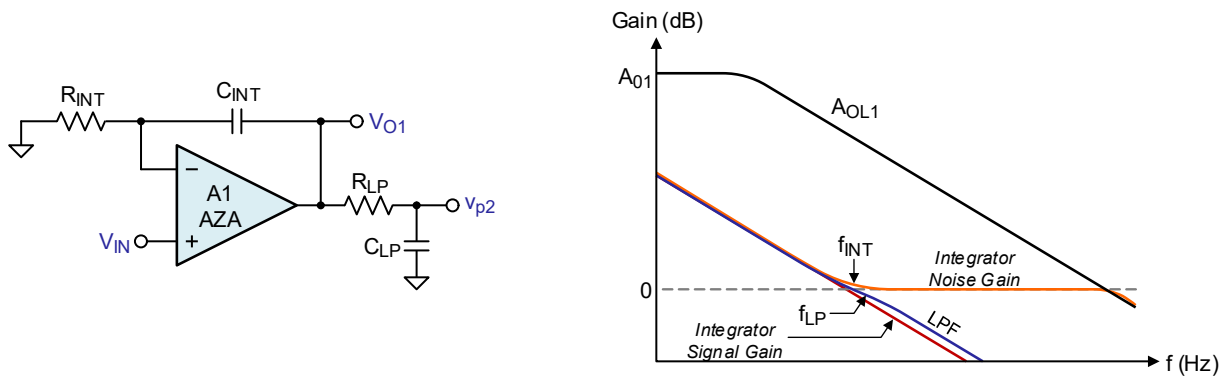

Figure 6. Measuring Integrator Noise Gain with and without a Low-Pass Filter (LPF)

2.5   What is the optimum integrator time constant?

The integrator time constant of 1ms is a compromise between offset cancellation time and RMS output noise. The simulation results in Figure 7 show that shorter time constants increase the RMS output noise voltage, while longer time constants increase the offset cancellation time.

Note: The black curve is the RMS output noise of the wideband amplifier without an offset cancellation network.
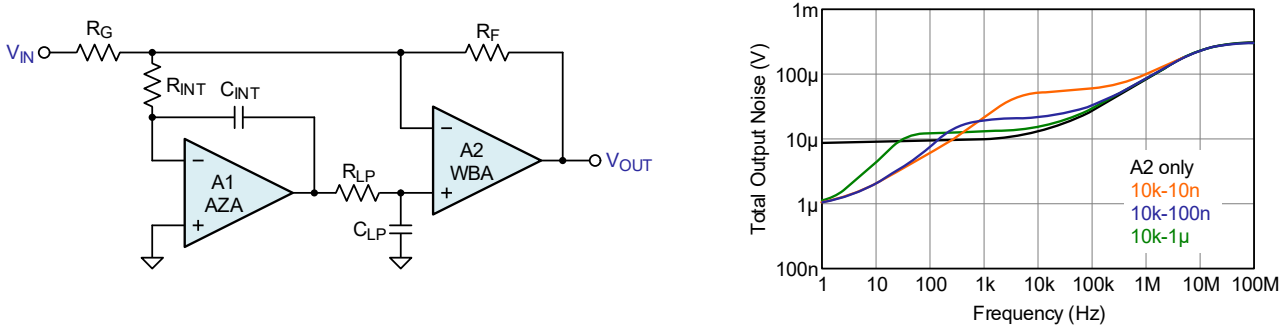

Figure 7. Simulating Total Circuit RMS Output Noise

2.6   How do I calculate the component values?

1. The application section in the datasheet of a wideband amplifier usually shows the preferred feedback resistor value to be in the lower kΩ range. For the EL8101, R_F = 1kΩ.

2. For a required closed-loop gain, in this case, ACL = 10V/V, calculate the gain resistor value with $R_G = R_F/A_{CL}$, which makes R_G = 100Ω.

3. Apply Equation 9, $R_{INT} \geq 100R_G$, to make R_{INT} = 10kΩ.

4. For a 1ms time constant, calculate the integrator capacitor with $C_{INT} = T/R_{INT}$, which gives C_{INT} = 100nF.

5. For the low-pass filter, making $R_{LP} = R_{INT}$ = 10kΩ, and applying Equation 10, $f_{LP} = 2f_{INT}$, makes C_{LP} = 47nF.

2.7   Can a fraction of the input signal reach the non-inverting input of the WBA?

Yes, but this fraction is tiny. Figure 8 expresses this fraction as an input signal suppression ratio, v_{p2}/V_{IN}. The minimum suppression is at DC. There, it is the difference (dB) between the circuit closed-loop gain, A_{CL}, and the wideband amplifier open-loop gain, A_{OL2}: $v_{p2}/V_{IN} = A_{CL} - A_{OL2}$. It continues at this level up to the geometric mean of the integrator and low-pass frequencies $\sqrt{f_{INT}f_{LP}}$. Beyond this frequency, the suppression increases.

However, the exact characteristic can differ between amplifier solutions primarily because of the varying output impedance characteristics of auto-zero amplifiers.
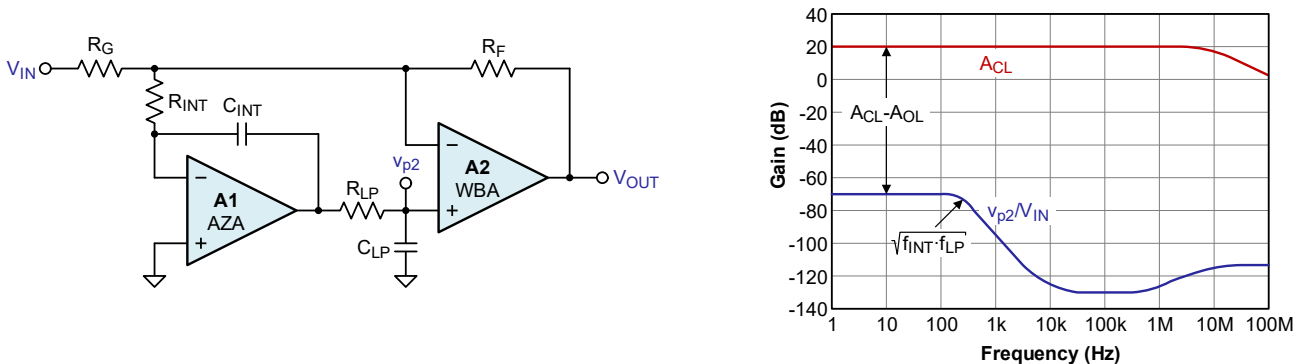

Figure 8. Input Signal Suppression at the Non-Inverting Input of the WBA

3. Conclusion

This version of an auto-zeroed wideband amplifier senses the input offset at the inverting input of the WBA. Unlike the many published circuits that sense the amplified offset at the output of the WBA, the approach presented here does not require high-pass filtering of the input signal to separate it from the offset eliminating the servo loop; therefore, allowing for the signal path to be DC-coupled.

4. Revision History

Revision	Date	Description
1.00	Jul 5, 2022	Initial release.

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