Abstract

The inputs of an operational amplifier (op-amp) must be DC-biased to ensure proper device operation. A basic requirement that many textbooks neglect to discuss in detail. Consequently, engineers new to op-amps might overlook this important requirement, which can lead to malfunctioning circuits.

This application note tries to rectify this shortcoming by explaining the need for input biasing and suggesting practical solutions that ensure proper circuit operation.

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1. The Need for Input Biasing

Figure 1 shows the differential input stage of an op-amp. The base terminals of transistors Q1 and Q2 form the non-inverting and inverting op-amp inputs, IN+ and IN-, respectively. For the op-amp to operate correctly, these inputs must be DC biased. That is, the DC bias currents (IB+ and IB-), must be able to flow into or out of the input terminals. The direction of the bias currents depends on the type of transistors. For NPN transistors and N-channel enhancement MOSFETs, the currents flow into the inputs, for PNP s and P-channel enhancement MOSFETs, the currents flow out of the inputs.

DC biasing is achieved by connecting the inputs via resistors to a reference potential, Vmid, which is the mid potential of the positive and negative supply voltages (VS+ and VS-) (Figure 1). The biasing resistors (RB+ and RB-), represent equivalent resistances, because they can consist of one or more resistors as depicted in the DC-coupled amplifier circuits of Figure 2.

In the DC-coupled buffer of Figure 2, IB+ flows through the signal source and its low source impedance into the noninverting input, while IB- flows from the output of the op-amp into the inverting input.

In the DC-coupled amplifier, IB- flows through RF and RG; therefore changing the DC voltage at the inverting input by IB- • (RF||RG). This introduces a DC offset voltage between the inputs. To minimize this offset, the DC potential at the non-inverting input must be adjusted by about the same amount. This is achieved by inserting a resistor (RIN), in series whose value matches the parallel combination of RF and RG, RIN = RF||RG.

2. AC-Coupled Amplifiers with Dual Supplies

While DC-coupled op-amp circuits receive their biasing through the signal source impedance, AC-coupled circuits have this bias path blocked by the input coupling capacitor (CIN). Figure 3 shows an AC-coupled amplifier without a path for the DC bias current to flow. In this case, IB+ charges the coupling capacitor until the common-mode voltage rating of the input circuit is exceeded, or its output is driven into saturation. Depending on the polarity of the bias current, the capacitor charges towards the positive or negative supply rail, with the resulting bias voltage being amplified by the closed-loop DC gain of the amplifier, 1 + RF/RG.
For a small bias current, such as 1pA, this process can take hours. Therefore, a casual lab test with an AC-coupled scope is more than likely to miss this problem, and the circuit does not fail until much later. Obviously, this problem must be avoided.

Figure 4 shows a simple solution by connecting the noninverting input through $R_{IN}$ to ground. This forms a new DC path for the bias current. Like in the DC-coupled case before, minimizing the offset due to bias currents requires $R_{IN} = R_F \parallel R_G$.

Note: AC-coupling forms a high-pass filter with a cutoff at $f_C = 1/(2\pi C_{IN} R_{IN})$, which sets the minimum input bandwidth of the amplifier. With the circuit gain being defined by the application, $R_C$ and $R_S$ can be determined. Their parallel value defines $R_{IN}$, which then allows the calculation of the input capacitor: $C_{IN} = 1/(2\pi f_C R_{IN})$.

Similar input biasing methods must be applied to the differential input stages of three-amp Instrumentation Amplifiers (INAs). Figure 5 shows INA circuits that are AC-coupled using either two capacitors or a transformer, without providing a DC bias path.

Correct biasing solutions for these circuits are shown in Figure 6, where a high-value resistor ($R_{IN}$) is added between each input and ground.

There is a small offset-voltage error due to mismatches between the input resistors and the input bias currents. To minimize this error, a third resistor, about 1/10th their value (yet still large compared to the differential source resistance), can be connected between the two in-amp inputs, therefore bridging both resistors.
3. AC-Coupled Amplifiers with Single Supply

The DC biasing of AC-coupled single-supply amplifiers also requires an input resistor connecting the noninverting input to the reference potential, $V_{mid}$. In single supply circuits however, $V_{mid}$ is derived from $V_S$ as $V_S/2$ to ensure a symmetrical, maximum dynamic range for both, input and output signals. This can make $V_S/2$ susceptible to supply noise.

The ideal design approach is to use an integrated voltage-reference chip (VREF) with high power-supply rejection ratio (PSRR). A VREF also provides low output impedance, allowing to provide $V_S/2$ potential to multiple reference locations within a larger circuit design (Figure 7).

For biasing a single-supply op-amp, a more convenient low-cost method exists in form of a voltage divider, whose output must be buffered with a large capacitor ($C_B$), to make $V_S/2$ less sensitive to supply noise, thus maintaining a high PSRR (Figure 8).

AC-coupled single-supply amplifiers also require the capacitive decoupling of the feedback path to ensure 0dB gain at DC. This prevents the $V_S/2$ potential at the noninverting input from being amplified by the passband gain, therefore saturating the output.

To calculate the component values, it is important to understand the interactions of the various time constants: $R_B$-$C_B$, $R_{IN}$-$C_{IN}$, and $R_G$-$C_G$. For clarity, their frequency responses are depicted in Figure 9, Figure 10, and Figure 11.

To the $V_S$ supply, the biasing voltage divider presents a low-pass filter whose frequency response is:

$$G_{SN} = \frac{\Delta V_{CC}}{\Delta V_{CC}} = \frac{1}{2 + \frac{1}{1 + j\omega C_B R_B/2}}, \text{ with a cutoff frequency of } f_{CB} = \frac{1}{2\pi C_B R_B/2}. \text{ Here, } G_{SN} \text{ denotes the supply-noise gain.}$$

$$G_{SN} = \frac{\Delta V_{S}/2}{\Delta V_S} = \frac{1}{2 + \frac{1}{1 + j\omega C_B R_B/2}} \text{ (Output Noise)}$$

Figure 9. Transfer Function and Frequency Response of the Supply-Noise Gain of the Voltage Divider
The $R_{IN}C_{IN}$ circuit has a high-pass response $G_{IN} = \frac{j\omega C_{IN}R_{IN}}{1+j\omega C_{IN}R_{IN}}$ with pole at $f_{CI} = \frac{1}{2\pi C_{IN}R_{IN}}$. This pole and the pole of the closed-loop gain, $f_{PG}$, determine the minimum input bandwidth of the amplifier circuit, $f_{CT}$.

![Figure 10. Transfer Function and Frequency Response of the AC-Coupled Input Stage](image)

The closed-loop gain of the op-amp is given with $A_{CL} = \frac{1+j\omega C_{G}(R_{G}+R_{F})}{1+j\omega C_{G}R_{G}}$. It has a pole and a zero-location due to the decoupling effect of $C_{G}$. The zero frequency, $f_{ZG} = \frac{1}{2\pi C_{G}(R_{G}+R_{F})}$, is the +3dB corner frequency above unity gain. This zero ensures that DC voltages at the noninverting input, such as $V_{S}/2$ and any offset errors, are amplified at a gain of 1V/V (0dB). The pole frequency, $f_{PG} = \frac{1}{2\pi C_{G}R_{G}}$, defines the -3dB cutoff of $A_{CL}$. This pole and the pole of the input stage, $f_{CI}$, determine the minimum input bandwidth of the amplifier circuit, $f_{CT}$.

![Figure 11. Transfer Function and Frequency Response of the Closed-Loop Gain of the Op-Amp, $A_{CL}$](image)

Depicting all discussed frequency responses, Figure 12 shows how $G_{IN}$ and $A_{CL}$ define the total gain ($G_T$), of the AC-coupled amplifier using $G_T = G_{IN} \cdot A_{CL}$, therefore moving the new -3dB cutoff, $f_{CT}$, to a higher frequency.

To ensure enough supply-noise suppression at this new cut-off frequency, practical experience suggests that the corner frequency of the voltage divider should be at least 1/10th of $f_{CT}$: $f_{CB} = 0.1 \cdot f_{CT}$. 

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**Operational Amplifiers**

**How to Bias Op-Amps Correctly**
3.1 Design Example

This section discusses the practical design of an actual amplifier circuit. As previously mentioned, with regard to component selection there are always compromises to be made between power consumption, resistor noise contribution, and supply noise rejection.

The design goal is to build an AC-coupled noninverting amplifier, operating from a single 5V supply, with an overall passband gain of 100V/V (40dB) and a minimum bandwidth of $f_{CT} = 16\text{Hz}$.

This requires the corner frequency of the voltage divider to be $f_{CB} \leq \frac{1}{10} f_{CT} = 1.6\text{Hz}$. To preserve the power in the voltage divider, a DC quiescent current of 50µA is allowed. This makes the value of $R_B$.

The buffer capacitor, $C_B$, can therefore be calculated with:

\[ C_B = \frac{1}{2\pi f_{CB}/2} = \frac{1}{2\pi \cdot 1.6\text{Hz} \cdot 25\text{k}\Omega} = 4\mu\text{F} \]

To ensure an earlier rather than later supply-noise roll-off, chose the next higher standard value with $C_B = 4.7\mu\text{F}$.

Aiming for equal corner frequencies for the input $R_{IN}C_{IN}$ stage and the closed-loop gain of the op-amp ($A_{CL}$), demands that $f_{CI} \sim f_{pG}$. These frequencies are derived from $f_{CT}$ using:

\[ f_{CI} = f_{pG} = \frac{f_{CT}}{\sqrt{1 + \sqrt{2}}} = \frac{16\text{Hz}}{1.554} = 10.3\text{Hz} \approx 10\text{Hz} \]

To see the detailed derivation of Equation 3, see “Appendix” on page 8.

For the noninverting amplifier, a passband gain of 100V/V requires the ratio $R_F/R_G = 99$. Here, the compromise between the feedback path loading, noise contribution from $R_F$, and the capacitor size of $C_G$ lead to resistor values of $R_F = 100\text{k}\Omega$ and $R_G = 1.01\text{k}\Omega$.

The capacitance ($C_G$), required to place the $A_{CL}$ pole at 10.3Hz can now be calculated with:

\[ C_G = \frac{1}{2\pi f_{pG} R_G} = \frac{1}{2\pi \cdot 10.3\text{Hz} \cdot 1.01\text{k}\Omega} = 15\mu\text{F} \]
To calculate $R_{IN}$, the offset due to bias current flow must be considered. While $I_B^-$ flows through $R_F$ only, as the DC path through $R_G$ is blocked by $C_G$, $I_B^+$ flows through $R_{IN}$ and the parallel circuit of the two $R_B$ Resistors. To minimize this offset, the sum of $R_{IN}$ and $R_B/2$ must equal $R_F$. $R_{IN}$ is the difference between $R_F$ and $R_B/2$:

\[
R_{IN} = R_F - R_B/2 = 100\,k\Omega - 25k\Omega = 75k\Omega
\]  

(EQ. 5)

Then, the input capacitance that produces the corner frequency derived in Equation 3 on page 6 is calculated with:

\[
C_{IN} = \frac{1}{2\pi f_{CL} R_{IN}} = \frac{1}{2\pi \cdot 10.3Hz \cdot 75k\Omega} = 206nF
\]  

(EQ. 6)

Again, to make the roll-off occur slightly earlier, choose the next higher standard value with $C_{IN} = 220nF$.

Figure 13 shows the final amplifier circuit and its corresponding frequency responses.

In the case of the inverting amplifier in Figure 14, the biasing voltage divider remains the same as in the noninverting case. The $V_S/2$ potential at the noninverting input also appears at the inverting input due to feedback action.

Note: The gain setting resistor ($R_G$), also represents the input resistance ($R_{IN}$). With that, the closed-loop gain of the amplifier becomes the overall circuit gain. Its pole frequency is the cutoff frequency of the input stage and therefore, the minimum input bandwidth of the circuit. This simplifies the calculation of component value enormously.

The main difference is, $I_B^+$ only flows through the parallel circuit of the two bias resistors. Therefore, reducing the offset due to bias current forces $R_F$ to drop in value to match $R_B/2$. For high gain applications with low input bandwidth, this necessitates a much smaller $R_G$ value, which in turn requires the increase of $C_{IN}$.

The simple design procedure would be to:

- Make $R_F = R_B/2$ for offset reduction,
- Then deriving the gain resistor with $R_G = R_F/G$
- And finally, calculating the input capacitor with $C_{IN} = 1/(2\pi f_{IN} R_G)$
Applying the above equations to an AC-coupled inverting amplifier with passband gain of $G = 100\,\text{V/V}$ and minimum input bandwidth of $f_{\text{IN}} = 16\,\text{Hz}$, while maintaining $R_B = 49.9\,\Omega$, results in $R_F = 24.9\,\Omega$, $R_G = 249\,\Omega$, and $C_{\text{IN}} = 47\,\mu\text{F}$.

Setting the supply-noise roll-off of the voltage divider to $1/10^{\text{th}}$ of $f_{\text{IN}}$ makes $f_{\text{CB}} = 1.6\,\text{Hz}$, which yields a capacitor value of $C_B = 1/(2\pi f_{\text{CB}} R_B/2) = 4\,\mu\text{F}$. Again, chose the next higher standard value of $4.7\,\mu\text{F}$.

**Figure 15** shows the actual circuit with its corresponding frequency responses.

![Figure 15. AC-Coupled Inverting Amplifier with ISL28134](image)

4. **Appendix**

4.1 **Deriving the -3dB Frequency ($f_{\text{CT}}$), of the Overall Circuit Gain ($G_T$)**

**Figure 16** shows the AC-signal equivalent schematic of the noninverting amplifier in **Figure 8 on page 4** and its individual and overall gain responses.

![Figure 16. AC-Signal Equivalent Circuit of Figure 8 and its Frequency Responses](image)

The gain responses for $G_{\text{IN}}$ and $A_{\text{CL}}$ are:

$$G_{\text{IN}} = \frac{j\omega C_{\text{IN}} R_{\text{IN}}}{1 + j\omega C_{\text{IN}} R_{\text{IN}}} \quad \text{with pole frequency } f_{\text{CL}} = \frac{1}{2\pi C_{\text{IN}} R_{\text{IN}}}.$$  

$$A_{\text{CL}} = \frac{1 + j\omega C_G (R_G + R_F)}{1 + j\omega C_G R_G} \quad \text{with pole frequency } f_{\text{PD}} = \frac{1}{2\pi C_G R_G}, \text{ zero frequency } f_{\text{ZG}} = \frac{1}{2\pi C_G (R_G + R_F)}, \text{ and passband gain } A_0 = \frac{R_G + R_F}{R_G}.$$  

Because the ratio of pole to zero frequency is: $\frac{f_{\text{PD}}}{f_{\text{ZG}}} = \frac{R_G + R_F}{R_G} = A_0$, it follows that $f_{\text{PD}} = \frac{f_{\text{ZG}}}{A_0}$.  


Expressing the transfer functions in their generic forms gives:

\[ G_{IN} = \frac{j\omega/\omega_{CI}}{1+j\omega/\omega_{CI}} \quad \text{and} \quad A_{CL} = \frac{1+j\omega/\omega_{pG}}{1+j\omega/\omega_{pG}} = \frac{1+jf/f_{zG}}{1+jf/f_{pG}} \]

This makes the overall gain: \( G_T = G_{IN} \cdot A_{CL} = \frac{jf/f_{CI} \cdot (1+jf/f_{zG})}{1+jf/f_{CI} \cdot (1+jf/f_{pG})} \)

And its magnitude function:

\[ |G_T| = \sqrt{\frac{(f/f_{pG})^2}{1+(f/f_{pG})^2} \cdot \frac{1+(f/A_0 f_{pG})^2}{1+(f/f_{pG})^2}} \]

By making \( f_{CI} = f_{pG} \) and replacing \( f_{zG} \) with \( f_{pG}/A_0 \) the magnitude function becomes:

\[ |G_T| = \sqrt{\frac{(f/f_{pG})^2}{1+(f/f_{pG})^2} \cdot \frac{1+(f\cdot A_0/f_{pG})^2}{1+(f/f_{pG})^2}} \]

To find \( f_{CT} \), \( |G_T| \) is set \( A_0/\sqrt{2} \) (the -3dB magnitude of \( A_0 \)) and the generic frequency, \( f \), is replaced with \( f_{CT} \):

\[ A_0 = \frac{f_{CT}^2 f_{pG}^2}{1+f_{CT}^2} \cdot \frac{1+f_{CT}^2 A_0 f_{pG}^2}{1+f_{CT}^2} \quad \Rightarrow \quad A_0^2 = \frac{f_{CT}^2}{f_{pG}^2} \left( \frac{1+f_{CT}^2 A_0^4}{f_{pG}^4} \right) \]

Then, solving for \( f_{CT} \) results in:

\[ f_{CT} = f_{pG} \sqrt{1-1/A_0^2 + (1-1/A_0^2)^2} \]

Therefore, for \( A_0 > 10 \), \( f_{CT} = f_{pG} \cdot \sqrt{1+\sqrt{2}} = 1.55 f_{pG} \).

Using 1.55 for \( A_0 \geq 2V/V \), yields less than 9% deviation from the theoretical \( f_{CT} \) value.

Plotting the factor of \( f_{pG} \) over \( A_0 \), one might simply use an average factor of 1.5.

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**Figure 17.** \( F_{CT}/F_{pG} \) Ratio vs Passband Gain
5. Revision History

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Corporate Headquarters
TOYOSU FORESIA, 3-2-24 Toyosu,
Koto-ku, Tokyo 135-0061, Japan
www.renesas.com

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