7.6
Op-Amp Transimpedance Amplifier

A transimpedance amplifier (TIA) converts a current to a voltage and is often used with current-based sensors like photodiodes. It's also a common building block that helps explain the performance and stability limits of many other op-amp circuits.

Our previous op-amp circuits have used voltage as a signal, but many real-world sensors like photodiodes are physically based on perceiving changes in current, not voltage. This makes it useful to be able to convert a current to a voltage drop.

Ohm's Law tells us that a simple resistor (i.e. an ordinary, real impedance) converts a current to a voltage drop, so why do we need to build an amplifier circuit to accomplish the same thing? Two answers:

1. A bare resistor as a current-to-voltage device will have low frequency bandwidth if there's significant input capacitance, which there often is.
2. Depending on the situation, we may be concerned with maximum signal transfer and minimizing interstage loading.

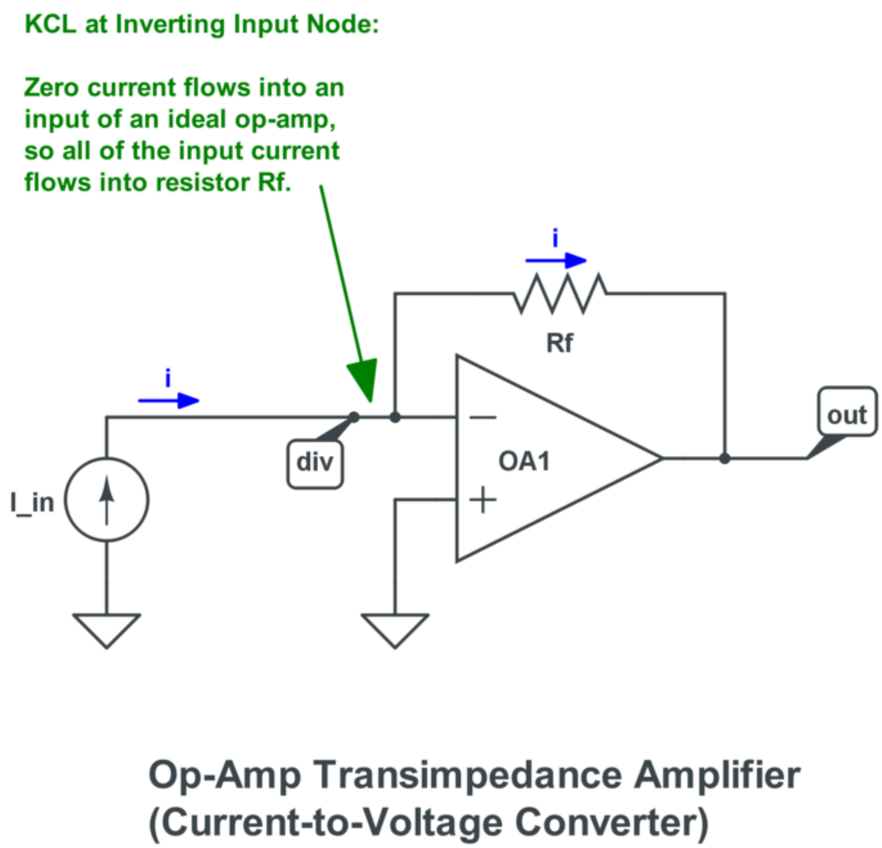
Both of these answers boil down to input impedance. A large resistor presents a large input impedance, when we really want our current-to-voltage converter to have a low (near zero) input impedance in order to maximize signal transfer from a high-impedance current signal source.

Fortunately, adding an ideal op-amp allows us to control both the input impedance and output impedance and make a much improved **current-to-voltage converter**. This overall circuit is called a **transimpedance amplifier** (abbreviated TIA).

It's a *trans*impedance, rather than an ordinary impedance, because the current at one place in the circuit becomes a voltage elsewhere in the circuit, rather than having the voltage affect the original current as happens in an ordinary impedance.

From Inverting Amp to Transimpedance Amp

The basic op-amp transimpedance amplifier looks like this, with the op-amp's non-inverting (+) input grounded, and a feedback resistor R_f between inverting (-) input and output:



Op-Amp Transimpedance Amplifier - Labeled
circuitlab.com/c9c575e8ks4rq

Edit - Simulate

The input current flows entirely through the feedback resistor, and the op-amp adjusts its voltage output to keep its inputs at equal voltages.

This is identical to the op-amp inverting amplifier except that we've removed the input resistor R_{in} and the input voltage V_{in} , and simply replaced them with a current source I_{in} .

In fact, it's reasonable to go back and think about the inverting amplifier as being composed of two stages: a voltage-to-current converter (resistor R_{in}), combined with a current-to-voltage transimpedance amplifier (op-amp OA1 and resistor R_{f}). By converting from voltage to current and then from current back to voltage, we achieve an overall voltage-to-voltage signal pathway. (The virtual ground behavior of the op-amp allows the voltage-to-current converter R_{in} to function correctly by keeping one of its terminals V_{div} at fixed voltage.)

Solving the Equations

From the perspective of Kirchhoff's Current Law at the op-amp's inverting input node, it is indistinguishable as to whether the current contribution comes from a resistor $\frac{V_{\text{in}}}{R_{\text{in}}} = i_{\text{in}}$ (from the previous op-amp inverting amplifier section) or simply from an externally-specified input current i_{in} as in the transimpedance amplifier.

Ohm's Law gives us an equation for the resistor:

$$V_{\text{div}} - V_{\text{out}} = i_{\text{in}} R_{\text{f}}$$

The ideal op-amp gives us one more equation, ensuring its two inputs to be equal when the feedback loop is closed:

$$V_{\text{div}} = 0$$

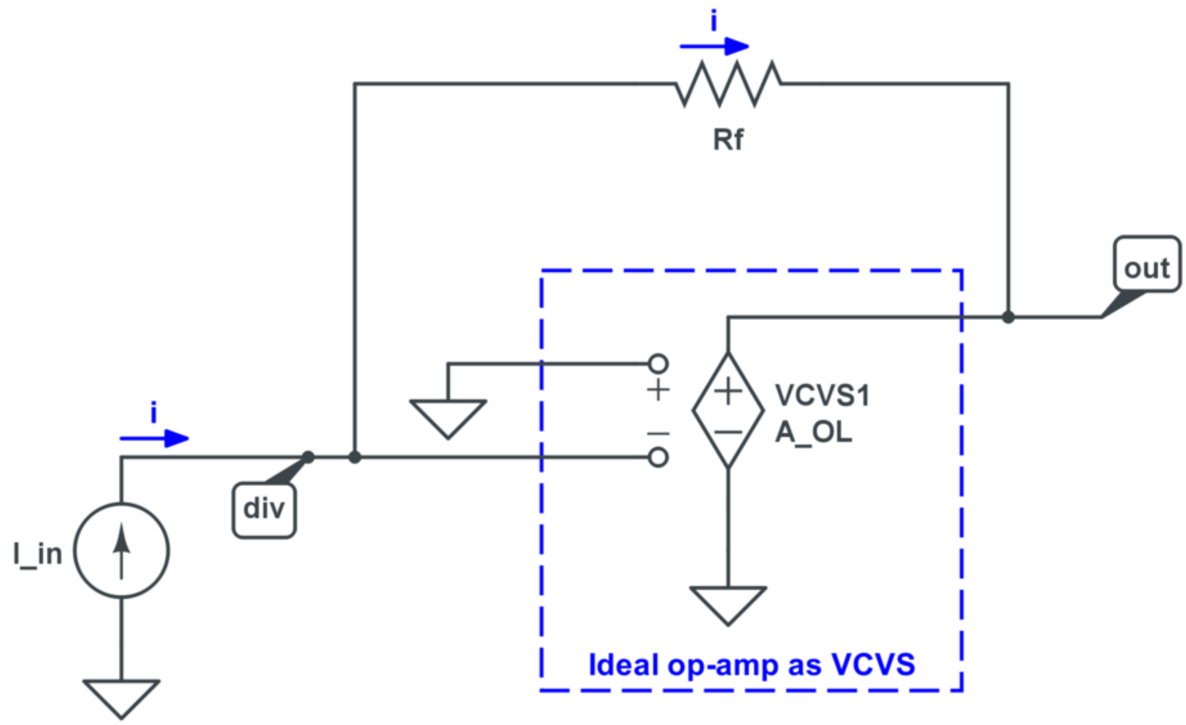
Combining these two, we find simply:

$$\begin{aligned} -V_{\text{out}} &= i_{\text{in}} R_{\text{f}} \\ V_{\text{out}} &= -i_{\text{in}} R_{\text{f}} \end{aligned}$$

The transimpedance of this amplifier is simply R_{f} . We might say it's $-R_{\text{f}}$, but generally wouldn't include the negative sign in our description of the transimpedance, because the sign is simply a function of our choice in labeling the direction of the current i_{in} and doesn't affect the behavior of the circuit. Regardless, as we've drawn it above, an increase in the current $+\Delta i_{\text{in}}$ causes a change in output voltage of $-R_{\text{f}} \Delta i_{\text{in}}$.

VCVS Model

A voltage-controlled voltage source (VCVS) model lets us examine more fine-grained behavior of the transimpedance amplifier and its limitations. As we did in the inverting amplifier section, we'll replace the ideal op-amp with a VCVS model. (For a more thoroughly worked solution, see the inverting amplifier section.)



Op-Amp Transimpedance Amplifier VCVS Model
circuitlab.com/cd5qtq6tcn2va

Edit - Simulate

The VCVS gives us one equation:

$$V_{\text{out}} = -A_{\text{OL}} V_{\text{div}}$$

Ohm's Law on the feedback resistor gives us a second:

$$V_{\text{div}} - V_{\text{out}} = i_{\text{in}} R_{\text{f}}$$

When we combine these two to eliminate V_{div} , we find:

$$\begin{aligned} V_{\text{out}} &= -A_{\text{OL}} V_{\text{div}} \\ V_{\text{out}} &= -A_{\text{OL}} \left(V_{\text{out}} + i_{\text{in}} R_{\text{f}} \right) \\ V_{\text{out}} (1 + A_{\text{OL}}) &= -A_{\text{OL}} i_{\text{in}} R_{\text{f}} \\ V_{\text{out}} &= -\left(\frac{A_{\text{OL}}}{1 + A_{\text{OL}}} \right) i_{\text{in}} R_{\text{f}} \end{aligned}$$

For an ideal op-amp we take the limit $A_{\text{OL}} \to \infty$, which causes the fraction $\frac{A_{\text{OL}}}{1+A_{\text{OL}}} \to 1$, again leaving us with:

$$V_{\text{out}} \approx -i_{\text{in}} R_{\text{f}}$$

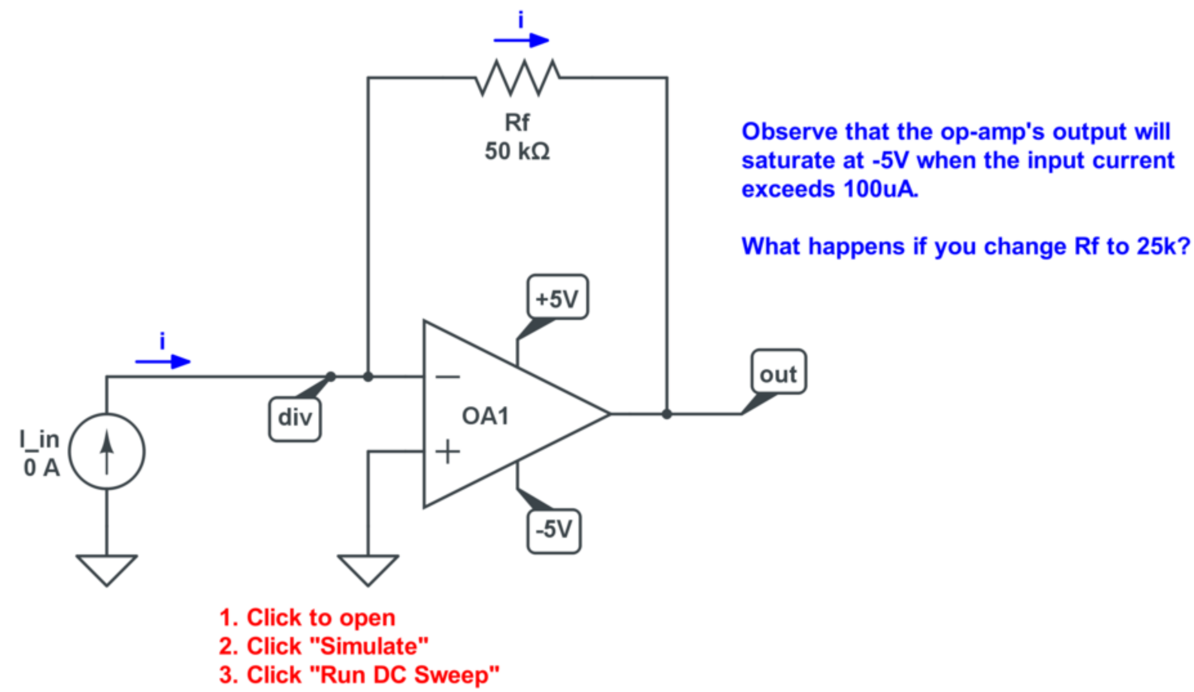
Choosing the Feedback Resistance

When considering the DC behavior, we only have one number to choose when designing a transimpedance amplifier: the resistance of the feedback resistor, R_f . We have constraints on both sides:

1. If R_f is **too large**, then our current signal will saturate the op-amp's output at either its positive or negative supply rail limits, causing clipping of the signal.
2. If R_f is **too small**, then our current signal will turn into too small of a voltage signal to be useful. Remember, our goal is to measure i_in . Creating a bigger change in V_out to measure by using a R_f that is sufficiently large enough will make it easier to detect small changes in our input current.

Therefore, we typically want to choose the largest possible resistance that just barely lets us cover the full range of input currents we want to measure.

For example, if our op-amp is powered by $\pm 5 \ \mathrm{V}$ and we want to measure currents up to $\pm 100 \ \mu\mathrm{A}$, then we should choose a resistance of about $R_\mathrm{f} = \frac{5 \ \mathrm{V}}{100 \ \mu\mathrm{A}} = 50 \ \mathrm{k}\Omega$:



Op-Amp Transimpedance Amplifier - 50k Resistance DC Sweep Simulation
circuitlab.com/cy56xh92e2v57

Edit - Simulate

Exercise Click to open and simulate the circuit above. Look at the plot of V_{out} vs. i_{in} . What happens if you change R_{f} to $25\ \text{k}\Omega$ and run the simulation again?

In a transimpedance amplifier, selecting the right resistance is a classic engineering **tradeoff between sensitivity and range**.

Advanced: Higher Dynamic Range

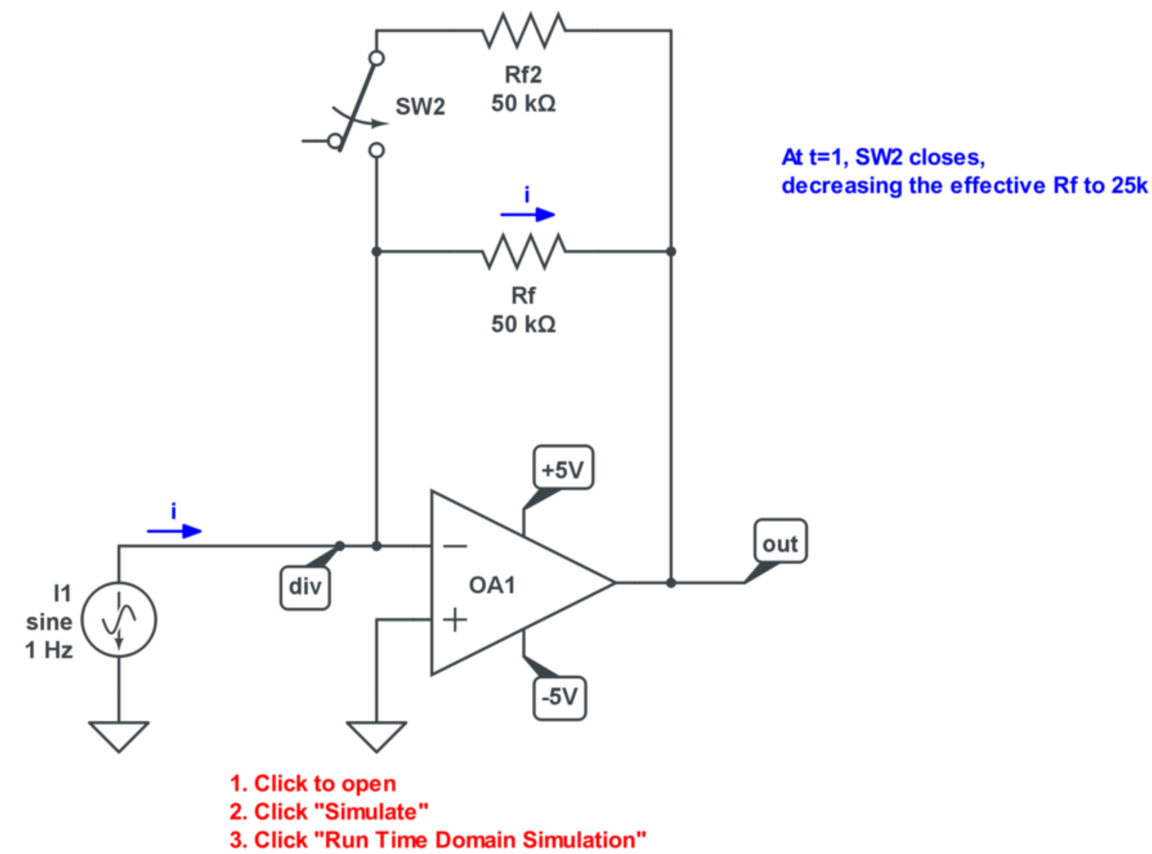
If you're looking to cover a very wide range of currents (high dynamic range, possibly over many orders of magnitude), there are two more advanced alternatives to consider:

Dynamically Adjust the Feedback Resistor

First, you might imagine being able to swap in different resistors for R_{f} to achieve different levels of transimpedance. If you observe that your output voltage is saturated, just swap in the next lower resistor. (This is how multimeters with user-selectable input ranges work manually!) These exist as advanced mixed-mode (analog and digital) ICs called

programmable gain transimpedance amplifiers.

Here's an example where the feedback resistance changes at $t = 1\ \mathrm{s}$:



Op-Amp Transimpedance Amplifier - Switched Resistance Simulation
circuitlab.com/c733z7nm5fbby

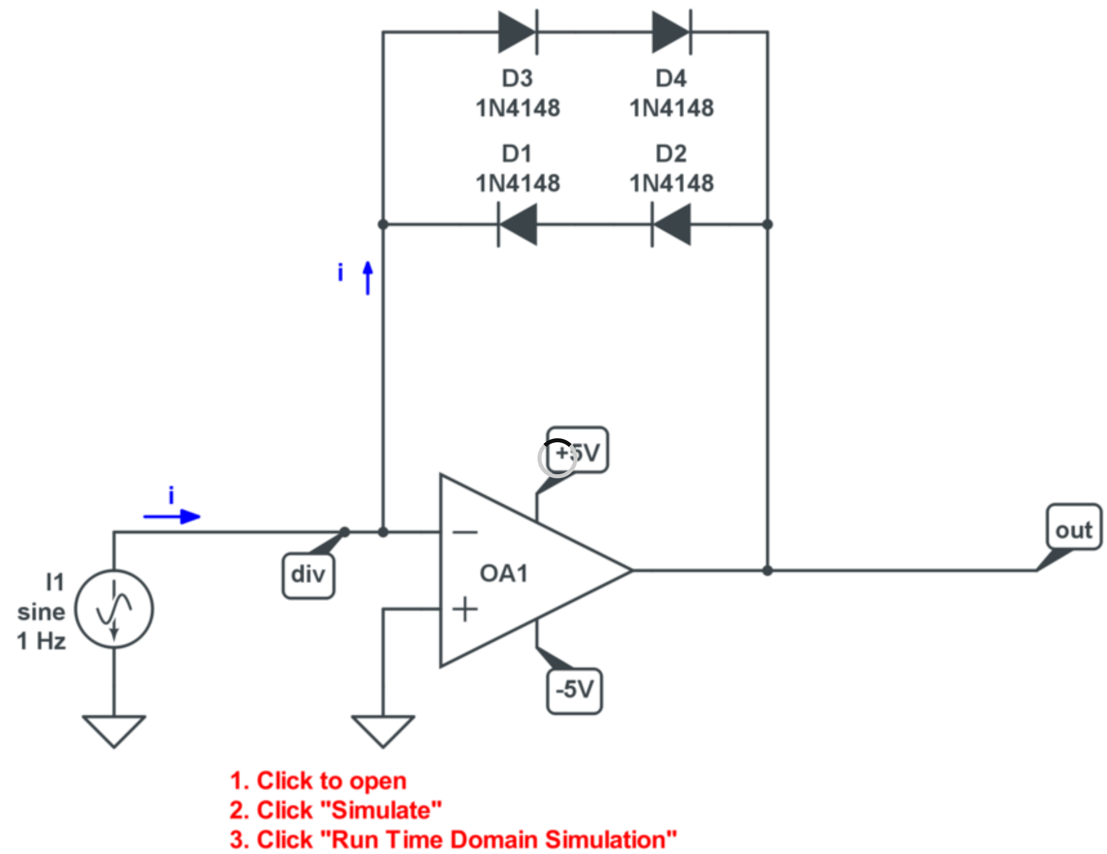
Edit - Simulate

Exercise Click to open and simulate the circuit above. How does the output for $t < 1\ \mathrm{s}$ look different than the output for $t > 1\ \mathrm{s}$?

Logarithmic Response

If you're willing to sacrifice linearity, you might consider using a **logarithmic** response to current. For example, an ordinary P-N junction diode has a voltage drop that changes with the log of the current through it, approximately $+60\ \mathrm{mV}$ for every 10x in current. By measuring the voltage across the diode, you can effectively measure currents over many orders of magnitude. (Place two diodes in series and you have doubled the sensitivity to $+120\ \mathrm{mV}$ per 10x in current!)

Here's an example with two diodes, instead of a feedback resistor:



Op-Amp Transimpedance Amplifier - Logarithmic Response
circuitlab.com/c6gpwqp963hr7

Edit - Simulate

Exercise Click to open and simulate the circuit above. Does the shape of $V_{\text{out}}(t)$ match your expectation for the sinusoidal input?

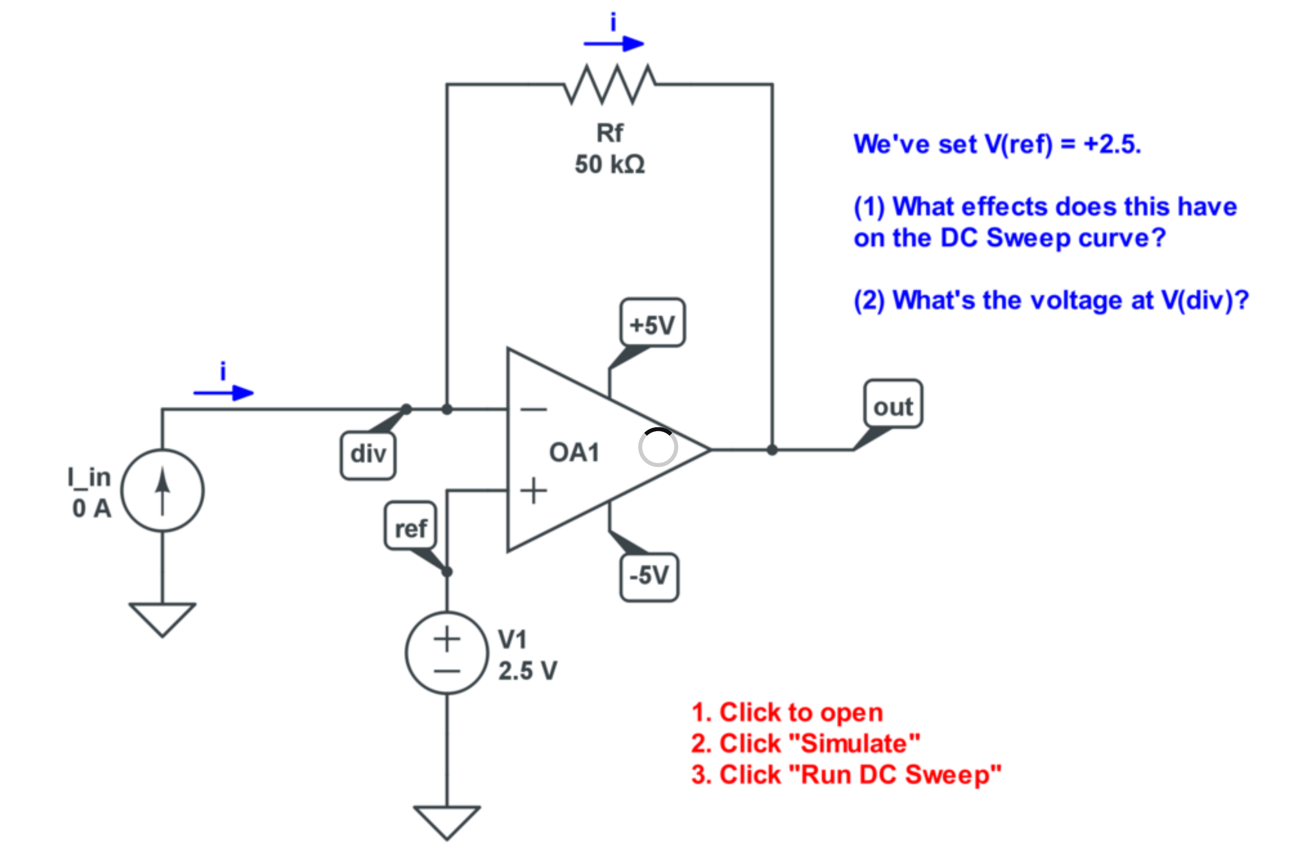
Note that we've used four diodes total (two diodes in each direction of current flow), but only one set is conducting at any given instant.

Moving the Bias Point

The op-amp transimpedance amplifier drawn earlier shows the op-amp's non-inverting (+) input connected to ground. As discussed in the Ground section, this is just a convenient labeling to indicate where our 0-voltage reference point is, but is otherwise nothing special.

It can be useful to pick a different voltage to be our reference. This is especially true if we're designing a transimpedance amplifier within a single supply system. This can be achieved by simply putting a different voltage on the op-amp's non-inverting (+) input, for example, $V_{\text{ref}} = +2.5\ \text{V}$. (In practice, this voltage might come from a voltage divider, an op-amp voltage reference, or an IC linear voltage regulator, for example.) What happens when we change V_{ref}

?



Op-Amp Transimpedance Amplifier - Moved Bias Point - DC Sweep Simulation
circuitlab.com/c823eq49s6h7k
Edit - Simulate

Exercise Click to open and simulate the circuit above. What's different about the DC Sweep I-V curve compared to the previous example? Next, try adding a second DC parameter sweep, changing `V1.V` from 0 to 5 in 0.5 V linear steps and look at the resulting simulation plot.

Changing V_{ref} has two important consequences:

First, it shifts the range of currents that the amplifier is able to measure before saturating. Previously, we could only measure $\pm 100\ \mu\text{A}$ before the amplifier would saturate at the supply rails. But now, we can go up to $+150\ \mu\text{A}$. We've sacrificed range on the negative side (now extending to only $-50\ \mu\text{A}$), but in most cases, we're more interested in measuring current in one direction than another, so that's a great tradeoff!

Second, changing V_{ref} also shifts the DC voltage that appears at V_{div} . Remember that the ideal op-amp adjusts its output to keep its input voltages precisely matched. Therefore, changing the fixed DC voltage at the non-inverting input leads to a change at the inverting input as well. This can be important because the input current source or sink may itself have a limited voltage range over which it can operate. Or, as we'll see in the photodiode example below, the behavior of the photodiode itself requires a particular bias voltage, one end of which is set by the op-amp's inverting input.

Moving the bias point doesn't change the fundamental current-to-voltage transimpedance slope of the circuit, but is a useful practical tool for moving the offset enough to make the overall circuit do what you want.

Input Impedance: Intuitive Model

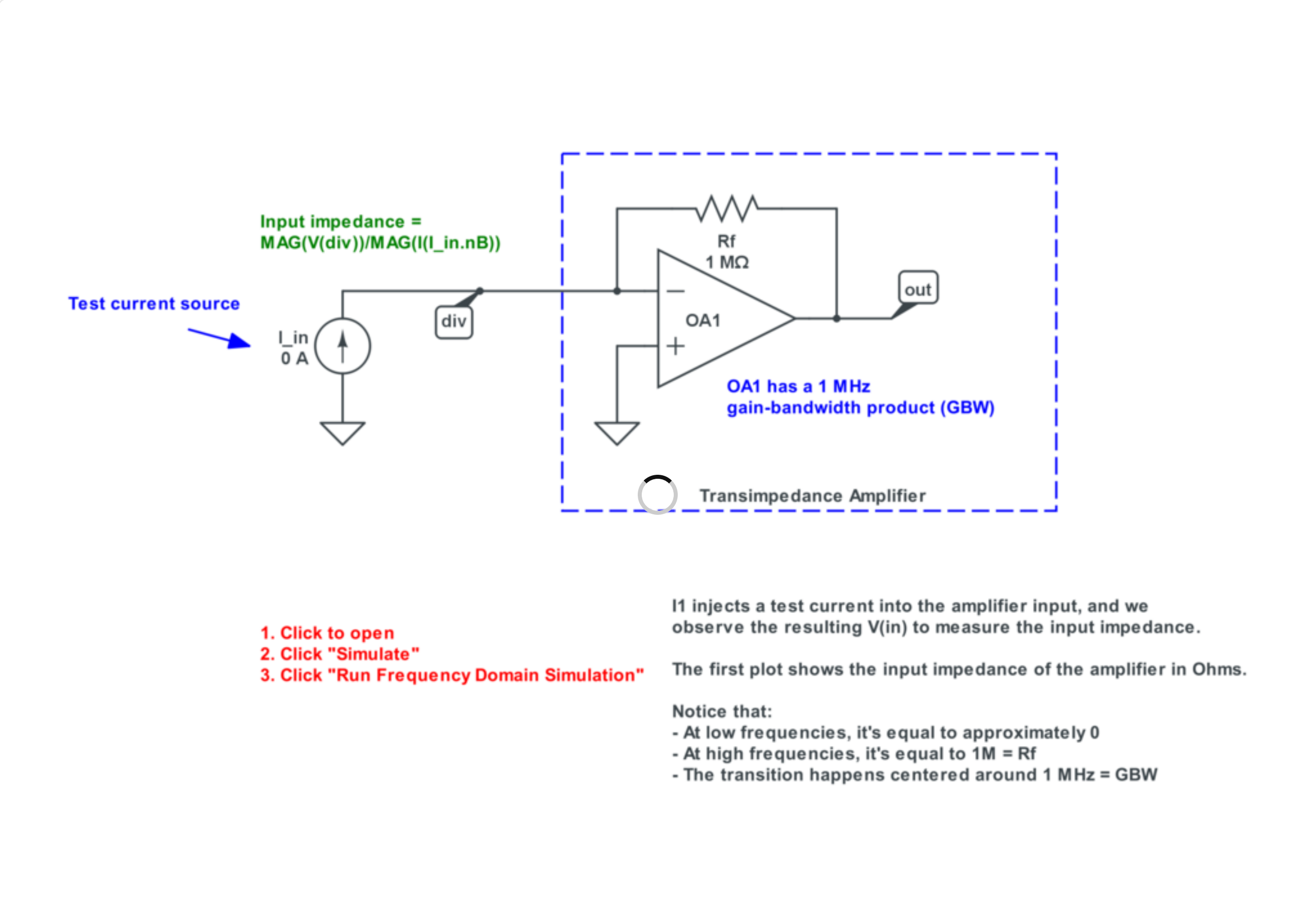
The transimpedance amplifier has interesting characteristics at different frequencies, and as we mentioned in this section's introduction, the input impedance of a theoretically perfect current-to-voltage converter would be $R_{\mathrm{in}} = 0$. In reality, however, an op-amp transimpedance amplifier's R_{in} varies widely with frequency, and this is the root of many issues we'll explore for the remainder of this section.

Understanding the input impedance of the op-amp transimpedance amplifier will not only help us manage the stability and bandwidth of the transimpedance amplifier itself, but will also help us design other closely related op-amp circuits like inverting amplifiers and differentiators.

The input impedance of a transimpedance amplifier varies tremendously with frequency.

- For frequencies much lower than the op-amp's gain-bandwidth product $f \ll \mathrm{GBW}$, the input impedance $R_{\mathrm{in}} \approx 0$.
- For frequencies much higher than the op-amp's gain-bandwidth product $f \gg \mathrm{GBW}$, the input impedance $R_{\mathrm{in}} \approx R_{\mathrm{f}}$.

We can see this easily through simulation:



Op-Amp Transimpedance Amplifier - Input Impedance
circuitlab.com/ce6tag6r6t3jj
Edit - Simulate

Exercise Click to open and simulate the circuit above. First, observe how the magnitude of the input impedance (top graph) changes with frequency. Then, click "Advanced Graphing" and change the magnitude plot from a linear to a logarithmic y-axis scale. Do you notice anything interesting about the shape of the plot? Finally, try changing the op-amp's GBW to be 1/10th as large. What happens to the input impedance graph?

In the simulation, notice also that the input impedance begins to rise 1-2 decades earlier than the GBW frequency. This is important because transimpedance amplifiers are often used to measure very small currents, and therefore have very large R_f , so even a small fraction of it can be large enough to compromise the behavior of the circuit at lower frequencies.

Intuitively, the crossover happens because at frequencies above the GBW frequency, variations in the op-amp's input voltage no longer cause equally large variations in the output. This causes the fraction $\frac{A_\mathrm{OL}(s)}{1+A_\mathrm{OL}(s)}$ to drop toward zero, since by definition $|A_\mathrm{OL}(j \cdot 2\pi \mathrm{GBW})| = 1$.

At low frequencies, a voltage increase at the input (caused by test current injection) is quickly cancelled out by a decrease in the output. But at high frequencies, the fluctuating voltage on the input is alternating too fast for the op-amp to respond, so it's *not* cancelled by an opposite fluctuation from the output. High input impedance at high frequencies is the result.

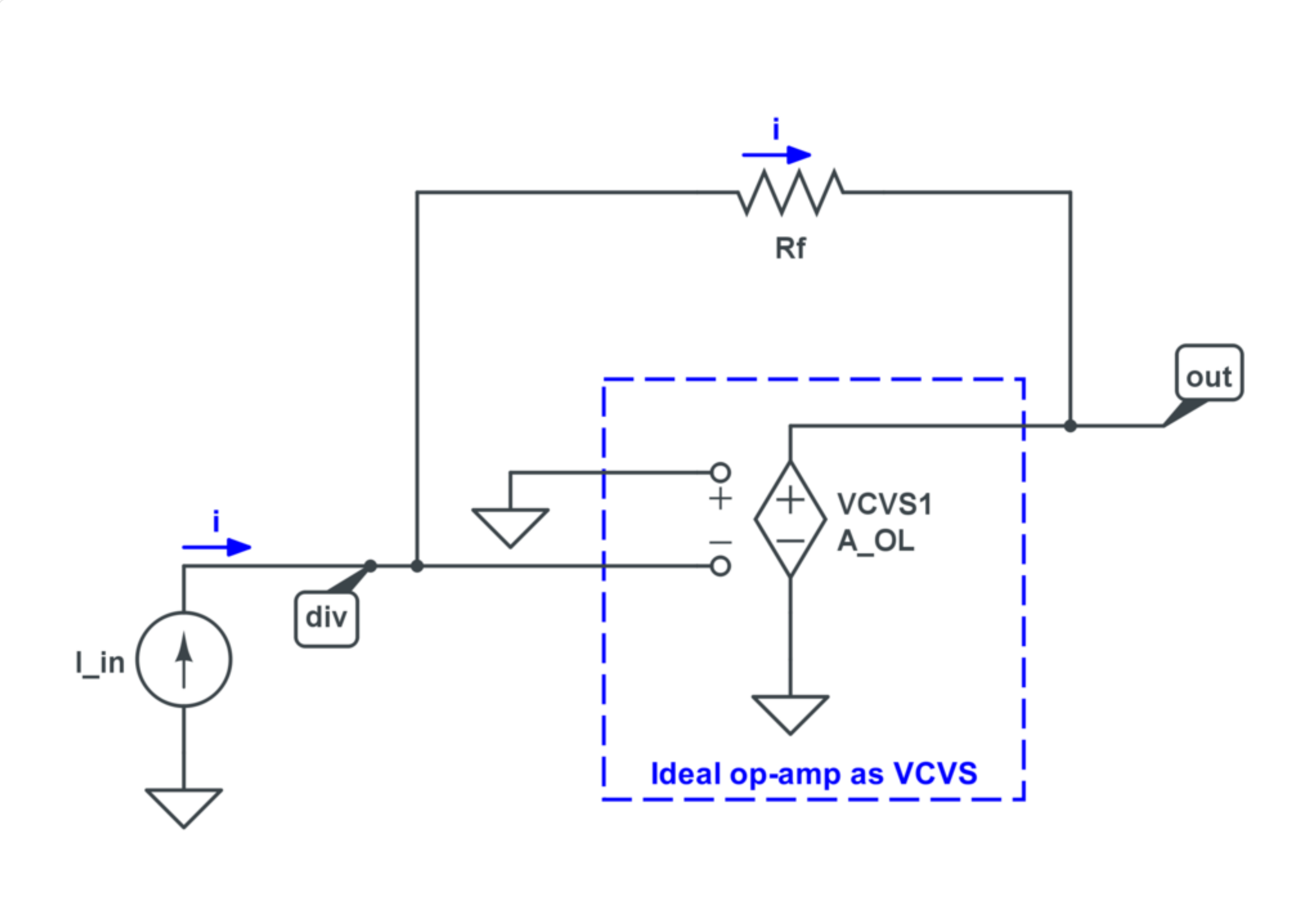
Input Impedance: VCVS Model and Equivalent Inductance

The transimpedance amplifier has a particularly interesting result when we find the input impedance Z_{in} algebraically, which helps explain some of the unusual behavior of this circuit. (We hinted at this when we plotted the input impedance on a logarithmic scale in the interactive exercise above.)

In our earlier VCVS solution above, we eliminated V_{div} to find an equation relating V_{out} to i_{in} . But now we're interested in input impedance:

$$Z_{\text{in}} = \frac{V_{\text{div}}}{i_{\text{in}}}$$

In this case, we want to take the same two circuit equations but use them to eliminate V_{out} instead. Let's refer to the same VCVS-based schematic as earlier:



Op-Amp Transimpedance Amplifier VCVS Model
circuitlab.com/cd5qtq6tcn2va
Edit - Simulate

As before, the VCVS gives us one equation:

$$V_{\text{out}} = -A_{\text{OL}} V_{\text{div}}$$

and Ohm's Law on the feedback resistor gives us a second:

$$V_{\text{div}} - V_{\text{out}} = i_{\text{in}} R_{\text{f}}$$

When we combine these two to eliminate V_{out} , we find:

$$\begin{aligned} V_{\text{div}} \left(1 + A_{\text{OL}}\right) &= i_{\text{in}} R_{\text{f}} \\ \frac{V_{\text{div}}}{i_{\text{in}}} &= \frac{R_{\text{f}}}{1 + A_{\text{OL}}} \\ Z_{\text{in}} &= \frac{R_{\text{f}}}{1 + A_{\text{OL}}} \end{aligned}$$

To capture the frequency dependent effects, we can substitute in the Laplace Transfer Function $G(s)$ of the one-pole op-amp model in place of the DC-only A_{OL} , which is:

$$G(s) = \frac{A_{\text{OL}}}{1 + s\left(\frac{A_{\text{OL}}}{2\pi \text{GBW}}\right)}$$

This $G(s)$ includes both the DC gain A_{OL} as well as the one-pole rolloff that dominates most of the frequency range. Recall that $s = j\omega = j2\pi f$:

$$G(f) = \frac{A_{\mathrm{OL}}}{1 + j2\pi f\left(\frac{A_{\mathrm{OL}}}{2\pi \mathrm{GBW}}\right)}$$
$$G(f) = \frac{A_{\mathrm{OL}}}{1 + jf\left(\frac{A_{\mathrm{OL}}}{\mathrm{GBW}}\right)}$$
$$G(f) = \frac{A_{\mathrm{OL}} \cdot \mathrm{GBW}}{\mathrm{GBW} + jfA_{\mathrm{OL}}}$$

At frequencies where $f \cdot A_{\mathrm{OL}} \gg \mathrm{GBW}$, which covers the vast majority of the operating bandwidth of most op-amps, we can make an algebraic approximation $G'(f)$ which is a simplified version of $G(f)$:

$$G'(f) = \frac{A_{\mathrm{OL}} \cdot \mathrm{GBW}}{jfA_{\mathrm{OL}}}$$
$$G'(f) = \frac{\mathrm{GBW}}{jf}$$
$$G'(f) = -j\frac{\mathrm{GBW}}{f}$$
$$G'(s) = \frac{2\pi \mathrm{GBW}}{s}$$

This tells us that the magnitude of the op-amp open-loop response is approximately $|G'(f)| = \frac{\mathrm{GBW}}{f}$ over a wide range of frequencies. (The $-j$ simply tells us about the relative phase of that response.) This is the definition of the gain-bandwidth product: the gain $|G'(f)|$ at a given frequency f is equal to the gain-bandwidth product GBW divided by the frequency f .

Substituting in our simplified $G'(s)$ in place of the original DC-only A_{OL} in our Z_{in} equation above, we find:

$$Z_{\mathrm{in}}(s) = \frac{R_{\mathrm{f}}}{1 + G'(s)}$$
$$Z_{\mathrm{in}}(s) = \frac{R_{\mathrm{f}}}{1 + \frac{2\pi \mathrm{GBW}}{s}}$$
$$Z_{\mathrm{in}}(s) = \frac{sR_{\mathrm{f}}}{2\pi \mathrm{GBW} + s}$$

As we did in our intuitive input impedance analysis, we can analyze this equation separately for both high and low frequencies.

At high frequencies where $f \gg \mathrm{GBW}$, then the $s = j2\pi f$ term dominates the denominator, and we have $Z_{\mathrm{in}}(s) \approx \frac{sR_{\mathrm{f}}}{s} = R_{\mathrm{f}}$. This matches our earlier work: at high frequencies, the input impedance looks like just the feedback resistor.

However, at low frequencies, things get quite interesting. When we assume $f \ll \mathrm{GBW}$, then the $s = j2\pi$ term in the denominator becomes insignificant, and we're left with:

$$Z_{\mathrm{in}}(s) \approx \frac{sR_{\mathrm{f}}}{2\pi\mathrm{GBW}}$$

This input impedance is proportional to frequency and in fact looks just like the impedance of an **inductor**, $Z_{\mathrm{L}} = sL$. The equivalent inductance is $L_{\mathrm{eq}} = \frac{R_{\mathrm{f}}}{2\pi\mathrm{GBW}}$. For a $R_{\mathrm{f}} = 5\ \mathrm{M\Omega}$ and $\mathrm{GBW} = 4\ \mathrm{MHz}$, the $L_{\mathrm{eq}} = 0.2\ \mathrm{H}$. This is large; the equivalent inductances can be quite significant for slow op-amps and large transimpedances.

The inductance is particularly a problem because once we combine an inductance and a capacitance, we have a second-order system and there can be a resonance, where energy repeatedly flows back and forth between the inductance and the capacitance. This resonance is undesirable, and next we'll see how it affects the response of the transimpedance amplifier, and finally learn how to mitigate it.

Input Capacitance and Frequency Response

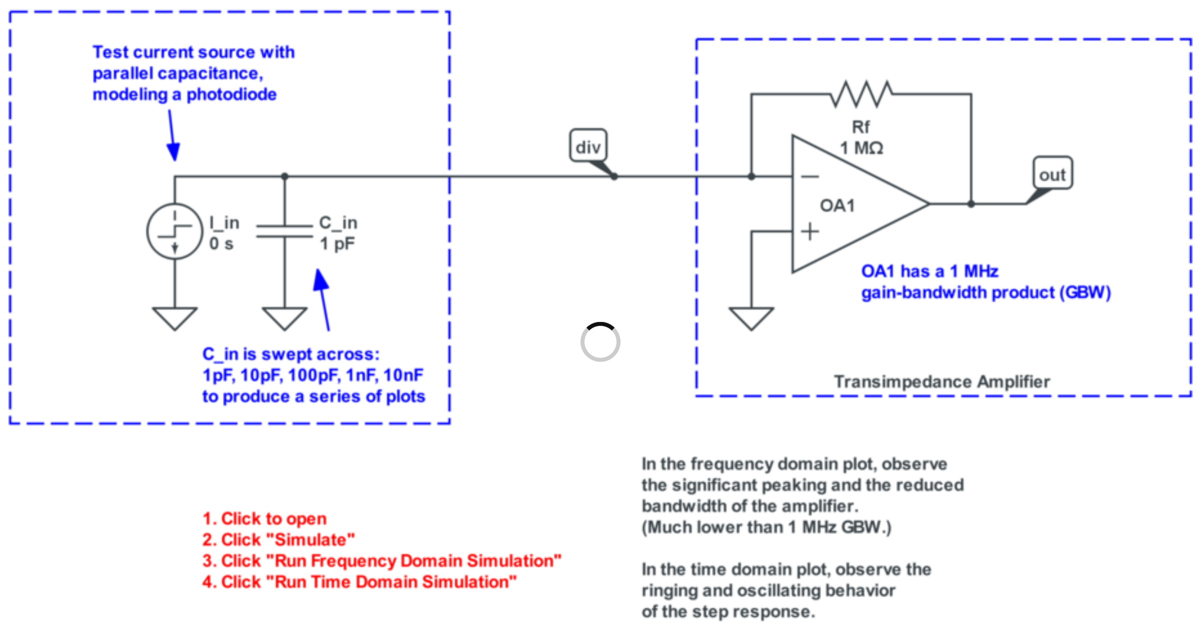
The DC input impedance of the transimpedance amplifier is approximately zero. However, when considering higher frequency effects, it would be wrong to assume the input impedance remains zero at higher frequencies because it actually rises drastically.

When a large impedance is combined with even a tiny amount of capacitance, the result is a large $\tau = RC$ time constant, significantly reducing the overall bandwidth of the amplifier circuit.

Even worse, when that impedance is inductive in phase, the result can be undesirable instability, resonance, and oscillation.

Further compounding the issue, transimpedance amplifiers are often used with photodiodes, where there's a large depletion capacitance in parallel with the photocurrent source.

This problem is most easily demonstrated with a quick simulation where we add small amounts of capacitance C_{in} at the input node:



Op-Amp Transimpedance Amplifier - Frequency Response and Input Capacitance
circuitlab.com/cv5d652s6w55e Edit - Simulate

Exercise Click to open and simulate the circuit above. In the frequency domain plot, observe the significant peaking, and see that the bandwidth of the amplifier is much, much less than the op-amp's 1 MHz GBW. In the time domain plot, observe the ringing, rather than a clean step response.

With even small amounts of capacitance, the simulation shows that this amplifier is only marginally stable. It's much slower than we'd like, and has far too much ringing. In the real world, it may even be unstable, oscillating uncontrollably.

Due to parasitic capacitance, there will always be some amount of capacitance present, even if unintentional. If we're designing a transimpedance amplifier, it's imperative that we consider this and attempt to compensate for its these effects.

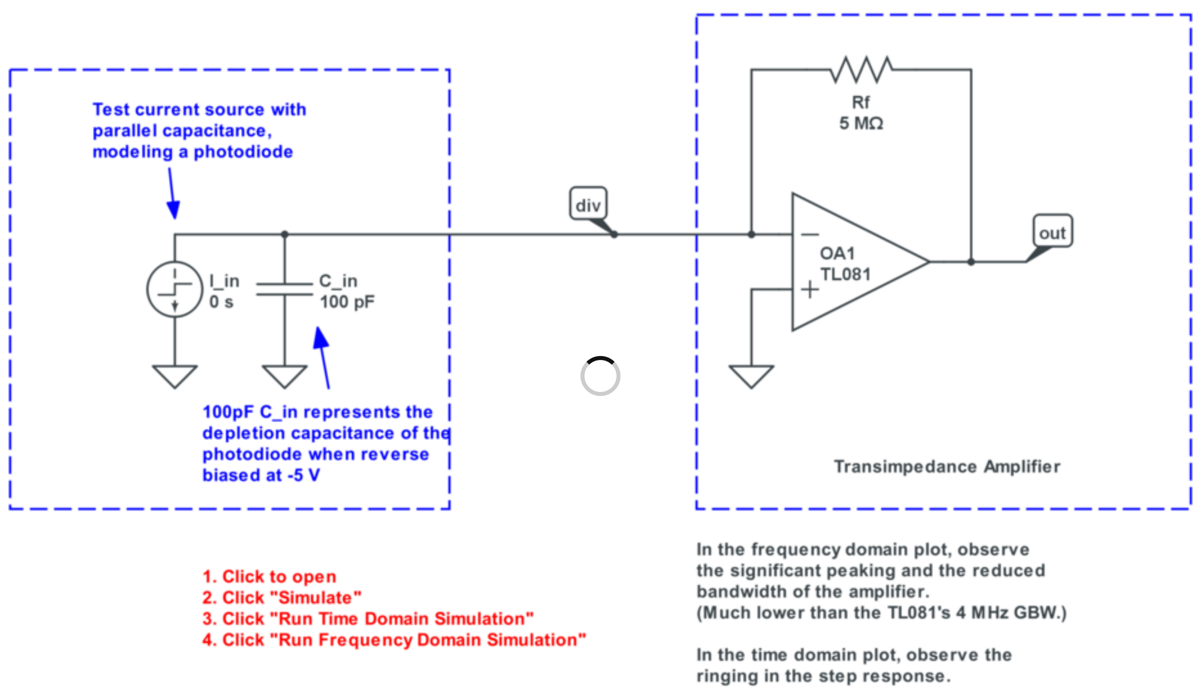
Stability and Compensation

Just as we addressed stability issues for the op-amp inverting amplifier and op-amp non-inverting amplifier circuits, we can correct for some of the bad behavior caused by input capacitance by adding a compensation network.

Relatively speaking, compared to the inverting and non-inverting amplifier compensation examples, compensation is more crucial for the typical transimpedance amplifier because the feedback resistance R_f and the input capacitance C_in both tend to be larger.

The simplest compensation network is simply to add a feedback capacitor C_f in parallel with R_f .

Let's consider an example where we're measuring a current of magnitude $|I_\mathrm{max}| = 1\ \mu\mathrm{A}$ and have $\pm 5\ \mathrm{V}$ voltage supplies, for an $R_\mathrm{f} = 5\ \mathrm{M\Omega}$. We'll also suppose that this is a photodiode, with depletion capacitance $C_\mathrm{in} = 100\ \mathrm{pF}$ when reverse biased at $-5\ \mathrm{V}$. What does the step response and frequency response look like when this transimpedance amplifier is uncompensated?



Op-Amp Transimpedance Amplifier - Photodiode Example - Uncompensated
circuitlab.com/cmy22d5hj37gq

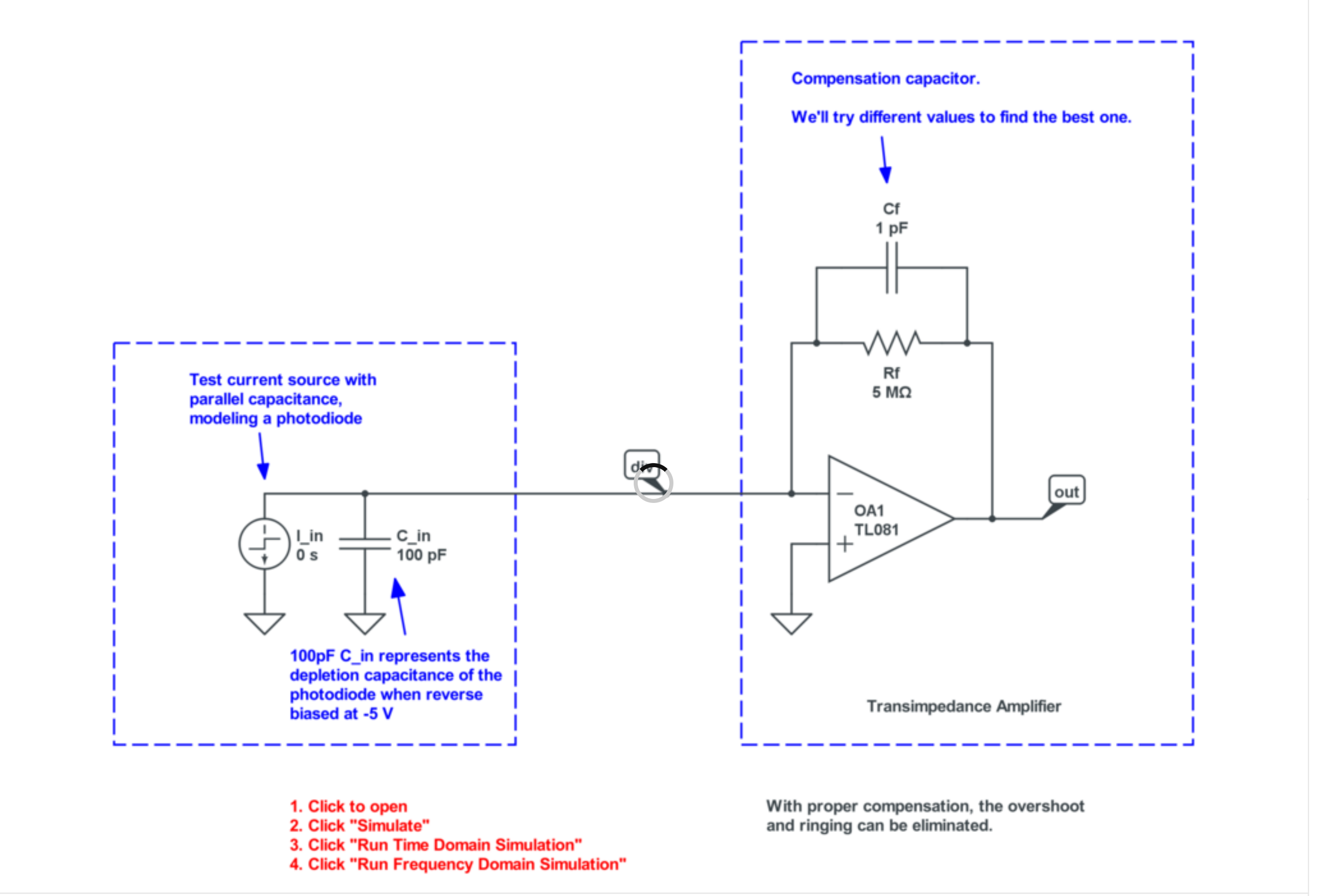
Edit - Simulate

Exercise Click to open and simulate the circuit above. Look at the ringing in the step response and peaking in the frequency response.

Uncompensated, this amplifier is marginally (i.e. barely) stable. The time domain simulation shows substantial ringing, rather than a nice clean step response.

Now, let's add a compensation capacitor C_f . How big should this capacitor be? The answer is complicated and involves interactions between the input capacitance, the feedback resistance, and the op-amp's transfer function. The easiest way to find the correct value is by simulation.

We'll use the simulator's parameter sweep to try different values, and observe the resulting time-domain step responses and frequency-domain responses:



Op-Amp Transimpedance Amplifier - Photodiode Example - With Compensation Capa... Edit - Simulate
circuitlab.com/cd57wr47kn9tj

Exercise Click to open and simulate the circuit above. Adjust the parameter sweep for `Cf.C` to compare multiple capacitor values.

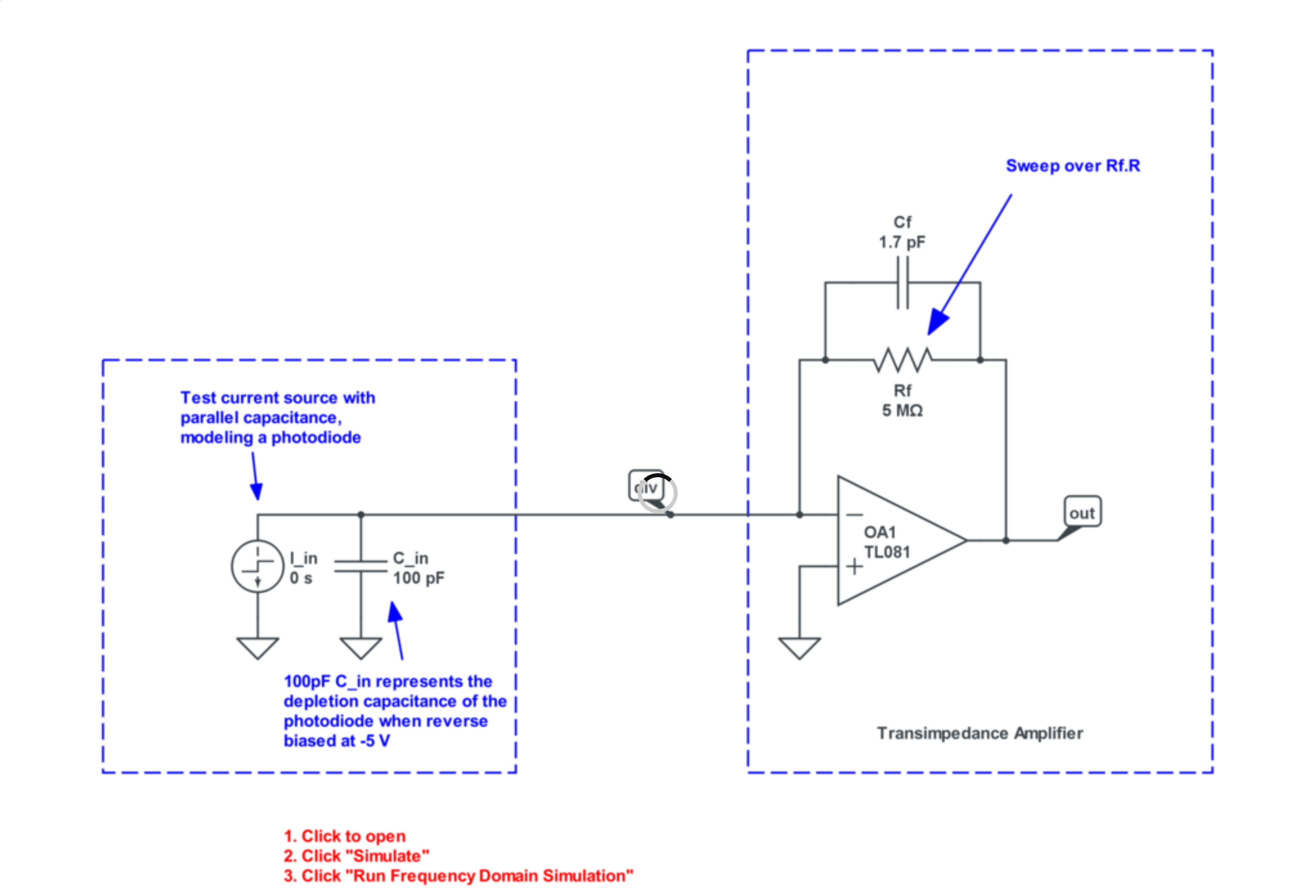
Note that even with $0.1\ \mathrm{pF}$ of capacitance (a tiny amount that even two adjacent wires might have!), the ringing of the step response is substantially reduced. But when we add just a bit more, we can get the overshoot to go away entirely. If you use the simulator's ability to sweep over the capacitance parameter, you can quickly compare step responses, and you'll find that somewhere in the neighborhood of $C_\mathrm{f} = 1.7\ \mathrm{pF}$ is probably optimal in this case.

Adding compensation to handle instability may be crucial if you want to avoid unintended oscillation in real circuits. In a circuit simulator, we know the value of the op-amp's parameters precisely. But in the real world, we don't. It's generally safer to **overcompensate**, adding a slightly larger C_f than predicted by simulation, to protect yourself from cases where the op-amp isn't performing up to spec. Overcompensation will lower the bandwidth, but will preserve stability and give a clean step response.

Improving the Frequency Response

While compensation does fix the instability, overshoot, and ringing, it does not address the reduced bandwidth of the amplifier. Even when properly compensated, the photodiode transimpedance amplifier example shown with $R_\mathrm{f} = 5\ \mathrm{M\Omega}$ and $C_\mathrm{f} = 1.7\ \mathrm{pF}$ only has a $-3\ \mathrm{dB}$ bandwidth of about $20\ \mathrm{kHz}$. This is quite low, especially when the op-amp has a $\mathrm{GBW} = 4\ \mathrm{MHz}$.

An easy way to see what's limiting the bandwidth is to use the simulator's parameter sweep to vary various values in the circuit and compare the resulting frequency-domain response plots. For example, let's sweep over different values of R_f :



Op-Amp Transimpedance Amplifier - Photodiode Example - Sweep over Rf
circuitlab.com/cw2bp62jnrs9u

Edit - Simulate

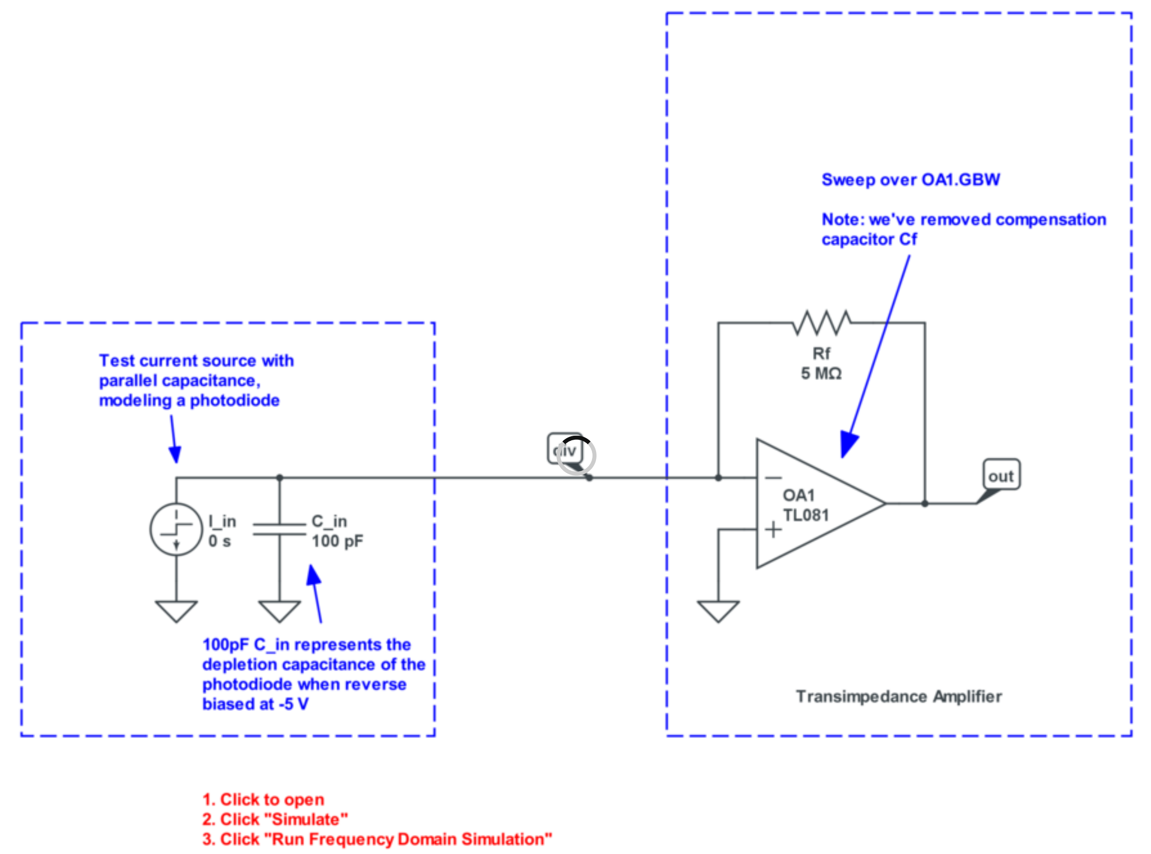
Exercise Click to open and simulate the circuit above. How does the circuit behave with different values for `Rf.R`?

Clearly, the simulation shows that changing R_f alters the DC transimpedance gain, and that lower values of transimpedance do correspond to higher bandwidths. However, this is a classic gain-bandwidth tradeoff: the magnitude plots all tail off and overlap at higher frequencies!

The horizontal section of the magnitude plot is set directly by the value of R_f , which varies for each parameter sweep run. In contrast, the declining section is set by $Z_{C_f} = \frac{1}{j\omega C_f}$, which is the same for each simulation run. At high frequencies, the parallel combination of $R_f // C_f$ is dominated by C_f . These facts explain the shape of the DC Sweep plot: separate traces at lower frequencies, overlapping and declining at higher frequencies.

In fact, this *must* be the case if the transimpedance amplifier is properly compensated. A larger C_f shifts the declining line down. A smaller C_f raises the line, until some other bandwidth-limiting element comes into play – in this case the op-amp itself.

What if we remove C_f but explore different op-amp parameters? We can consider whether a faster op-amp will help:



Op-Amp Transimpedance Amplifier - Photodiode Example - Sweep over GBW
circuitlab.com/c3wdb33g7f7r4

Edit - Simulate

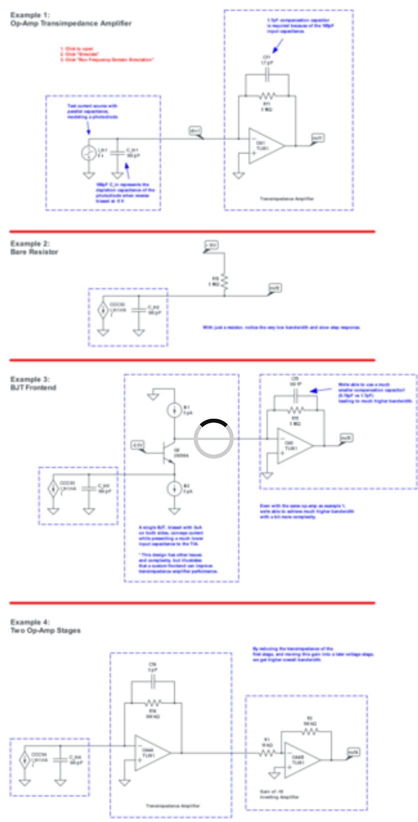
Exercise Click to open and simulate the circuit above. How does the circuit behave with different values for `OA1.GBW`?

Clearly, a faster op-amp (higher gain-bandwidth product) extends the bandwidth of the transimpedance amplifier. While we do see significant peaking in the magnitude plot, we know how to fix that by adding a compensation capacitor, which will be smaller for faster amplifiers. Even with a very fast hypothetical $\mathrm{GBW} = 1\ \mathrm{GHz}$ op-amp, the transimpedance amplifier has an overall bandwidth of barely $870\ \mathrm{kHz}$.

The key takeaway is that if you need to measure small currents at high bandwidth, you'll probably need either a very fast op-amp (properly compensated, of course), or to reduce your transimpedance and move that gain into a later voltage gain stage.

Advanced: Custom Frontends

If you need even more bandwidth, it is possible, even without faster op-amps. When performance matters, a single op-amp alone is not the answer. A simple demonstration here compares four ways of turning our photodiode current into a voltage, all with $5\ \mathrm{M\Omega}$ of overall transimpedance:



Op-Amp Transimpedance Amplifier - Photodiode Example - Alternatives
circuitlab.com/c8ypf3766ubz3

Edit - Simulate

Exercise Click to open and simulate the circuit above. How does the bandwidth compare for our four examples?

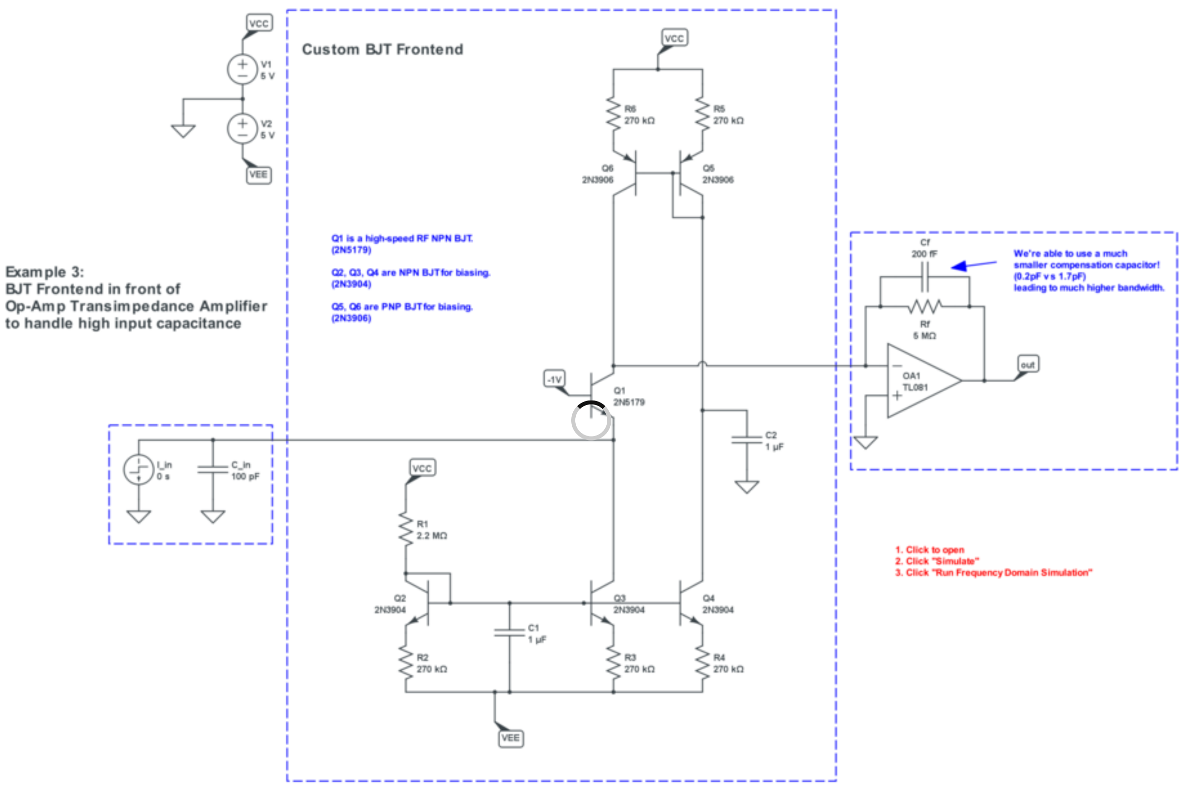
Example 1, our op-amp transimpedance amplifier, has around $24\ \mathrm{kHz}$ bandwidth.

Example 2, just a resistor (with no op-amp), is far worse with only about $0.3\ \mathrm{kHz}$ bandwidth. (Crucially, this demonstrates the need for a transimpedance amplifier in the first place! When an input capacitance is present, a resistor alone makes a poor current-to-voltage converter.)

Example 3 inserts a small BJT circuit between between the photodiode and the op-amp transimpedance amplifier, and achieves an impressive $182\ \mathrm{kHz}$ bandwidth, 7.5x that of example 1! It's not an especially practical frontend to build, and adds a lot of complexity to actually implement the biasing sources, but it does demonstrate that it's possible to dramatically improve the performance of a transimpedance amplifier by building a custom frontend.

Example 4 combines an op-amp transimpedance amplifier with $500\ \mathrm{k\Omega}$ with a second stage op-amp inverting amplifier with a gain of -10 , achieving an overall bandwidth of $82\ \mathrm{kHz}$. This is not as good as example 3, but is a much simpler solution.

We can develop Example 3 into a real-world example by using a higher speed BJT for Q1 and adding the rest of the circuit around it for biasing:



Op-Amp Transimpedance Amplifier - Photodiode Example - BJT Frequency Extension
circuitlab.com/cue38246z862s

Edit - Simulate

Exercise Click to open and simulate the circuit above. What's the $-3\ \mathrm{dB}$ bandwidth of this configuration?

With just a few discrete transistors and resistors, we've extended the bandwidth of our transimpedance amplifier from $24\ \mathrm{kHz}$ (example 1, op-amp only) out to $170\ \mathrm{kHz}$. A bit more cost, complexity, and power consumption, but a tremendous gain in performance!

The high-bandwidth current-to-voltage conversion problem (with nonzero input capacitance) is widespread in high-speed communication systems. If you're curious, search the literature and you'll find designs for BJT, JFET, and MOSFET-based transimpedance amplifier frontends.

What's Next

Return to the Table of Contents to continue. *More sections coming soon!*
