

SOT-23 10MHz Rail-to-Rail Op Amp Saves Board Space and Power

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Introduction

The new LT1797 op amp offers 10MHz gain bandwidth and rail-to-rail inputs and outputs in a tiny SOT-23 package. The usual bandwidth vs supply current trade-off has been optimized by designing on a 6GHz f_T process and running the transistors at reduced quiescent currents. The result is a fast, tiny, rail-to-rail device that brings to the table all the beauty that is the operational amplifier, while consuming minimal board space and power. The LT1797 is available in both I grade and C grade.

When comparing the LT1797 to competing products, don't stop at input offset voltage and input offset current. Be sure to look at open-loop gain, PSRR, CMRR, AC distortion, noise power, supply voltage range and output voltage swing. Table 1 shows some of the LT1797's specifications at a glance.

Applications

Fast Settling, Compact, -48V Current Sense

Figure 1 shows a compact, fast-settling current sense circuit designed for -48V supplies. Fast settling is required in applications that involve circuit-breaking or power rerouting. The circuit operates as follows. Current flowing through the sense resistor creates a voltage across it of $V = I_{SENSE} \times R_{SENSE}$. The LT1797's output rises until an equal voltage appears across R3, with the current flowing from Q1's emitter ($= I_{SENSE} \times R_{SENSE}/R3$). Q1's emitter current comes from its collector and appears as an easily measured voltage across R2. R2 is connected to a 3V supply, so the output voltage equation becomes $V_{OUT} = 3V - I_{SENSE} \times R_{SENSE} \times R2/R3$. With the values shown, the output response is -100mV per amp. R1 is a 5% resis-

Table 1. LT1797 specifications

Symbol	Parameter	Typical	Worst Case Over Temperature*
V_{OS}	Input Offset Voltage	1mV	2.5mV (I grade: 3mV)
I_{OS}	Input Offset Current	10nA	25nA
GBW	Gain Bandwidth Product	10MHz	5MHz (I grade: 4.5MHz)
A_{OL}	Open-Loop Gain	1000V/mV	150V/mV
PSRR	Power Supply Rejection Ratio	90dB	80dB
CMRR	Common Mode Rejection Ratio	96dB	88dB
THD	Total Harmonic Distortion, $f = 1\text{kHz}, A_V = 1$	0.001%	—
V_S	Supply Voltage Range	2.5V to 12.6V	2.7V to 12V
I_S	Supply Current, $V_S = 3V$	1.1mA	2.0mA
P_N	Noise Power, $f = 10\text{kHz} (e_N \cdot i_N)$	$4.6 \times 10^{-21}\text{W/Hz}$	—
V_{OH}	Output Voltage Swing High, $V_S = 3V, I_{OUT} = 5\text{mA}$	2.8V	2.7V
V_{OL}	Output Voltage Swing Low, $V_S = 3V, I_{SINK} = 5\text{mA}$	80mV	100mV

*C grade: 0°C to 70°C; I grade: -45°C to 85°C

tor that reduces Q1's power dissipation at higher currents. The LT1797, the Zener diode and the transistor are both SOT-23 packages. Q1 was chosen for its high breakdown voltage and high β at low currents. Settling

time of better than 2 μ s to 1% was measured on a 1V output step. Total error is shared approximately equally between the 1% resistors, the transistor β and the LT1797 input offset voltage.

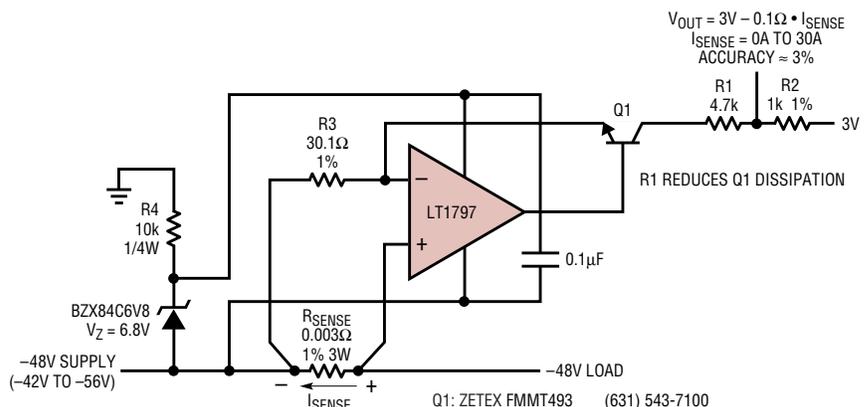
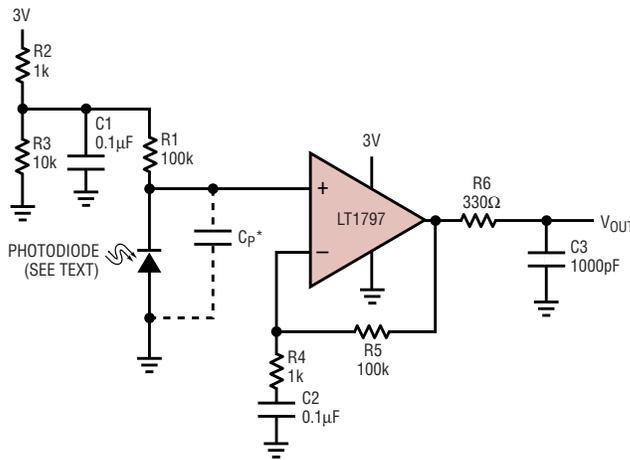


Figure 1. Fast, compact -48V current sense



*C_p = SUM OF PHOTODIODE CAPACITANCE, PARASITIC LAYOUT CAPACITANCE AND LT1797 INPUT CAPACITANCE

Figure 2. Simple, high gain 65kHz photodiode amplifier

Long-Range Photodiode Receiver

Much has been written about photodiode amplifiers, but relatively little has been published showing real circuits with achieved results. The circuit of Figure 2 is a simple photodiode amplifier that was optimized for range (or, if you prefer, sensitivity) with about 65kHz of bandwidth¹. The keys to achieving range are high gain and low noise. The lowest noise I/V converter is a resistor, so the simplest thing to do is to place a resistor in series with the photodiode (R1 of Figure 2). The problem is that putting a resistor in series with the photodiode reduces the bandwidth, but is infinite bandwidth needed? Maximizing this series resistance improves the inherent signal to noise ratio because the signal is

proportional to R1, whereas the resistor noise is proportional to $\sqrt{R1}$. R1 was chosen at 100kΩ to support 100kHz of bandwidth with 16pF² of combined photodiode and parasitic capacitance ($f_{-3dB} = 1/2\pi RC_p$). This achieves 100kΩ of transimpedance gain before even getting to the amplifier (pretty good!). R2, R3 and C1 bias the photodiode near the upper rail, reducing its capacitance while decoupling it from power supply noise and keeping the AC operation well ground referenced. Biasing the LT1797 near its upper rail also has the effect of favoring its NPN input stage, which improves the data sheet's 0.23pA/√Hz current noise specification by a factor of about $\sqrt{3}$. With a large input resistance of 100kΩ, the

current noise of the LT1797 would otherwise begin to dominate the circuit's noise performance. (It will be shown later that amplifier noise is not the limiting factor in many applications, anyway.)

R4, R5 and C2 place the LT1797 in an AC gain of 101. This high gain is not a problem for the LT1797 because of its high open-loop gain, and simply has the effect of reducing the bandwidth to 100kHz. R4 and C2 roll off the gain below 1.6kHz. This cancels the effect of 1/f noise and prevents amplification of the total DC offset, which by this point is several millivolts due to the 50nA bias current and the 100kΩ source impedance. R6 and C3 set a reliable upper limit on noise bandwidth at 500kHz, with minimal effect on signal bandwidth. When all is said and done, this single stage gives a total transimpedance gain of 10MΩ, down 3dB at 65kHz.³

The total output noise was measured with the circuit battery-powered in a sealed cookie tin with the low capacitance SFH213 photodiode installed. Using a Hewlett Packard 3403C true RMS meter, the output noise was 1.84mV_{RMS}. This result is in approximate agreement with the calculated RMS sum of the total noise sources: voltage noise, current noise $\times R_s$, and R_s Johnson noise.

$$V_{NOISE_RMS} = gain \times skirt_factor \times \sqrt{BW} \times \sqrt{(e_n^2 + (i_n \times R_s)^2 + 1.7e-20 \times R_s)}$$

where gain = 101, skirt factor = 1.3 and BW = 80kHz⁴, $e_n = 20nV/\sqrt{Hz}$, $i_n = 0.14pA/\sqrt{Hz}$, and $R_s = 101k\Omega$.

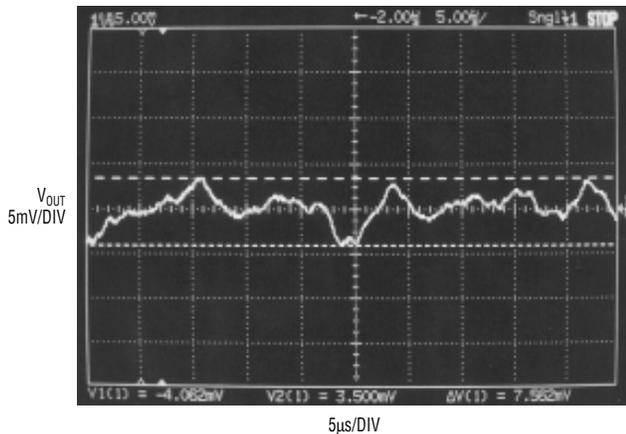


Figure 3. Total amplifier output noise is 7.6mV_{p.p}. The upper limit on measurement bandwidth is dictated by op amp gain bandwidth and R6 \times C3. Lower limit is 20kHz, set by the 50µs time window.

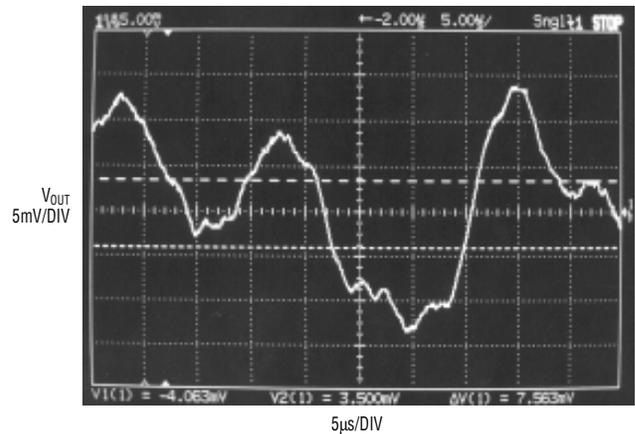


Figure 4. The same circuit exposed indirectly to fluorescent light. Cursors show amplifier noise measurement for comparison. The photodiode used is the least susceptible to this problem.

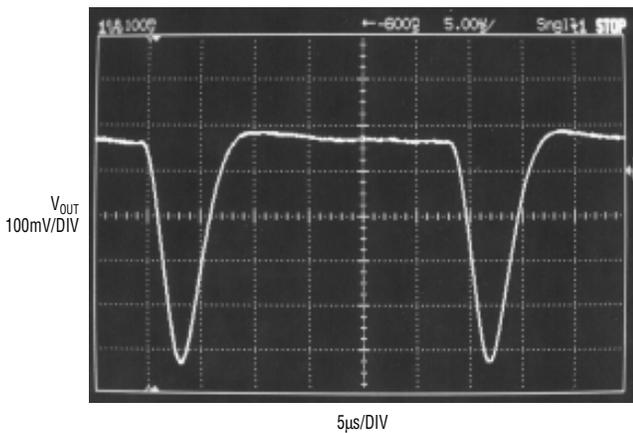


Figure 5. Reception at 16 feet using the SFH213FA. The transmitter is an HSDL-4220 IR LED pulsed with 330mA for 3.5 μ s.

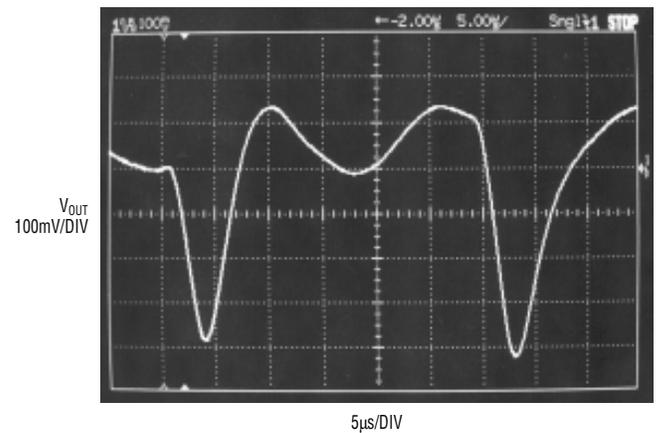


Figure 6. The same conditions as Figure 5, but without an IR filter. The photodiode is an SFH213.

This comes out to 1.8mV_{RMS}, very close to the 1.84mV_{RMS} measured. Figure 3 shows the peak-to-peak output noise at 7.6mV_{P-P}⁵, about four times the RMS value. This output voltage noise represents a total input-referred current noise, and hence resolution, of 0.76nA_{P-P}. Note that the method of calculating and/or measuring total output noise and then converting to an equivalent input current noise is more realistic than approaches that place undue initial emphasis on low amplifier input current noise. Placing undue emphasis on amplifier input current noise leads one to believe that JFET input amplifiers are the only option.

Although it is possible, and maybe even easy, to improve upon the noise figures achieved, in many through-air transmission applications it would be pointless. This is illustrated in

Figure 4, where the circuit has been exposed to light (with the least susceptible photodiode, an SFH213FA). The disgusting results are due to the fluorescent lights in the laboratory; their interference swamps any concerns about amplifier input noise. The problem appears to be a combination of two effects; visible light getting past the imperfect “black plastic” IR filter and the presence of actual IR or near-IR emissions from the lamps. Some lamps are worse than others, with efficient “warmer” lamps being the big culprits. Know your environment.

Overall results with three different photodiodes are shown in Figures 5 through 7. The transmitter is an Agilent HSDL-4220 IR LED placed sixteen feet away, with a drive current of 330mA for 3.5 μ s using a 10% duty cycle. Figures 5 and 6 show the

LT1797 receiver circuit output with the fast, narrow-angle Siemens/Infineon SFH213FA and SFH213 PIN photodiodes. The only difference between these devices is that the FA version has a black plastic filter that eliminates much of the visible fluorescent interference (the 50kHz ripple apparent in Figure 6), while only slightly attenuating the signal. Figure 7 shows results using a wide angle Vishay/Telefunken BPV22NF PIN photodiode. Signal amplitude is reduced not because the BPV22NF takes less optical gain, as this is compensated by its seven times increase in die area over the high optical gain SFH213 types. The reduction in amplitude is primarily due to the bandwidth reduction caused by the factor of seven increase in capacitance. The presence of the fluorescent interference in the case of the BPV22NF is not due to the lack of a visible light filter: it has one. Rather, it is due to its wide angle of acceptance. So, although the wider angle allows reception over a broader angle of transmitter locations, it also opens up its eyes to a lot more background interference through the (always imperfect) filter. At any rate, in none of the three cases is amplifier noise the dominating limit to achieving range.

The circuit works well even at close range because the LT1797 has good output clipping recovery. However, the amplitude of the photodiode current at close range can be fairly

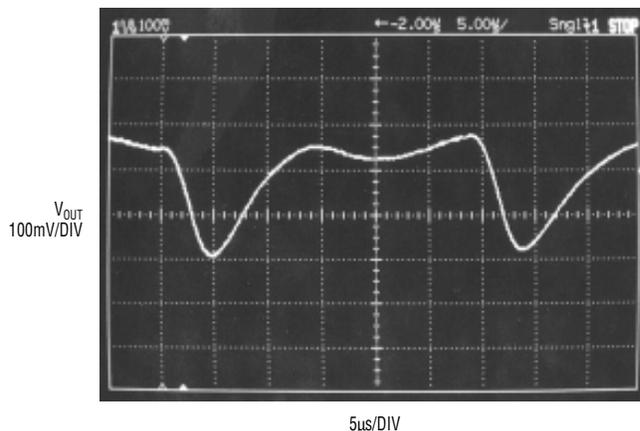


Figure 7. Same conditions as Figures 5 and 6, but using a wider-angle, larger-die device, the BPV22NF. The IR filter attempts to reject some of the fluorescent light; capacitance slows the response.

high, in which case the recovery time will rise and the output pulse width will increase. The higher capacitance BPV22NF shows this effect more than does the SFH213. This circuit is not suited to pulse width modulation schemes unless physical transmitter motion will be below the frequency of interest and the steady-state pulse width is noncritical.

Convert Your Favorite Op Amp to a Rail-to-Rail Output

Many of the world's greatest op amps were not originally intended for operation on reduced supply voltages, the ultralow noise LT1028 being a good example. The LT1797 can help remedy this situation by converting the output stage of one of these amplifiers to a rail-to-rail output stage. Figure 8 shows the method. The LT1028 output drives the noninverting input of the LT1797, which is placed in a gain of three by R1 and R2. The feedback resistors R3 and R4 put the entire loop in a gain of 500, forcing the LT1028 to provide a gain of 167. This combination of the two amplifiers takes advantage of the ultralow noise, precision front end of the LT1028 and the rail-to-rail output of the LT1797. The circuit is stable from a gain-phase point of view without compensation components R5 and C1. However, when the input

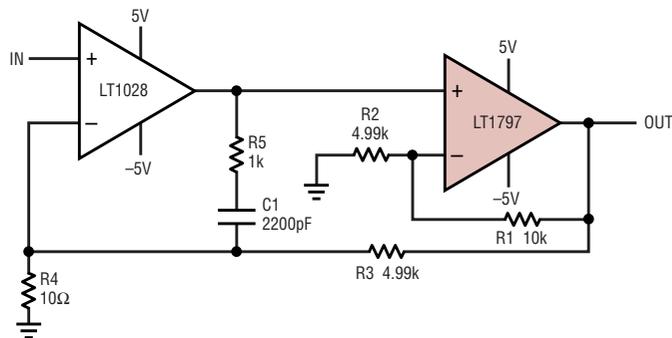


Figure 8. Converting the LT1028 to a $\pm 5\text{V}$ supply with rail-to-rail output; $A_v = 500$

receives a transient or the output hits a rail, the two op amps begin a usually unrecoverable slew-rate contest. R5 and C1 fix this by slowing down the LT1028.

Conclusion

The LT1797 is a compelling choice where minimal footprint or rail-to-rail 10MHz gain bandwidth are essential. The efficient nature of the LT1797 design also makes it suitable for applications where power is at a premium and wide bandwidth and output drive are also required. 

Notes:

- ¹ To cut to the chase, results will be given with sixteen feet of transmitter-receiver separation.
- ² Some of the photodiodes tested had more capacitance than this, and some had less. Although it is tempting to place a trimpot at R1, the parasitic

capacitance of a bulky trimpot would quickly complicate the matter.

- ³ Cascading the two 100kHz -3dB bandwidths results in a net bandwidth of 65kHz. However, the -3dB that is due to the photodiode capacitance and R1 will be more or less dependent on the photodiode used, and this will have an effect on the net bandwidth.

- ⁴ The bandwidth is chosen at about 80kHz because the low capacitance photodiode will not reduce the 100kHz bandwidth as much as would the design value of 16pF. For additional complexity, the bandwidth reduction due to input capacitance has effect on current noise and Johnson noise but not voltage noise. Also, the fact that measurements are made over a finite period of time introduces an inherent highpass characteristic. The skirt factor is next to impossible to determine because of the complexity of the various roll-off mechanisms. The value of 1.3 is a compromise.

- ⁵ Taking 100 measurements using a 50 μs window, the average peak-to-peak noise was 7.7mV_{p-p} with a standard deviation of 1.2mV_{p-p}. Note that a 50 μs window has a highpass effect above about 15kHz.