OPTOELECTRONICS CIRCUIT COLLECTION

By Neil Albaugh

The following collection of analog circuits may be useful in electro-optics applications such as optical networking systems. This page summarizes their salient characteristics.

AVALANCHE PHOTODIODE BIAS SUPPLY 1

Provides an output voltage of 0V to +80V for reverse biasing an avalanche photodiode to control its gain. This circuit can also be reconfigured to supply a 0V to –80V output.

LINEAR TEC DRIVER–1

This is a bridge-tied load (BTL) linear amplifier for driving a thermoelectric cooler (TEC). It operates on a single +5V supply and can drive ±2A into a common TEC.

LINEAR TEC DRIVER–2

This is very similar to DRIVER–1 but its power output stage was modified to operate from a single +3.3V supply in order to increase its efficiency. Driving this amplifier from a standard +2.5V referenced signal causes the output transistors to have unequal power dissipation.

LINEAR TEC DRIVER–3

This BTL TEC driver power output stage achieves very high efficiency by swinging very close to its supply rails, ±2.5V. This driver can also drive ±2A into a common TEC. Operation is shown with the power output stage operating on ±1.5V supplies. Under these conditions, this linear amplifier can achieve very high efficiency.

LASER DIODE DRIVER–1

A single-ended voltage–controlled current source is shown here. This circuit operates from a single +3.3V supply and it can drive from 0A to 2A into a laser diode with a 0V to 2V input from a Digital-to-Analog (D/A) converter. Its input can be also be configured for a 0V to –2V input voltage.

LASER DIODE DRIVER–2

Similar to the previous circuit except that it operates from ±5V supplies and it is inverting; i.e., drives from 0A to 2A into a laser diode with a 0V to –2V input. A very low noise bipolar input op amp allows this circuit to achieve a low noise output current, an important consideration in dense channel spacing systems. As drawn, neither terminal of the laser diode is grounded.

TEMPERATURE UNDER–AND OVER–RANGE SENSING WITH A WINDOW COMPARATOR

This circuit is useful to monitor the temperature of a TEC thermistor, to sense an out-of-range condition, or if the thermistor is shorted or open. Current sources make the threshold adjustments non-iterative. The supply voltage is +5V.

AVALANCHE PHOTODIODE BIAS SUPPLY–1

An avalanche photodiode (APD) is commonly used in optical detector circuits that require high sensitivity and wide bandwidth. A high reverse bias voltage across the photodiode junction creates avalanche gain, and varying the reverse bias voltage can control this gain. Although some APDs require a bias of a few hundred volts, many InGaAs and Si APDs require only 60V to 80V.

The circuit shown in Figure 1 can provide a positive bias voltage of up to +80V to an APD. The 0V to +2V input control voltage can be a DAC output or from an analog source.

The OPA445 high-voltage op amp is rated to operate with up to ±45V supplies and it can provide up to ±15mA if its power dissipation limits are observed. To obtain a high positive output voltage from this op amp, it can be operated from unequal supplies as long as the voltage difference between the supplies is 90V or less and the amplifier's common-mode input voltage stays within its specified range.

To allow the OPA445's output voltage to swing to zero, a negative supply of –5V was chosen. Staying within the 90V supply voltage difference specification, a +85V positive supply was chosen. This allows the OPA445 output voltage to swing up to +80V.

R_{IN} and R_F set the gain of this noninverting op amp circuit to 40V/V. As illustrated in Figure 2 an input of 0V to +2V results in an output of 0V to +80V.

Other gains can be calculated by the equation:

$$Av = \left(\frac{R_F}{R_{IN}}\right) + 1$$

Other op amps can be used in this circuit if lower output voltages are desired. An OPA551 or OPA552 can be used for a 0V to +50V bias supply if its positive supply voltage is reduced to +55V.

Bandwidth of this circuit is about 60kHz, as seen in Figure 3. To reduce noise, a capacitor can be placed in parallel with R_F. In addition, low-pass filtering can be accomplished by adding a passive RC low-pass filter (LPF) on the DAC output, the OPA445 output, or both. Low noise on the bias supply is important, noise on an APD bias supply will feed through into the following stage–usually a transimpedance amplifier.

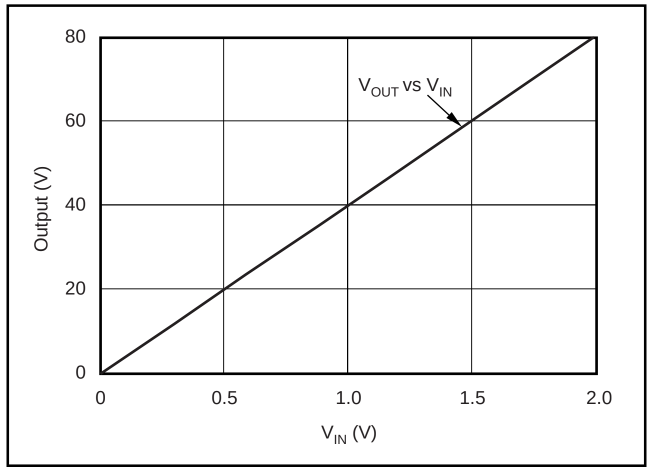

FIGURE 2. APD Bias Supply DC Output versus Input Voltage.

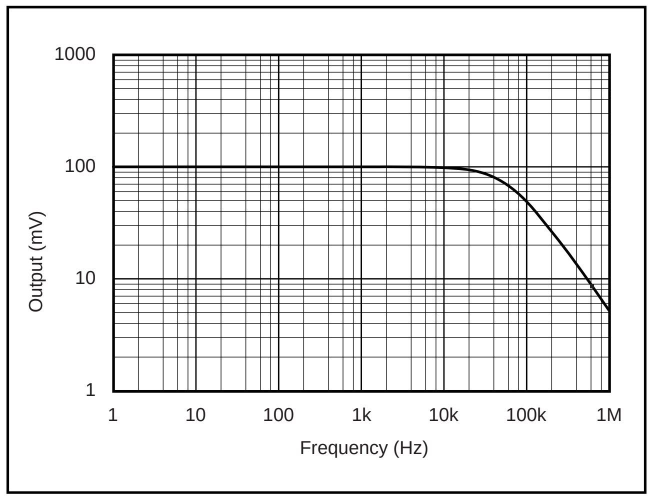

FIGURE 3. APD Bias Supply Small Signal Bandwidth.

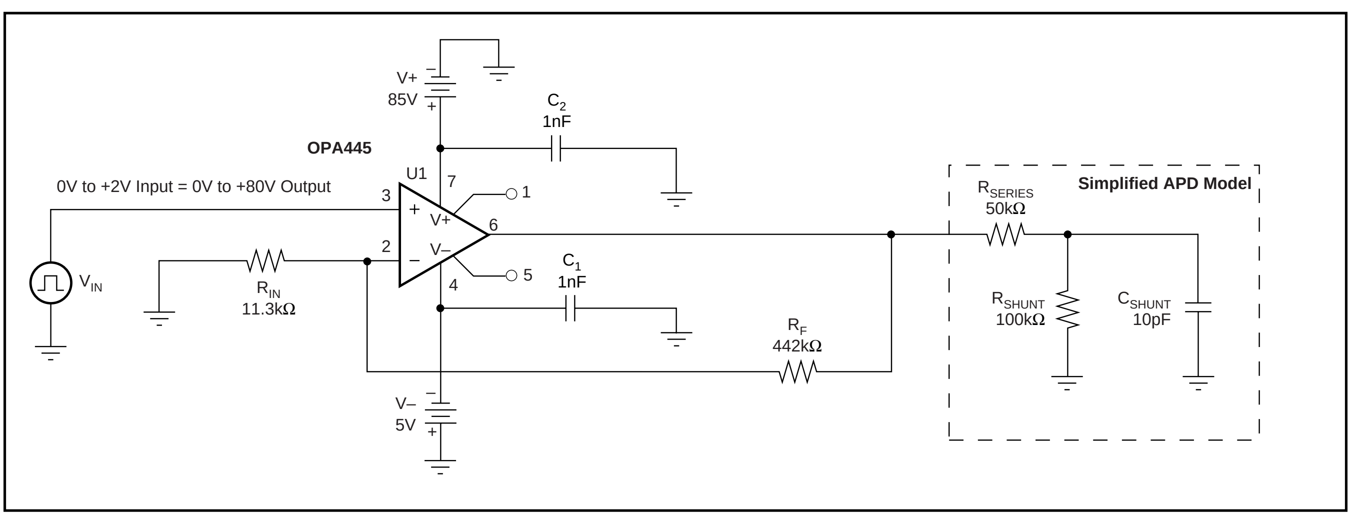

FIGURE 1. Positive Bias Supply Circuit Diagram.

The circuit's transient response in Figure 4 shows a clean step response with no peaking or overshoot.

The OPA445 is now offered in a SO-8 surface–mount package, so adding a programmable APD bias supply into a corner of a larger PCB layout is now feasible.

This bias supply circuit can easily be reconfigured to supply a 0V to –80V output by changing the op amp's supplies to +5V and –85V. A 0V to –2V input is required, but a 0V to +2V input can be used if the input is connected to R_{IN} and the op amp noninverting input is grounded. This changes the op amp to an inverting configuration. Note that the circuit's input impedance is now lower (equal to R_{IN}) if the OPA445 is used as an inverter. The familiar inverting op amp gain equation is:

$$Av = \left(-\frac{R_F}{R_{IN}}\right)$$

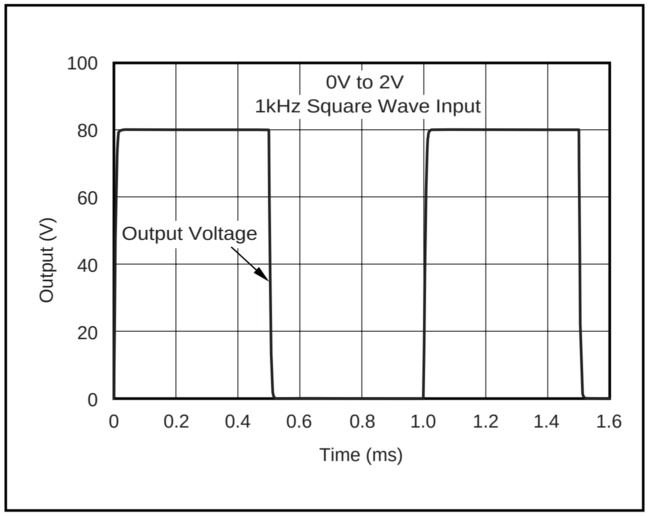

FIGURE 4. APD Bias Supply Transient Response.

LASER DIODE DRIVER–1

The voltage-controlled current source circuit shown in Figure 5 can be used to drive a constant current into a signal or pump laser diode. This simple linear driver provides a cleaner drive current into a laser diode than switching PWM drivers.

The basic circuit is that of a Howland current pump with a current booster (Q_1) on the output of a R-R CMOS OPA350 op amp (U_2). Laser diode current is sensed by differentially measuring the voltage drop across a shunt resistor (R_{SHUNT}) in series with the laser diode. The output current is controlled by the input voltage (V_{IN}) that may be from an analog voltage source or from a voltage-output DAC. As shown, the scale factor is 1V input equals 1A output.

The scale factor (V_{IN}/I_{OUT}) can be set to other values by choosing appropriate resistor values using the equation:

$$\frac{V_{IN}}{I_{OUT}} = \frac{R_3}{R_4} \bullet R_{SHUNT} \text{ and, } R_1 = R_3,\ R_2 = R_4$$

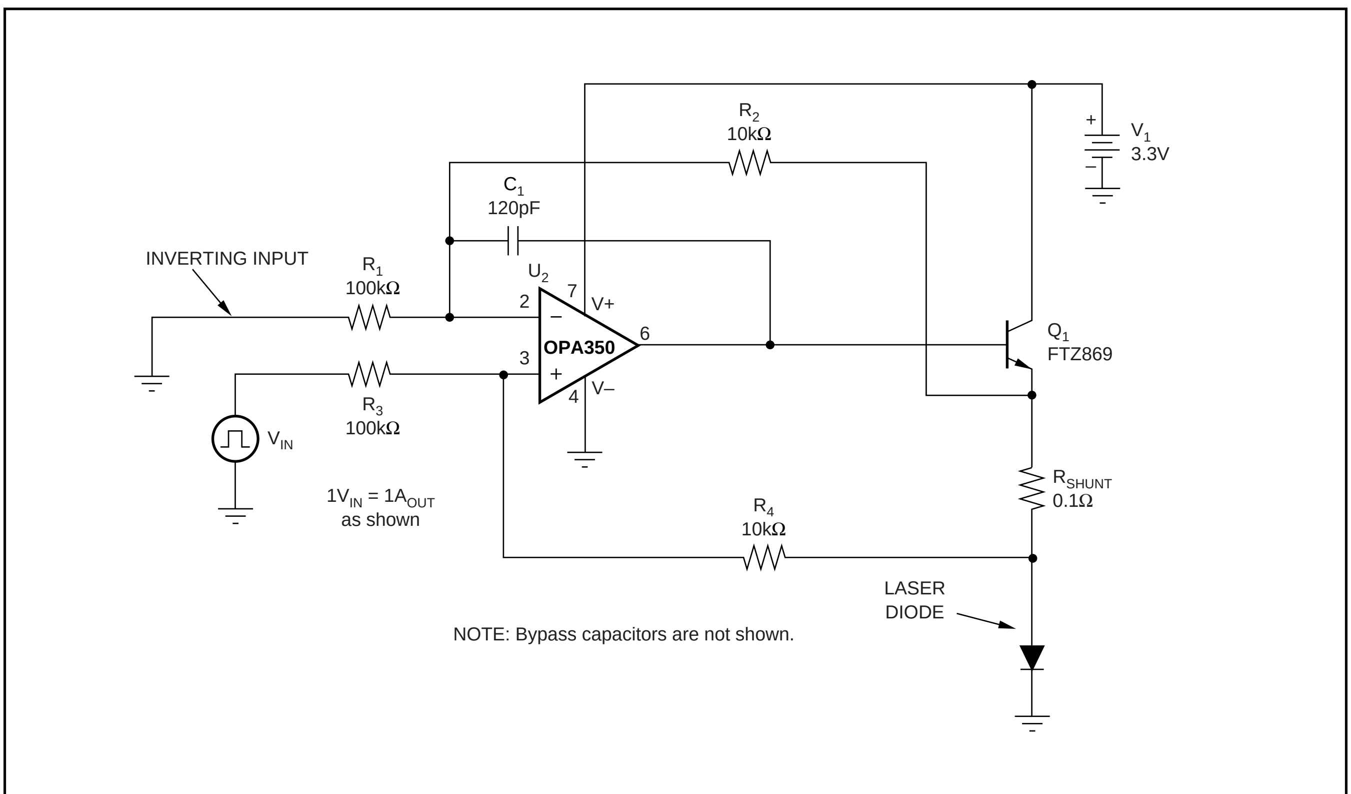

FIGURE 5. Laser Diode Constant–Current Driver–1, Circuit Diagram.

A P–Spice simulation (DC sweep) was performed on V_{IN}, sweeping the input voltage from 0V to 2V. The lower curve, shown in Figure 6, shows the relationship of the laser diode current to the input voltage. Power dissipation of Q_1 is shown in the upper curve.

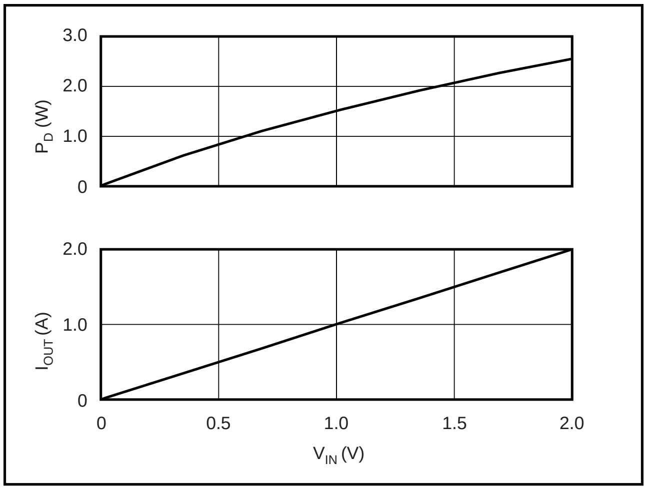

FIGURE 6. Output Current and Q_1 Power Dissipation versus Input Voltage with 3.3V Supply.

Operating on a supply voltage of 3.3V, Q_1 dissipates only 1.5W at an output current of 1A. This is well within the capability of the FTZ transistor, as its SOT-223 package can dissipate heat into the copper traces on a PC board.

Similar curves are shown in Figure 7 for operation on a 5V supply voltage. This clearly shows the power advantage in operating the current source on a low supply voltage. The higher 5V supply voltage is advantageous if a higher output compliance voltage is required. A laser diode macromodel was unavailable, so a laser diode junction was simulated by a series connection of three silicon diodes.

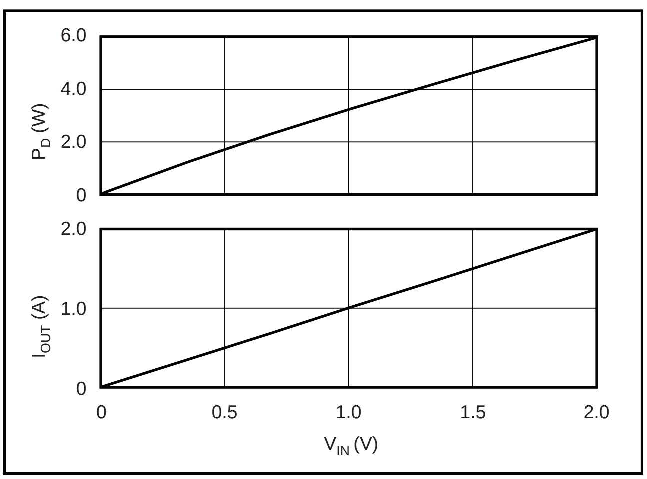

FIGURE 7. Output Current and Q_1 Power Dissipation versus Input Voltage with 5V Supply.

The P–Spice Probe output of a transient response simulation is shown in Figure 8. The input voltage pulse amplitude was stepped from 10mV to 500mV to 10mV to 2V, the current output waveform was plotted. The circuit displays clean response on both the rising and falling edges of the pulse. Figure 9 also shows clean pulse response operating on a 5V supply.

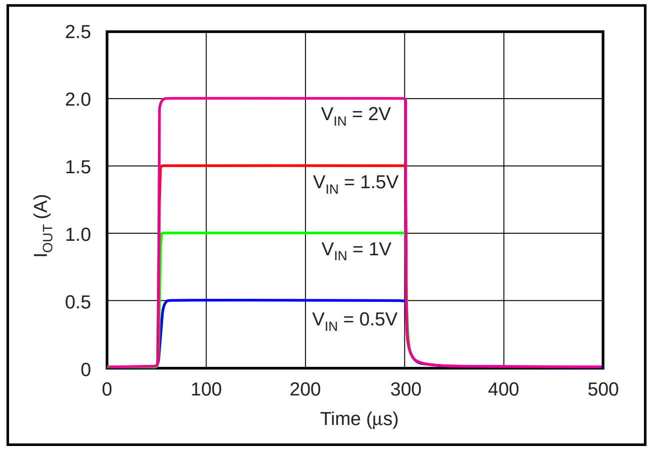

FIGURE 8. Laser Driver Transient Response with +3.3V Supply.

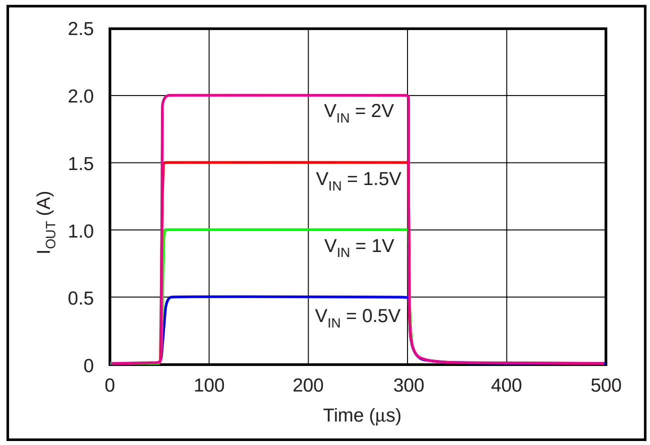

FIGURE 9. Laser Driver Transient Response with +5V Supply.

Changing the value of the shunt resistor or the scale factor will necessitate changing the compensation capacitor C_1. Verify that your circuit is stable before connecting an expensive laser diode to the output.

If a negative output control voltage is available, it can be applied to R_1 and R_3 is then tied to ground. The amplifier is then configured as an inverting amplifier.

The power booster used for Q_1 is a very high gain single NPN transistor, and is not a Darlington, it has a Beta of over 300 at a collector current of 1A, allowing the CMOS OPA350 op amp to easily drive it to high currents. Zetex rates its continuous collector current as 6.5A but SOA limits are reached before approaching this current.

If a unidirectional output current is acceptable, this circuit can be used to drive a TEC for cooling a laser diode or an APD. Adding a mechanical switch or a low on-resistance H-bridge will allow the TEC polarity to be switched and changed from heating to cooling.

Satisfactory operation of this circuit should be verified in your actual application by breadboarding and testing.

LASER DIODE DRIVER–2

The voltage-controlled current source circuit shown in Figure 10 can be used to drive a very low-noise constant current into a signal or pump laser diode. This simple linear driver provides a far cleaner drive current into a laser diode than a switching PWM driver can achieve.

The basic circuit is that of an NPN transistor current booster (Q_1) on the output of (U_1), a very low noise bipolar op amp OPA227. Laser diode current is sensed by measuring the voltage drop across a shunt resistor (R_{SHUNT}) in the emitter of Q_1. The output (laser) current is controlled by the input voltage (V_{IN}) that may be from an analog voltage source or from a voltage-output DAC. As shown, the scale factor is –1V input equals 1A output.

The scale factor (V_{IN}/I_{OUT}) can be set to other values by choosing appropriate resistor values using the equation:

$$\frac{V_{IN}}{I_{OUT}} = \frac{R_1}{R_2} \bullet R_{SHUNT}$$

A P–Spice simulation (DC sweep) was performed on V_{IN}, sweeping the input voltage from 0V to –2V. The lower curve shown in Figure 11 shows the relationship of the laser diode current to the input voltage. Power dissipation of Q_1 is shown in the upper curve. Operating on a supply voltage of 3.3V, Q_1 dissipates only 1.5W at an output current of 1A. This is well within the capability of the FTZ851 transistor, as its SOT-223 package can dissipate heat into the copper traces on a PC board.

A similar DC sweep output current (I_{OUT}) versus V_{IN} curve is obtained with Q_1 operating on a 5V supply. The higher 5V supply voltage is advantageous if a higher output compliance voltage is required. Op amp U_1 supplies can be from ±5V to ±15V.

See Figure 12 for the P–Spice Probe output of a transient response simulation. The input voltage pulse amplitude was stepped from –10mV to –500mV to –10mV to –2V, the current output waveform was plotted. The circuit displays clean response on both the rising and falling edges of the pulse. The circuit simulation also exhibited clean pulse response with Q_1 operating on a 5V supply.

Changing the value of the shunt resistor or the scale factor will necessitate changing the compensation capacitor C_1. Verify that your circuit is stable before connecting an expensive laser diode to the output.

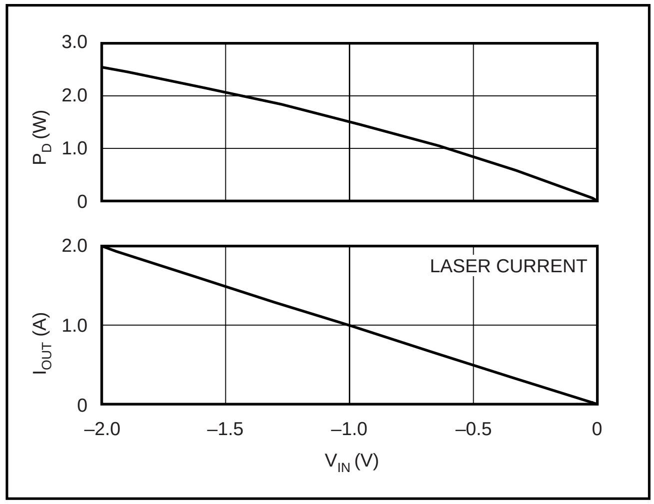

FIGURE 11. Output Current and Q_1 Power Dissipation Versus Input Voltage with 3.3V Supply.

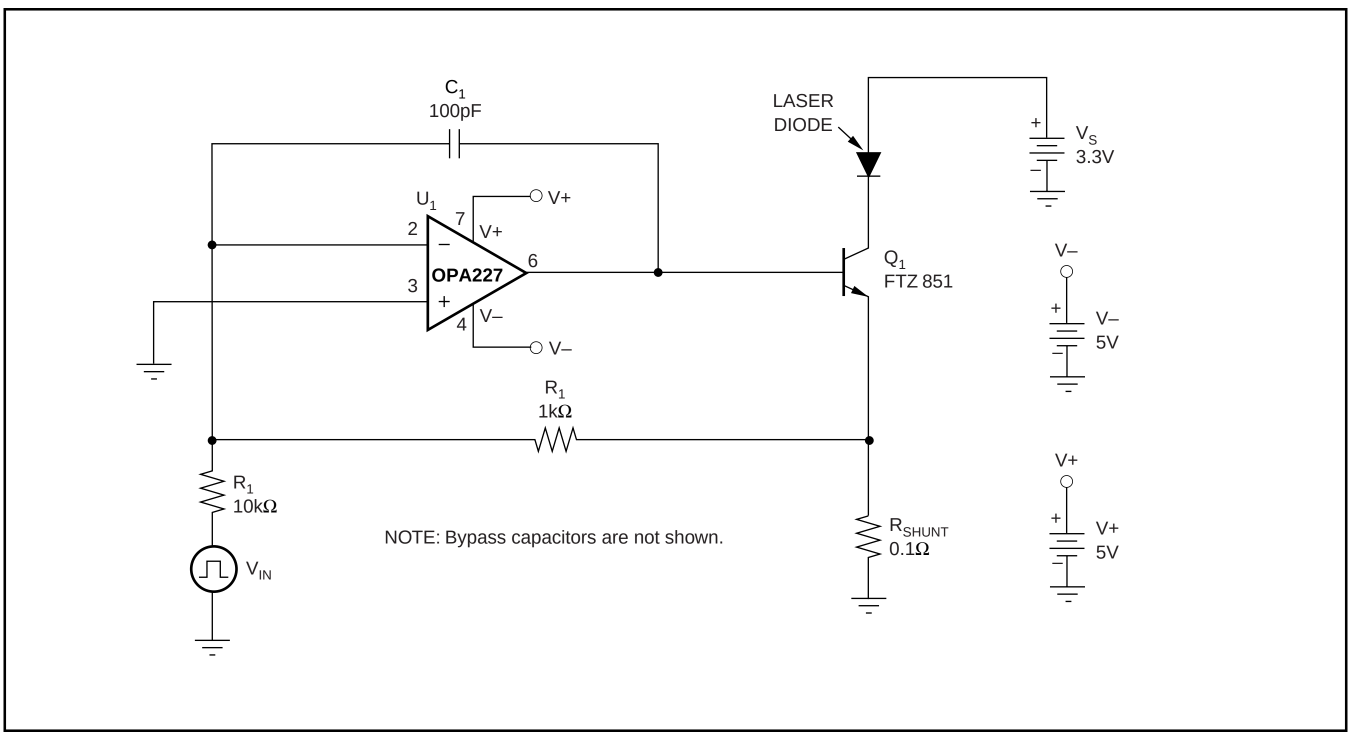

FIGURE 10. Laser Diode Constant-Current Driver–2 Circuit Diagram.

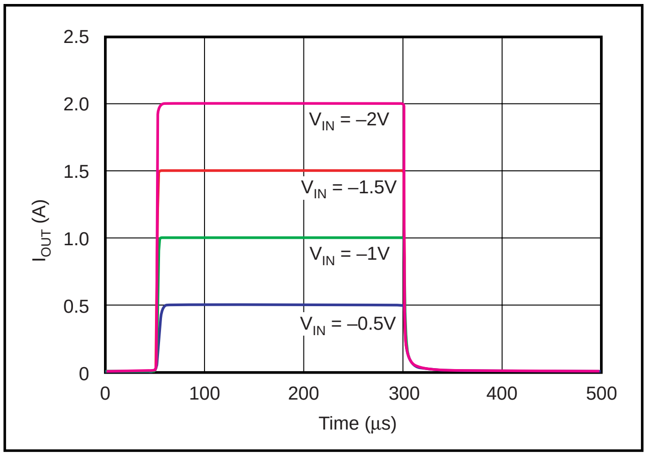

FIGURE 12. Laser Driver–2 Transient Response with +3.3V Supply.

A laser diode macromodel was unavailable so a laser diode junction was simulated by a series connection of three silicon diodes. This driver circuit requires both the anode and cathode of the laser diode to be floating.

The power booster used for Q_1 is a very high gain single NPN transistor, and is not a Darlington, it has a Beta of over 100 at a collector current of 2A, allowing the OPA227 op amp to easily drive it to high currents. Zetex rates its continuous collector current as 6A, but SOA limits are reached before approaching this current.

Satisfactory operation of this circuit should be verified in your actual application by breadboarding and testing.

LINEAR TEC DRIVER–1

The linear thermoelectric cooler (TEC) driver circuit is capable of driving ±2A into a TEC (see Figure 15). The circuit operates on a single +5V supply and drives the TEC in the highly desirable "constant-current" mode. A bridge-tied load (BTL) amplifier topology achieves bidirectional current output. This type of amplifier drives its load differentially, so the TEC must not be grounded on either end. An input offset of 1/2 supply voltage (in this case, an offset of ±2.5V) is used to allow the amplifiers to swing in both directions and to interface with a single-supply input voltage source. This is represented in the circuit by V_{OS}.

In Figure 15, voltage V_{IN} is amplified by a R-R CMOS op amp U_1, with a class B power output stage (formed by the addition of a complimentary power transistor pair Q_1 and Q_3) that drives one end of a TEC load through shunt resistor R_4. A CMOS instrumentation amplifier (IA) U_3 senses the voltage drop across the shunt resistor and amplifies it by 50, it then feeds it back to the input of U_1. This feedback approach forces the output TEC current to be a function of V_{IN}. Shunt resistance and IA gain determines the scale factor of the circuit.

$$\frac{V_{IN}}{I_{OUT}} = A_V \bullet R_4 \quad \text{where } A_V \text{ is the IA gain in V/V}$$

A P-Spice simulation (DC sweep) was performed on V_{IN}, sweeping the input voltage from –2.5V to +2.5V. This is equivalent to an input voltage of 0V to +5V from an external voltage source.

The P-Spice Probe output current to the TEC is shown in Figure 13. TEC current is shown for three different sizes; 1Ω, 1.5Ω, and a 2Ω TEC. When operating on a single +5V supply, this driver is capable of driving a 1Ω or 1.5Ω TEC to 2A. Output voltage compliance limits the 2Ω TEC current to about 1.6A.

As Figure 13 illustrates, this TEC driver amplifier is a voltage-controlled current source. Constant-current drive assures that TEC drive current is independent of production variations in TEC junctions or long-term aging. Constant-current drive also eliminates the effect of thermal "back EMF" on current through the TEC under dynamic temperature control conditions.

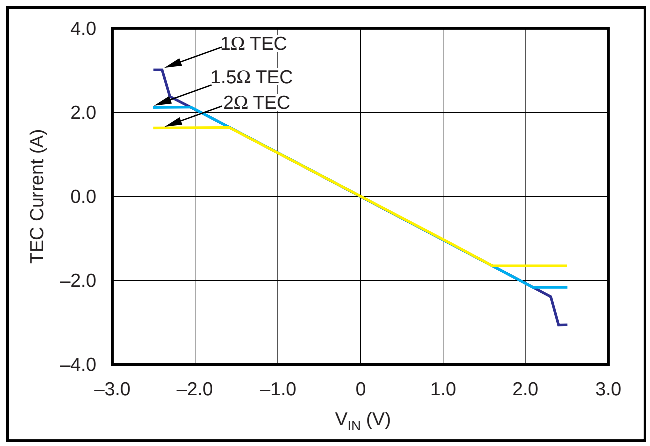

FIGURE 13. TEC Current Versus Input Voltage V_{IN}.

To determine the circuit's power dissipation and the requirements for heat sinking the SOT-223 output power transistors, a simulation was run by sweeping the DC input voltage as before, and using the same size TECs. The results are shown in Figure 14.

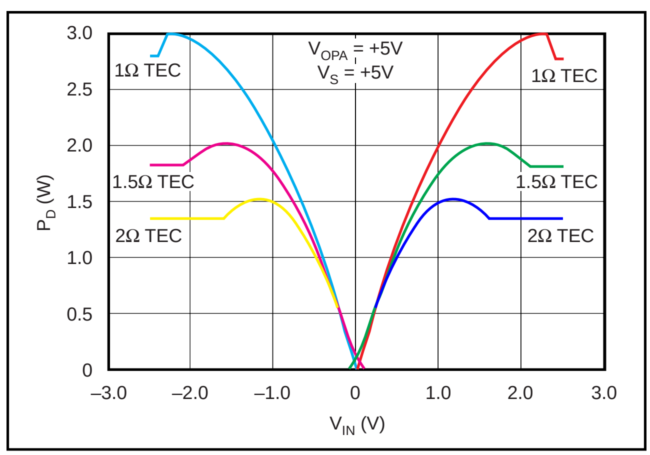

FIGURE 14. Output Transistor Power Dissipation.

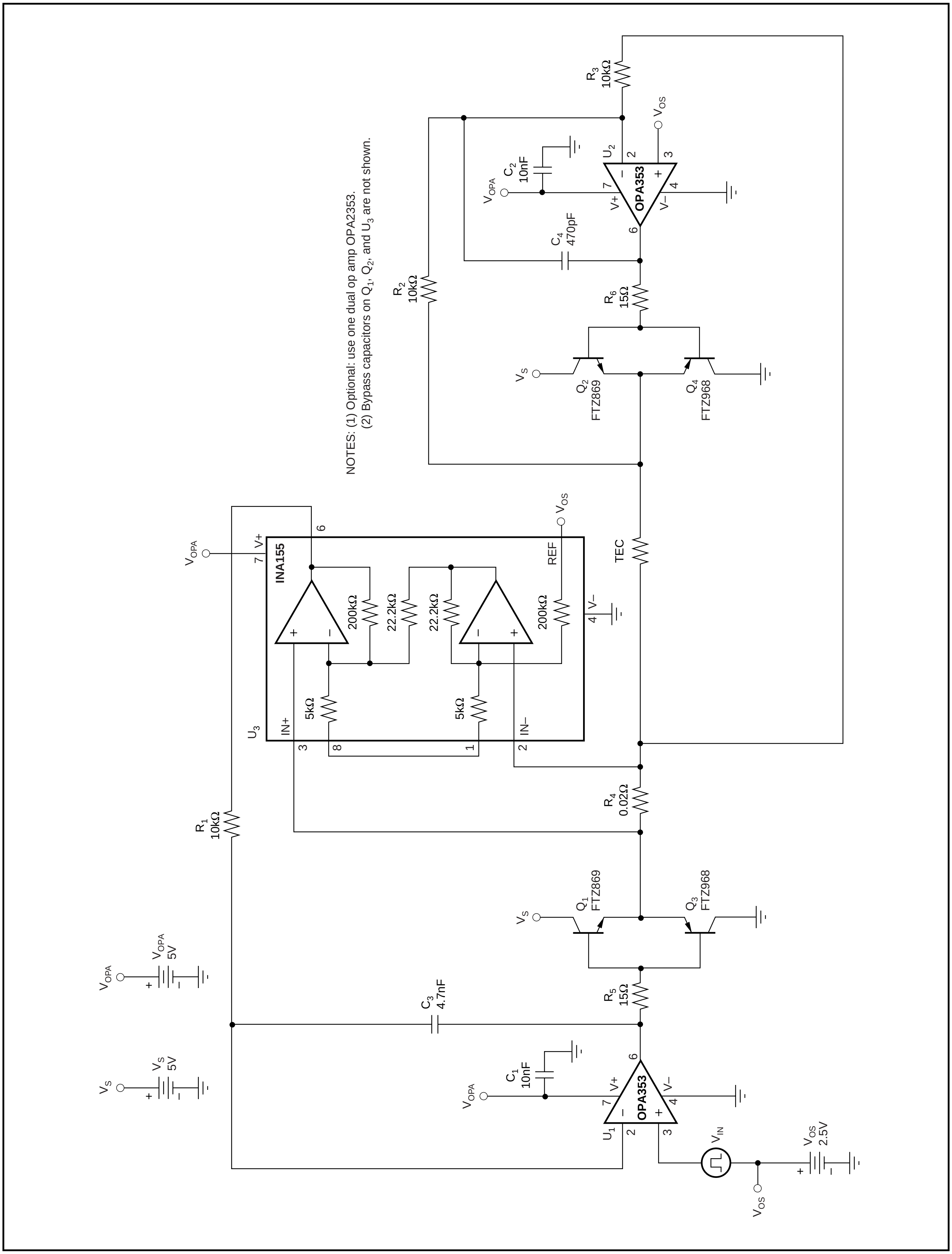

FIGURE 15. TEC Driver–1 Circuit Diagram

Power dissipation of one NPN Q_1 and one PNP Q_3 power output transistor is shown in this sweep. The dissipation of the devices in the other half of the bridge (NPN = Q_2 and PNP = Q_4) will be the same, as shown in Figure 16. Depending on whether the TEC is in its cooling or heating mode, power is dissipated in Q_1 and Q_4 or in Q_2 and Q_3.

Driver efficiency is usually a concern in large multi-channel systems due to limitations on the total power dissipation of the system. Linear amplifiers do not reach the efficiencies of PWM switching types, but they do offer important advantages, principally their very low noise. Switching noise interference in laser and APD circuits is not a concern with linear drivers.

The DC simulation data was used to plot the efficiency of the driver. To simplify the calculation, only the power output stage was considered. The CMOS OPA353 op amp power dissipation is only about 26mW, so deleting it contributes little error to the overall efficiency calculation. In this calculation, efficiency is considered to be the ratio of the power delivered to the load TEC to the power supplied to the driver. For example, a current of 1A into a 1Ω TEC represents a power P_{OUT} of 1W dissipated in the load ($P_{OUT} = I^2R$). The power supplied to the driver P_{IN} is 1A from the = 5V supply, or 5W ($P_{IN} = E \bullet I$).

Therefore:

$$\text{Eff}(\%) = \frac{P_{OUT}}{P_{IN}} \bullet 100,\ \text{or Eff}(\%) = \frac{I^2 \bullet R}{V_S \bullet I_S} \bullet 100 = \frac{1A^2 \bullet 1\Omega}{5V \bullet 1A} \bullet 100 = 20\%$$

Swinging the output voltage close to the supply rails minimizes the voltage across the output transistor, thus reducing its power dissipation. Likewise, it maximizes the voltage across the load. As can be seen from the equation above, this increases the efficiency of a linear driver. In fact, as seen in Figure 16, this circuit can reach an efficiency of over 60% under favorable conditions. The discussion of TEC Drivers 2 and 3 investigates this further.

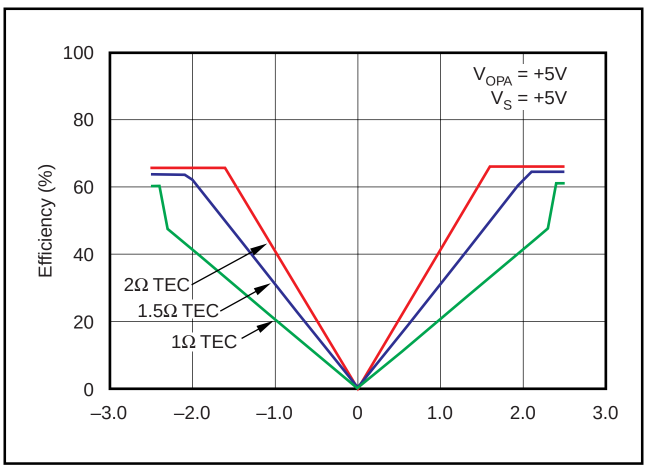

FIGURE 16. Driver Efficiency with 1Ω, 1.5Ω, and 2Ω Loads.

Driver–amplifier loop stability was investigated by running an AC and Transient simulation. Results of the AC simulation are shown in Figure 17.

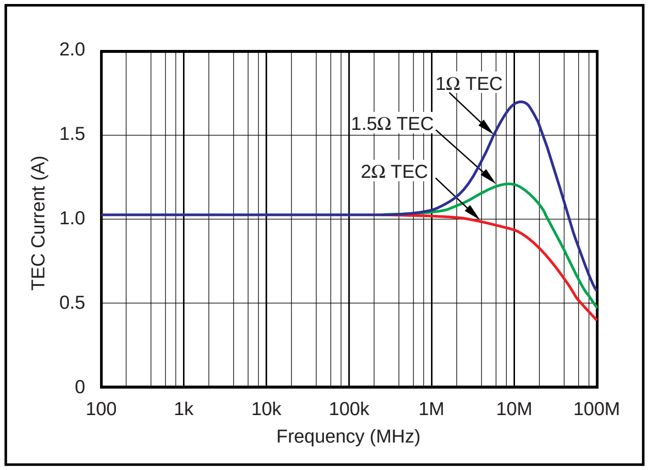

FIGURE 17. TEC Driver Amplifier Frequency Response.

Some peaking is shown in the frequency response curves of the 1Ω and 1.5Ω TEC but increasing the capacitance of the compensation capacitors C_3 and C_4 can eliminate this. Notice that the capacitance of C_3 and C_4 are not the same. Amplifier U_1 requires more capacitance than U_2 because of the presence of feedback gain provided by the instrumentation amplifier, U_3.

Similar results are seen in Figure 18, a P–Spice transient simulation of this TEC driver amplifier. Slight peaking is noted for heavy loads, as predicted by the frequency response curves.

Also evident in the transient response waveform is a small crossover distortion "glitch" around 0A output current due to the delay between turning off one transistor before its complimentary transistor turns on. For example, the op amp driving the power transistors must slew quickly between the voltage at which the NPN transistor turns off and the PNP transistor turns on. Due to the fact that this amplifier uses a Class B output stage, this region is $2V_{BE}$. A fast op amp, such as the OPA353, minimizes the crossover distortion and enhances stability about the crossover region. For applications such as an audio amplifier, the output stage transistors could be biased slightly into conduction (class AB1) which would eliminate the crossover region altogether. Applications for driving loads such as a thermoelectric cooler do not warrant the increased complexity of a Class AB1 output stage.

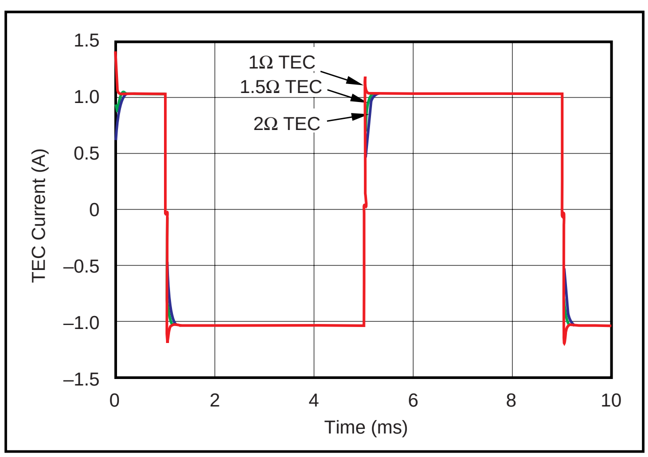

FIGURE 18. TEC Driver Transient Response.

LINEAR TEC DRIVER–2

The linear TEC driver circuit (see Figure 21) is capable of driving +1.5A and –1A into a TEC. The small-signal amplifier circuit operates on a single +5V supply and the power output stage operates on a single +3.3V supply to minimize its power dissipation. This circuit drives the TE cooler in the highly desirable "constant-current" mode. A BTL amplifier topology achieves bi-directional current output. This type of amplifier drives its load differentially, so the TEC must not be grounded on either end.

An input offset of 1/2 the op amp supply voltage V_{OPA} (in this case, an offset of +2.5V) is used to allow the amplifiers to swing in both directions and to interface with a single-supply input voltage source. This is represented in the circuit by V_{OS}.

In the schematic, voltage V_{IN} is amplified by a R-R CMOS op amp U_1 with a class B power output stage (formed by the addition of a complimentary power transistor pair Q_1 and Q_3) that drives one end of a TEC load through shunt resistor R_4. A CMOS INA155 instrumentation amplifier (IA) U_3 senses the voltage drop across the shunt resistor and amplifies it by 50; it then feeds it back to the input of U_1. This feedback approach forces the output TEC current to be a function of V_{IN}. Shunt resistance and IA gain determines the scale factor of the circuit.

$$\frac{V_{IN}}{I_{OUT}} = A_V \bullet R_4 \quad \text{where } A_V \text{ is the IA gain in V/V.}$$

A P–Spice simulation (DC sweep) was performed on V_{IN}, sweeping the input voltage from –2.5V to +2.5V. This is equivalent to an input voltage of 0V to +5V from an external voltage source.

The P–Spice Probe output current to the TEC is shown in Figure 19.

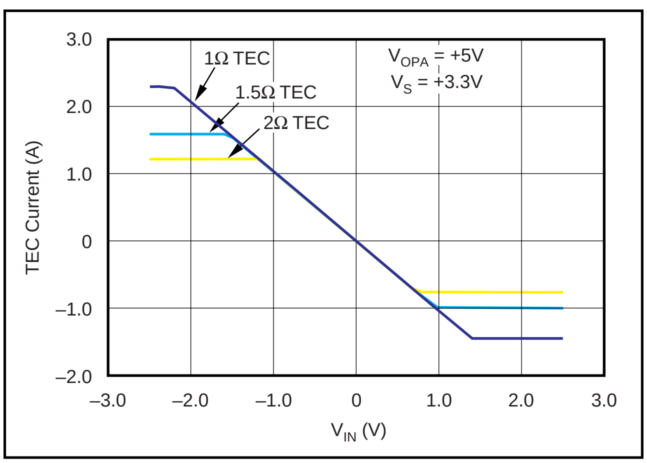

FIGURE 19. TEC Current Versus Input Voltage V_{IN}.

TEC current is shown for three different sizes; 1Ω, 1.5Ω, and a 2Ω TEC. With the output transistors operating on a single +3.3V supply, this driver is capable of driving a 1Ω TEC to over 2A or 2Ω TEC to over 1A in one direction but its output current capability is lower in the opposite direction. Output voltage compliance limits are asymmetrical due to the $1/2V_{OPA}$ offset voltage applied to the op amp input.

While this is the optimum offset to allow the op amp to swing symmetrically to its supply rails, it creates a problem for the output transistors. To swing optimally to its rails, it would be biased to $1/2V_S$ to half its supply of only +3.3V or an offset of 1.65V. The 2.5V offset that is used creates an unbalance in the output transistor bridge, causing some problems as we shall see.

As Figure 19 illustrates, this TEC driver amplifier is a voltage-controlled current source. Constant-current drive assures that TEC drive current is independent of production variations in TEC junctions or long-term aging. Constant-current drive also eliminates the effect of thermal "back EMF" on current through the TEC under dynamic temperature control conditions.

To determine the circuit's power dissipation and the requirements for heat sinking the SOT-223 output power transistors, a simulation was run by sweeping the DC input voltage as before, using the same sizes of TECs. The results are shown in Figure 20.

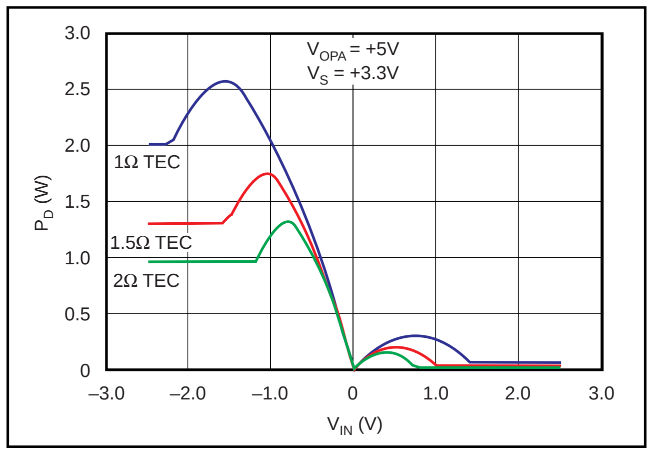

FIGURE 20. Output Transistor Q_1 and Q_3 Power Dissipation.

Power dissipation of one NPN (Q_1) and one PNP (Q_3) power output transistor is shown in this sweep. Note the difference in power dissipation between the NPN and PNP transistor. This is a direct result of the unbalance caused by the offset voltage problem previously noted. The dissipation of the devices in the other half of the bridge (NPN = Q_2 and PNP = Q_4) are similar but a "mirror image" as seen in Figure 4. Depending on whether the TEC is in its cooling or heating mode, power is primarily dissipated in the PNP transistor Q_3 or in Q_4. The power dissipated in NPNs, Q_1 and Q_2, is relatively minor.

The effects of limited output compliance voltage are also clearly seen in the curve of Figure 22.

This compliance limit can also be seen in the power transistor output voltage driving the each end of the TEC load (see Figure 23). Note that the amplifier's output voltage polarity crossover point is at the 2.5V input bias point. Biasing the output transistors at their optimum 1.65V point would allow the output voltage to swing symmetrically to the supply rails, +3.3V and ground.

Driver efficiency was not simulated for this TEC driver amplifier.

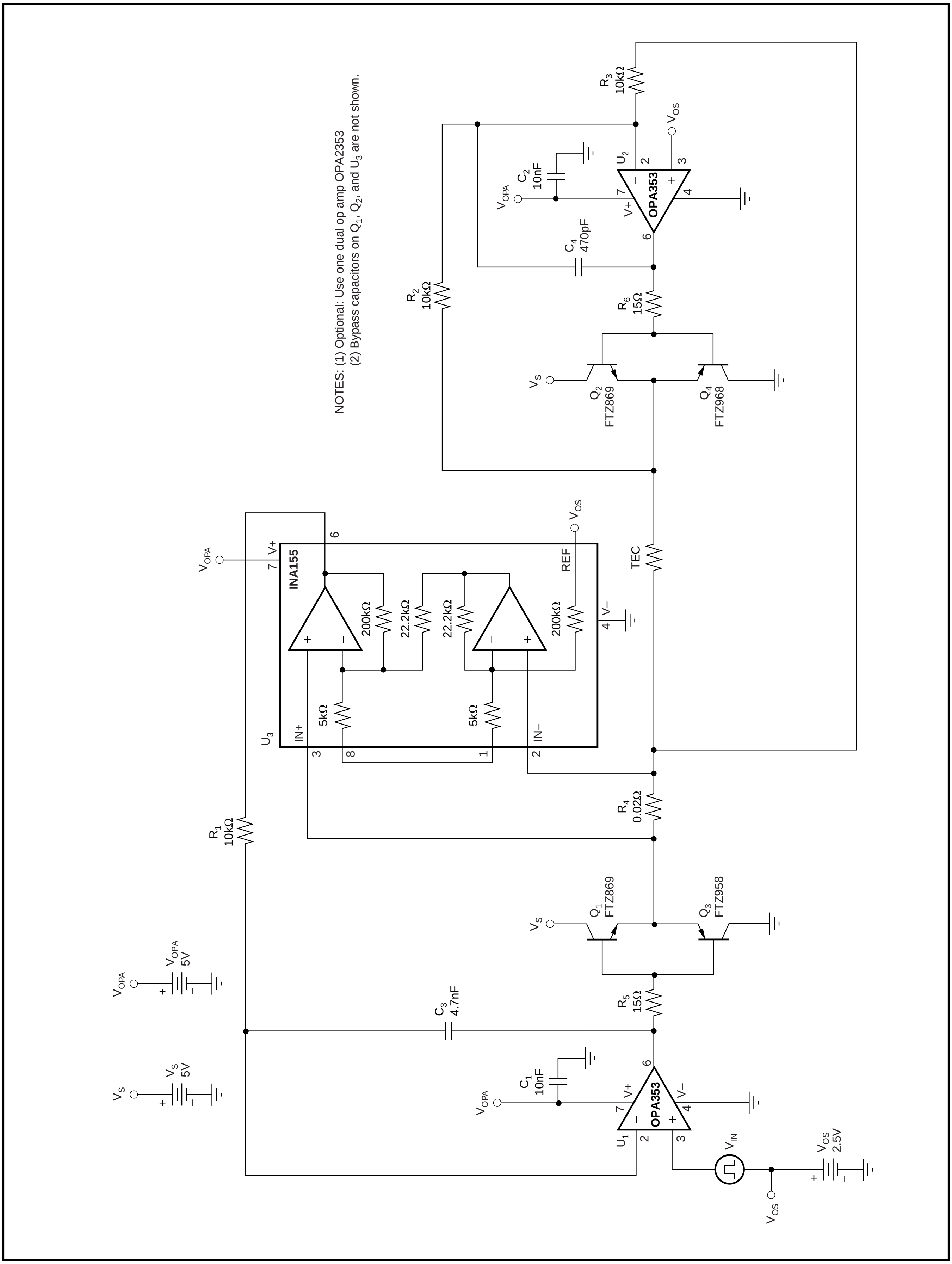

FIGURE 21. TEC Driver–2 Circuit Diagram.

Linear amplifiers do not reach the efficiencies of PWM switching types, but they do offer important advantages, principally their very low noise. Switching noise interference in laser and APD circuits is not a concern with linear drivers.

Driver amplifier loop stability was investigated by running an AC and Transient simulation. Results of the AC simulation are shown in Figure 24.

Some peaking is shown in the frequency response of the 1Ω and 1.5Ω TEC but increasing the capacitance of the compensation capacitors C_3 and C_4 can eliminate this. Notice that the capacitance of C_3 and C_4 are not the same. Amplifier U_1 requires more capacitance than U_2 because of the presence of feedback gain provided by the instrumentation amplifier, U_3.

Similar results are seen in Figure 25, a P–Spice transient simulation of this TEC driver amplifier. Slight peaking is noted for heavy loads as predicted by the frequency response curves.

Also evident in the transient response waveform is a small crossover distortion “glitch” around 0A output current due to the delay between turning off one transistor before its complimentary transistor turns on. For example, the op amp driving the power transistors must slew quickly between the voltage at which the NPN transistor turns off and the PNP transistor turns on. Due to the fact that this amplifier uses a Class B output stage, this region is $2V_{BE}$. A fast op amp such as the OPA353 minimizes the crossover distortion and enhances stability about the crossover region. For applications such as an audio amplifier, the output stage transistors could be biased slightly into conduction (class AB1) which would eliminate the crossover region altogether. Applications for driving loads such as a thermoelectric cooler do not warrant the increased complexity of a Class AB1 output stage.

Due to the asymmetry of output stage of this driver, it may not be the best choice for a TEC driver amplifier. For better efficiency, see the last section of “Linear TEC Driver–3.”

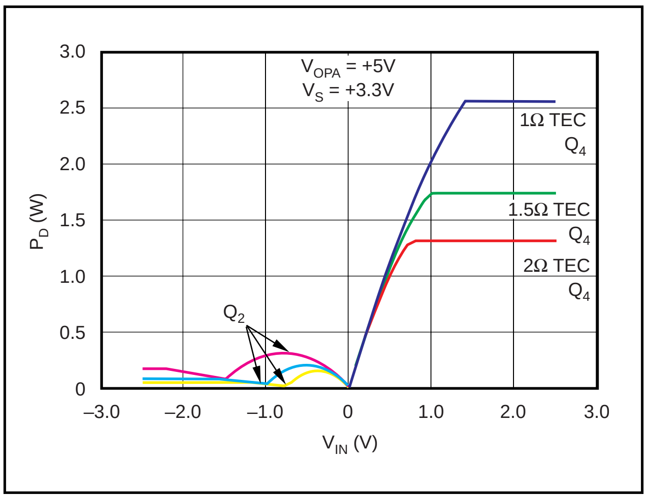

FIGURE 22. Output Transistor Q_2 and Q_4 Power Dissipation.

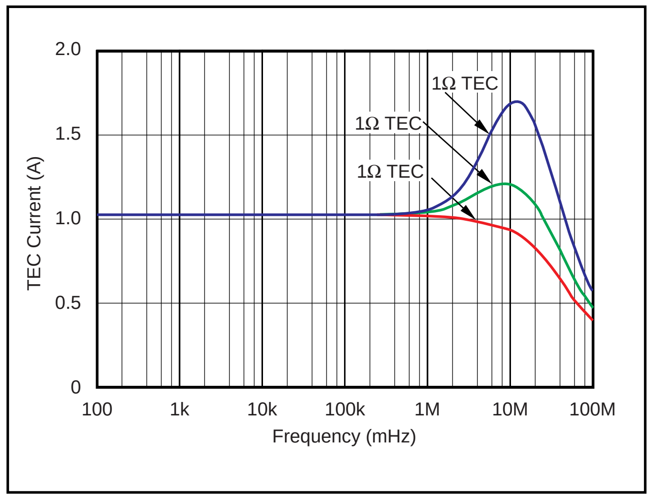

FIGURE 24. TEC Driver–2 Amplifier Frequency Response.

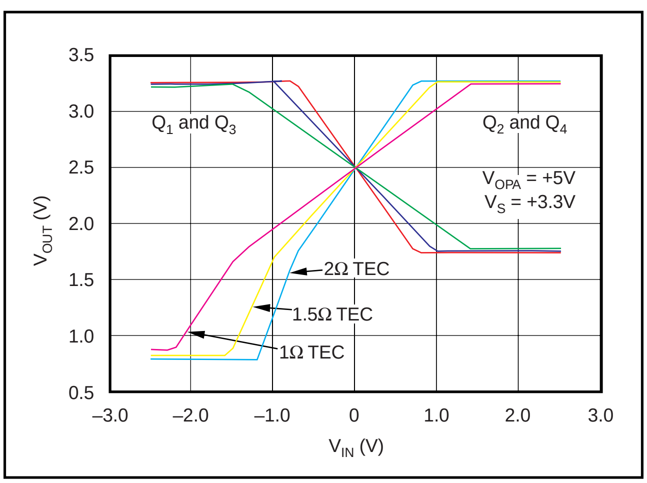

FIGURE 23. TEC Driver–2 Amplifier Output Voltage Swing.

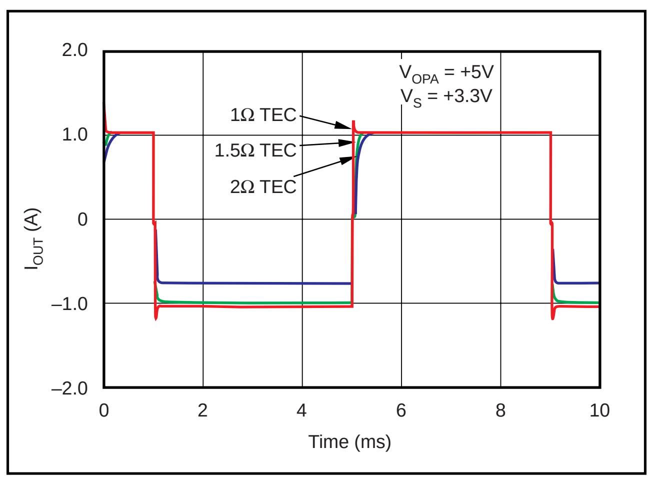

FIGURE 25. TEC Driver–2 Amplifier Transient Response.

LINEAR TEC DRIVER–3

The linear TEC driver circuit (see Figure 28) is capable of driving over ±2A into a typical TEC. The circuit operates on a bipolar ±2.5V supply and drives the TE cooler in the highly desirable “constant-current” mode. A BTL amplifier topology achieves bidirectional current output. This type of amplifier drives its load differentially, so the TEC must not be grounded on either end.

With bipolar supplies, an input offset voltage is not necessary to allow the amplifier outputs to swing in both directions. A level shift circuit will be necessary to interface with a single-supply input voltage source that is biased to $1/2V_{CC}$.

In the schematic, voltage V_{IN} is amplified by a R-R CMOS op amp U_1 with a class B power output stage (formed by the addition of a complimentary power transistor pair Q_1 and Q_3) that drives one end of a TEC load through shunt resistor R_4. A CMOS instrumentation amplifier (IA) U_3 senses the voltage drop across the shunt resistor and amplifies it by 50, it then feeds it back to the input of U_1. This feedback approach forces the output (TEC) current to be a function of V_{IN}. Shunt resistance and IA gain determines the scale factor of the circuit.

$$\frac{V_{IN}}{I_{OUT}} = A_V \bullet R_4, \text{ where } A_V \text{ is the IA gain in V/V.}$$

A P–Spice simulation (DC sweep) was performed on V_{IN}, sweeping the input voltage from –2.5V to +2.5V. The amplifier output current to the TEC is shown in Figure 26. This is a P–Spice Probe file output.

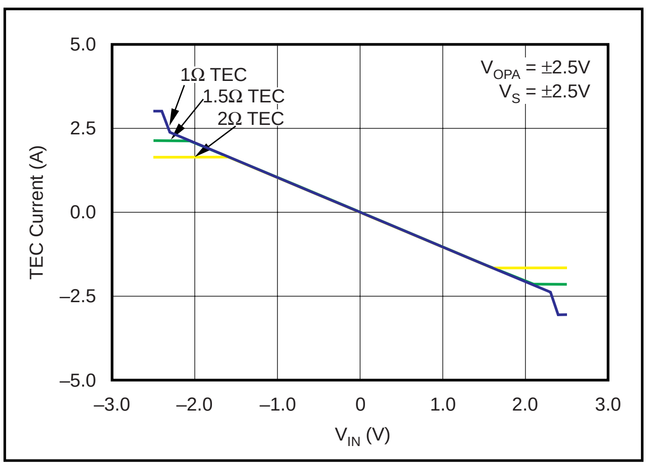

FIGURE 26. TEC Current Versus Input Voltage V_{IN}.

TEC current is shown for three different TEC sizes; 1Ω, 1.5Ω, and a 2Ω TEC. Operating on a bipolar ±2.5V supply, this driver is capable of driving a 1Ω or 1.5Ω TEC to 2A. Output voltage compliance limits the 2Ω TEC current to about 1.6A. This output capability is the same as TEC Driver–1, essentially the same circuit operating on a single +5V supply.

As Figure 26 illustrates, this TEC driver amplifier is a voltage-controlled current source. Constant-current drive assures that TEC drive current is independent of production variations in TEC junctions or long-term aging. Constant-current drive also eliminates the effect of thermal “back EMF” on current through the TEC under dynamic temperature control conditions.

Current limiting in a voltage-controlled current source becomes a simple matter of clamping the maximum input voltage to the amplifier. If driven by a rail-to-rail op amp, a voltage divider or pot will set the current limit. When the R-R op amp hits its rail, that voltage is divided down to an appropriate voltage that represents the maximum desired TEC current. The R-R op amp can’t swing past its rail, so this clamps the input voltage to the TEC driver amplifier.

To determine the circuit’s power dissipation and the requirements for heat sinking the SOT-223 output power transistors, a simulation was run by sweeping the DC input voltage as before using the same sizes of TECs. The results are shown in Figure 27.

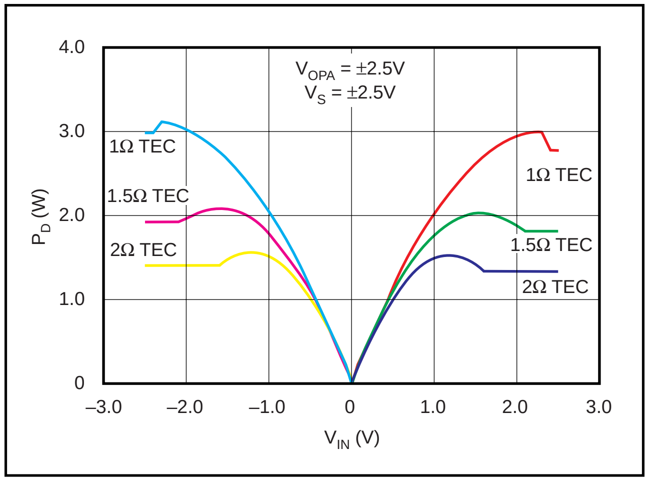

FIGURE 27. Output Transistor Power Dissipation.

Power dissipation of one NPN (Q_1) and one PNP (Q_3) power output transistor is shown in this sweep. The dissipation of the devices in the other half of the bridge (NPN = Q_2 and PNP = Q_4) will be the same as those shown in Figure 28. Depending on whether the TEC is in its cooling or heating mode, power is dissipated in Q_1 and Q_4 or in Q_2 and Q_3.

Driver efficiency is usually a concern in large multi-channel systems due to limitations on the total power dissipation of the system. Linear amplifiers do not reach the efficiencies of PWM switching types but they do offer important advantages, principally their very low noise. Switching noise interference in laser and APD circuits is not a concern with linear drivers.

The DC simulation data was used to plot the efficiency of the driver. To simplify the calculation, only the power output stage was considered. The CMOS OPA353 op amp power dissipation is only about 26mW, so deleting it contributes little error to the overall efficiency calculation. In this calculation, efficiency is considered to be the ratio of the power delivered to the load TEC to the power supplied to the driver. For example, a current of 1A into a 1Ω TEC represents a power P_{OUT} of 1W dissipated in the load ($P_{OUT} = I^2R$). The power supplied to the driver P_{IN} is 1A from each ±2.5V supply, or 5W ($P_{IN} = E \bullet I$).

Therefore:

$$\text{Eff}(\%) = \frac{P_{OUT}}{P_{IN}} \bullet 100, \text{ or Eff}(\%) = \frac{I^2 \bullet R}{V_S \bullet I_S} \bullet 100 = \frac{1A^2 \bullet 1\Omega}{5V \bullet 1A} \bullet 100 = 20\%$$

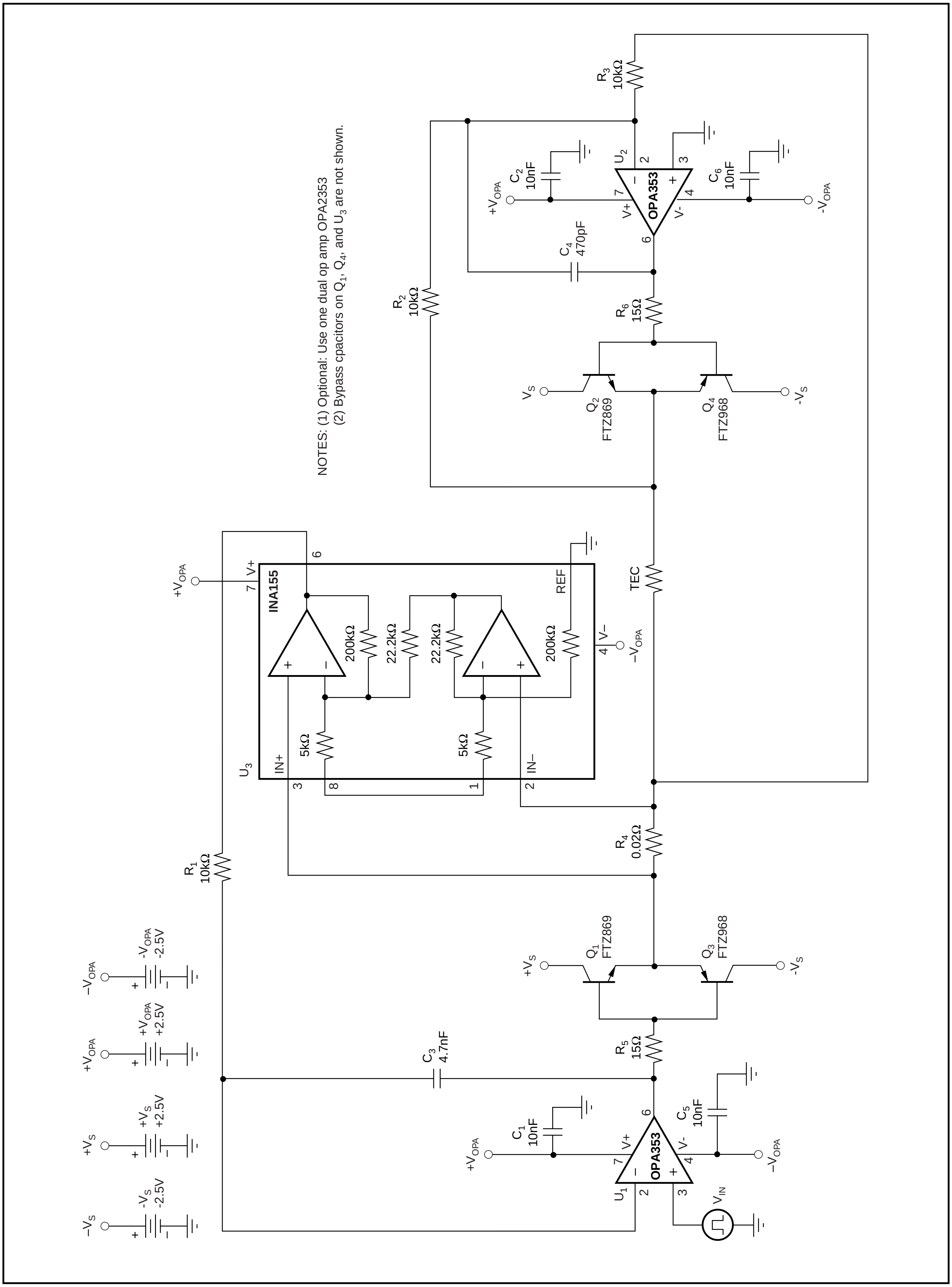

FIGURE 28. TEC Driver–3 Circuit Diagram.

TEXAS INSTRUMENTS

Swinging the output voltage close to the supply rails minimizes the voltage across the output transistor, thus reducing its power dissipation. Likewise, it maximizes the voltage across the load. As can be seen from the previous equation, this increases the efficiency of a linear driver. In fact, as seen in Figure 29, this circuit can reach an efficiency of over 60% under favorable conditions. The dissipation of TEC Driver–3 on ±2.5V is exactly the same as TEC Driver–1 on +5V.

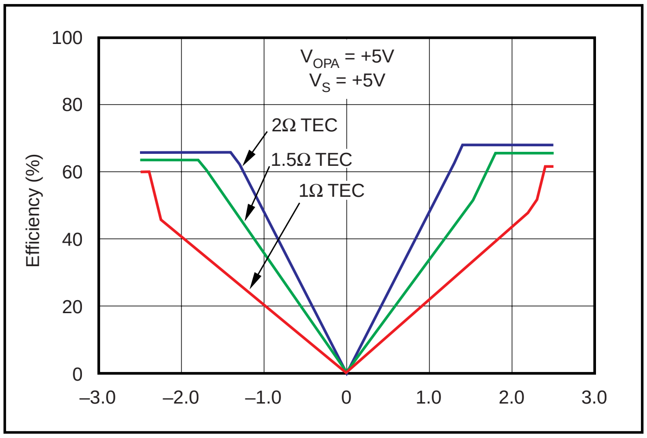

FIGURE 29. Driver–1 (+5V supply) and Driver–3 (±2.5V supplies) efficiency with 1Ω, 1.5Ω and 2Ω TEC loads.

Driver amplifier loop stability was investigated by running an AC and Transient simulation. Results of the AC simulation are shown in Figure 30.

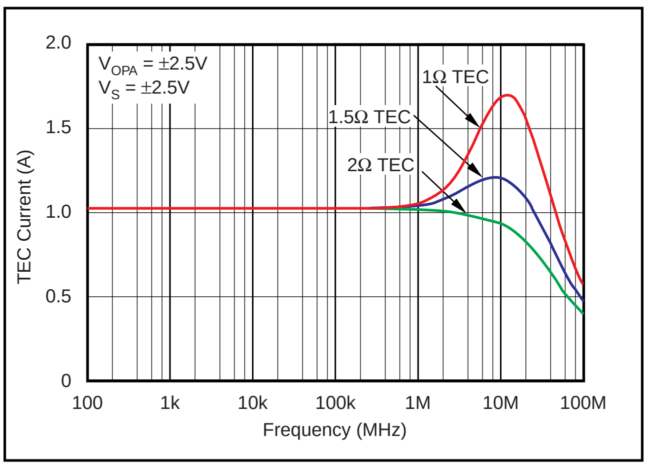

FIGURE 30. TEC Driver Amplifier Frequency Response.

Some peaking is shown in the frequency response curves of the 1Ω and 1.5Ω TEC but increasing the capacitance of the compensation capacitors C_3 and C_4 can eliminate this. Notice that the capacitance of C_3 and C_4 are not the same. Amplifier U_1 requires more capacitance than U_2 because of the presence of feedback gain, that is provided by the instrumentation amplifier, U_3.

Similar results are seen in Figure 31, a P–Spice transient simulation of this TEC driver amplifier. Slight peaking is noted for heavy loads as predicted by the frequency response curves.

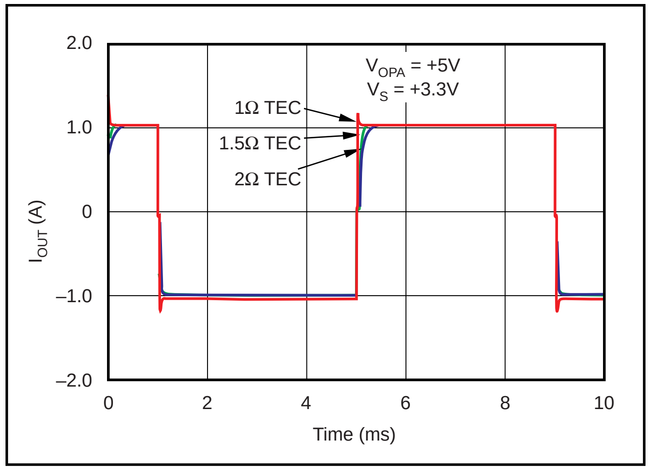

FIGURE 31. TEC Driver Amplifier Transient Response.

Also evident in the transient response waveform is a small crossover distortion "glitch" around 0A output current due to the delay between turning off one transistor before its complimentary transistor turns on. For example, the op amp driving the power transistors must slew quickly between the voltage at which the NPN transistor turns off and the PNP transistor turns on. Due to the fact that this amplifier uses a Class B output stage, this region is $2V_{BE}$. A fast op amp, such as the OPA353, minimizes the crossover distortion and enhances stability about the crossover region. For applications such as an audio amplifier, the output stage transistors could be biased slightly into conduction (class AB1) which would eliminate the crossover region altogether. Applications for driving loads such as a thermoelectric cooler do not warrant the increased complexity of a Class AB1 output stage.

Reducing the power output transistor supplies from ±2.5V to ±1.5V increases the overall efficiency of the driver amplifier by reducing the power dissipation in the output transistors. As expected, lower transistor V_{CE} results in lower power dissipation at the same collector current. Running the output stage on as low voltage as possible increases efficiency—especially if the amplifier output voltage can swing close to the rail.

Although the low-level circuits are operating on different supply voltages (±2.5V) than the output stage (±1.5V), this does not cause the same asymmetry problems as discussed in "TEC Driver–2." This is because the inputs are ground-referenced as are all of the supplies; both the op amp and the output stage can swing symmetrically both positive and negative.

See Figure 32 for output current into three sizes of TECs.

TEXAS INSTRUMENTS

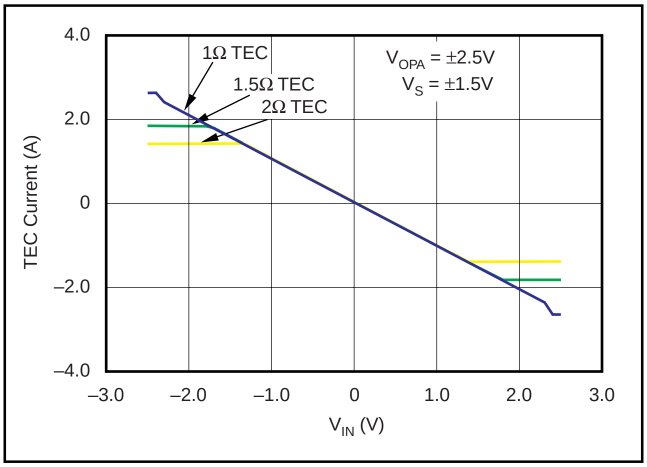

FIGURE 32. TEC Driver Output Current with ±1.5V Output Stage Supplies.

Operation on low voltage supplies does sacrifice output voltage compliance somewhat. This reduces the maximum drive current into a 2Ω or 1.5Ω TEC but lower resistance 1Ω TECs are unaffected.

Worthwhile gains in power dissipation are also realized by operating the output stage on ±1.5V supplies, as shown in Figure 33.

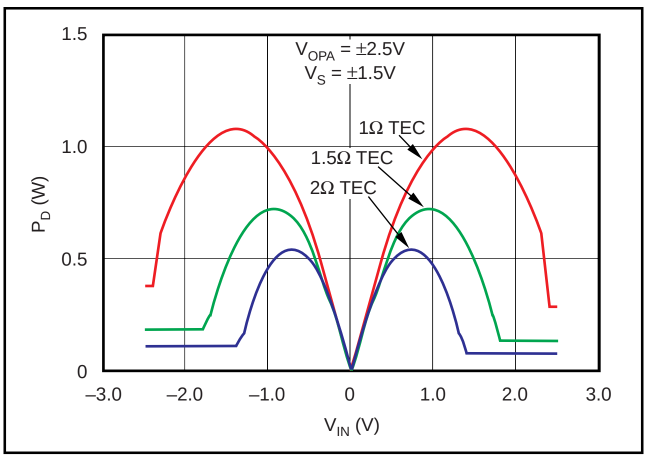

FIGURE 33. Output Transistor Power Dissipation with ±1.5V Supplies.

Smaller transistor heat sinks are required and reductions in cooling capacity are possible in large systems.

Figure 34 shows the very high efficiency that can be achieved by a linear TEC driver when it swings very close to its supply rails. Driving a 2Ω TEC to its maximum current of about 1.3A, this amplifier achieves an efficiency of about 90%. Driving a 1Ω TEC to its maximum current of about 1.3A, this amplifier achieves an efficiency of about 80%. Lower output currents achieve lower efficiencies but, at the same time, power dissipation is also lower.

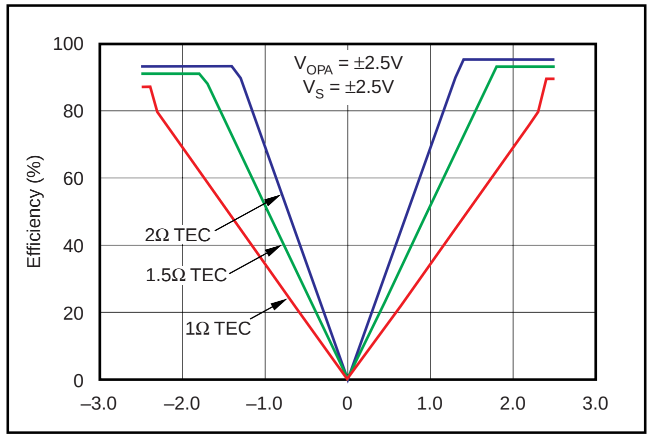

FIGURE 34. Improved Amplifier Efficiency with Output Stage Operating on ±1.5V Supplies.

The key to achieving good efficiency with a linear driver is to match the TEC driver amplifier characteristics with appropriate power supplies for your TEC. Thermoelectric coolers are available with a wide range of voltage and current characteristics; it is important to choose a TEC that requires a drive voltage that is very close to the voltage(s) available from your existing power supply. In new designs, it may be possible to chose a TEC and then specify a power supply voltage that optimizes the performance of the TEC and driver amplifier.

Satisfactory operation of these circuits should be verified in your actual application by breadboarding and testing.

TEMPERATURE UNDER–AND OVER–RANGE SENSING WITH A WINDOW COMPARATOR

The window comparator circuit shown in Figure 35 has been used to monitor a TEC operating temperature and indicate an out-of-range condition. Separate outputs are provided for indicating an over-temperature or under-temperature condition. A logic HI at the output of U_1 or U_2 indicates that the thermistor temperature is outside of the set points of the window comparator.

A logic LOW at both outputs indicates that the TEC temperature is within its safe operating range. A dual comparator with open-collector or drain outputs can be used in a “wired-OR” configuration but the single OPA340 or dual OPA2340 CMOS op amp offers higher accuracy.

In the schematic, voltage V_{IN} is used to simulate the voltage appearing across a thermistor that measures the laser/TEC temperature. If a 10kΩ at 25°C thermistor is excited by 100µA from a REF200 Current Reference, it will read 1.000V across the thermistor at 25°C. By using this constant-current source, the thermistor’s output voltage is a direct function of its temperature and this is easily converted to °C by using the thermistor’s calibration chart.

A P–Spice simulation (DC sweep) was performed on V_{IN}, sweeping the thermistor voltage from 0V to 2V. The lower temp limit was set to 500mV with R_{LOW}, a 5kΩ resistor. A 10kΩ resistor, R_{DELTA}, set the difference between the low and high thresholds to 1V; therefore, the high threshold was 500mV + 1.000V = 1.500V.

The P–Spice Probe output is shown in Figure 36. As the thermistor voltage crosses each threshold, U_1 and U_2 indicate their status with a TTL HIGH or LOW output.

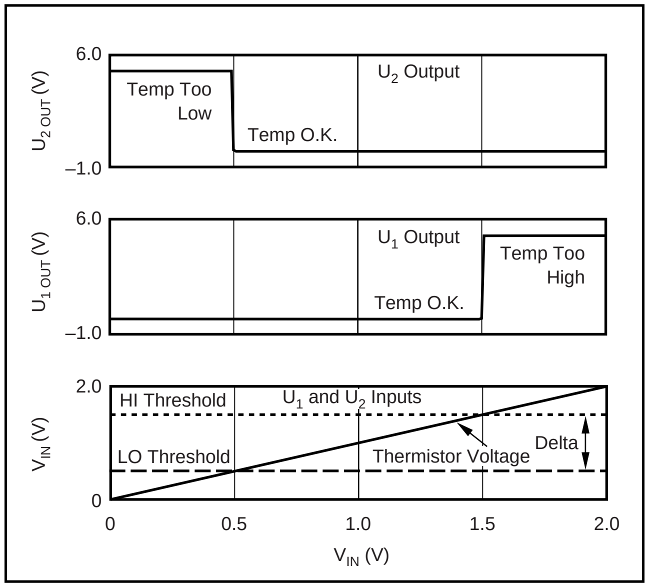

FIGURE 36. Input Voltage, HIGH and LOW Thresholds, and Ouput Voltages.

To prevent loading of the thermistor, a low bias current precision CMOS op amp was used as a comparator. Operating on a single +5V supply, the R/R op amp outputs are TTL/CMOS compatible. A dual OPA2340 is ideal for low–speed window comparator applications.

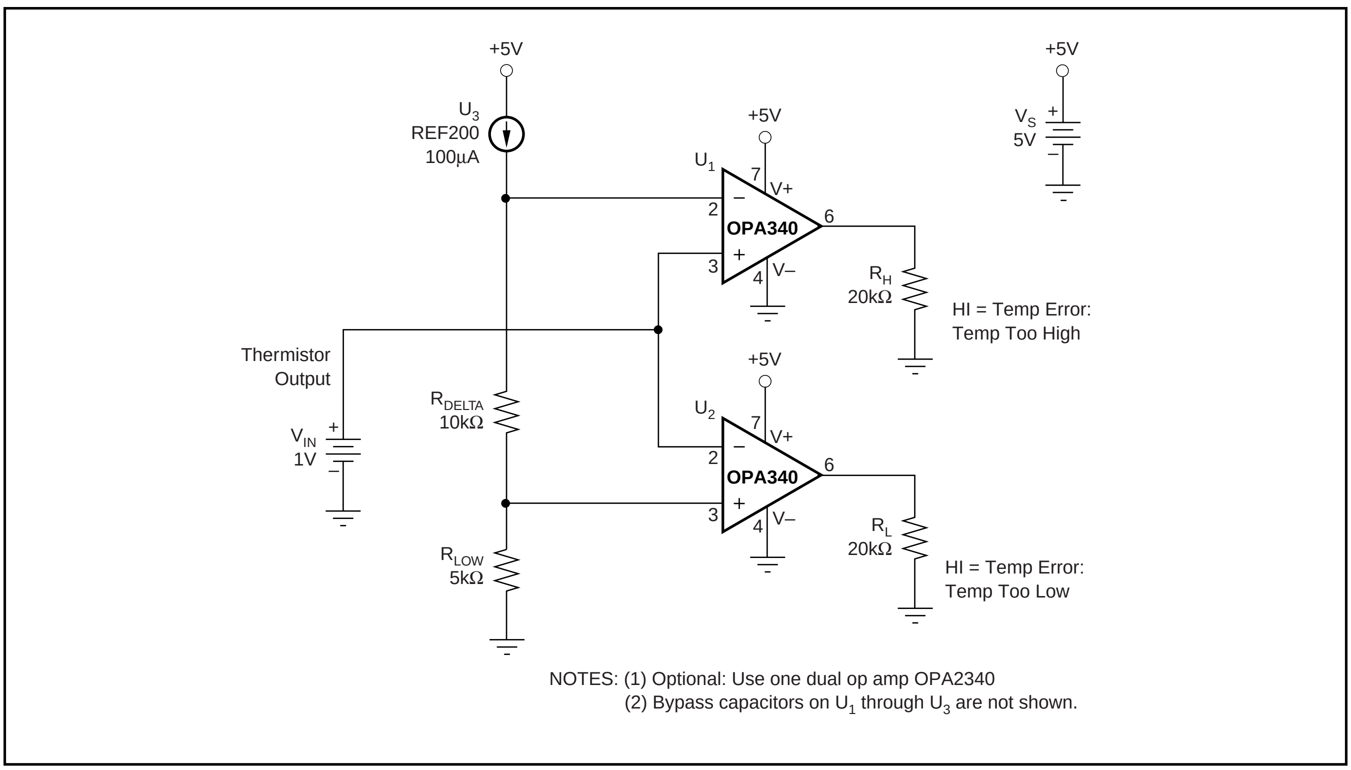

FIGURE 35. Precision Window Comparator Circuit Diagram.

One caution–the REF200 Current Reference requires about 2.5V of headroom to operate properly. This means that the upper threshold should not exceed 2.5V if this circuit is operated on the recommended +5V supply.

By setting the thresholds appropriately, a thermistor can also be monitored for an open or shorted condition.

This circuit can also perform a continuous check of whether the temp control system is operating within specification. Setting the window low temperature trip point to the lower desired operating temperature limit with R_{LOW} and its upper temperature trip point to the upper desired operating temperature limit by R_{DELTA} will allow this circuit to indicate when the TEC is within the desired control range. A digital HIGH appears at the output of U_1 or U_2 if the temperature exceeds either the upper or lower preset error bounds.

Additional sensitivity can be obtained by sensing an instrumentation amplifier (IA) output (this is usually the amplified error signal that is used by the control loop) instead of connecting the window comparator directly to the thermistor. Since the IA usually has gain, a larger signal is available to drive the comparator. In this case, the comparator is monitoring the amplified difference between the thermistor voltage and the temp set voltage rather than the thermistor directly. This can monitor the temperature control loop error and a second window comparator can monitor the thermistor temperature as described above.

Ordinarily, accurately setting both thresholds of a window comparator is a tedious process. Voltage excitation of the commonly-used three-resistor divider guarantees interaction between thresholds. To adjust thresholds, first one pot is adjusted—then the other. The second adjustment changes the first threshold point, etc. Threshold trimming is an iterative process for the standard window comparator circuit.

By using current excitation, only two fixed resistors or pots are needed. R_{LOW} sets the lower threshold and R_{DELTA} sets the “width” of the “window”. Adjusting the R_{DELTA} does not change the lower threshold and adjusting the lower threshold does not change the voltage *difference* between thresholds. Thus the adjustment procedure is greatly simplified.

Noisy environments may require a small amount of hysteresis (positive feedback) to prevent “chatter” on the outputs at the comparator switching points.

Satisfactory operation of this circuit should be verified in your actual application by breadboarding and testing.

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Mailing Address:

Texas Instruments
Post Office Box 655303
Dallas, Texas 75265