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:

\[ A_v = \left( \frac{R_f}{R_{in}} \right) + 1 \]

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\textsubscript{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\textsubscript{IN}) if the OPA445 is used as an inverter. The familiar inverting op amp gain equation is:

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

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\textsubscript{1}) on the output of a R–R CMOS OPA350 op amp (U\textsubscript{2}). Laser diode current is sensed by differentially measuring the voltage drop across a shunt resistor (R\textsubscript{SHUNT}) in series with the laser diode. The output current is controlled by the input voltage (V\textsubscript{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\textsubscript{IN}/I\textsubscript{OUT}) can be set to other values by choosing appropriate resistor values using the equation:

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

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.

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.

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.

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.
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 (Q₁) on the output of (U₁), 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 Q₁. 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:

\[ \frac{V_{IN}}{I_{OUT}} = \frac{R_1}{R_2} \cdot R_{SHUNT} \]

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 Q₁ is shown in the upper curve. Operating on a supply voltage of 3.3V, Q₁ 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 Q₁ operating on a 5V supply. The higher 5V supply voltage is advantageous if a higher output compliance voltage is required. Op amp U₁ 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 Q₁ operating on a 5V supply.

Changing the value of the shunt resistor or the scale factor will necessitate changing the compensation capacitor C₁. Verify that your circuit is stable before connecting an expensive laser diode to the output.
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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.

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

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.

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.

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.
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 $Q_1$ or $Q_4$ 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 $1\,\Omega$ 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 $5\,\text{V}$ supply, or $5\,\text{W}$ ($P_{IN} = E \times I$).

Therefore:

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

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.

Some peaking is shown in the frequency response curves of the $1\,\Omega$ and $1.5\,\Omega$ 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.

Driver–amplifier loop stability was investigated by running an AC and Transient simulation. Results of the AC simulation are shown in Figure 17.
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 U1 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 U1. 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.

$$\frac{V_{IN}}{I_{OUT}} = A_V \cdot R_4$$

where $A_V$ 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.

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.

Driver efficiency was not simulated for this TEC driver amplifier.

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.

$$V_{IN} = +5V$$

$V_{S} = +3.3V$

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.

NOTES: (1) Optional: Use one dual op amp OPA2353
(2) Bypass capacitors on Q₁, Q₂, and U₃ are not shown.
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.”
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 1/2VCC. A CMOS instrumentation amplifier (IA) U3 senses the voltage drop across the shunt resistor and amplifies it by 50, it then feeds it back to the input of U1. This feedback approach forces the output (TEC) current to be a function of VIN. Shunt resistance and IA gain determines the scale factor of the circuit.

\[ \frac{V_{IN}}{I_{OUT}} = A_v \cdot R_4, \text{ where } A_v \text{ is the IA gain in V/V.} \]

A P–Spice simulation (DC sweep) was performed on VIN, 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}](image)

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](image)

Power dissipation of one NPN (Q1) and one PNP (Q3) power output transistor is shown in this sweep. The dissipation of the devices in the other half of the bridge (NPN = Q2 and PNP = Q4) 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 Q1 and Q4 or in Q2 and Q3.

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 \cdot I).

Therefore:

\[ \text{Eff(\%)} = \frac{P_{OUT}}{P_{IN}} \times 100, \text{ or Eff(\%)} = \frac{I^2 \cdot R}{V_s \cdot I_s} \times 100 = \frac{1A^2 \cdot 1\Omega}{5V \cdot 1A} = 20\% \]
FIGURE 28. TEC Driver–3 Circuit Diagram.
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.

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

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.

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. 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.
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.

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.

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 U₁ or U₂ 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, U₁ and U₂ indicate their status with a TTL HIGH or LOW output.

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.**

**FIGURE 36.** Input Voltage, HIGH and LOW Thresholds, and Output Voltages.

NOTES: (1) Optional: Use one dual op amp OPA2340
(2) Bypass capacitors on U₁ through U₂ are not shown.
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 $R_{\text{LOW}}$ and its upper temperature trip point to the upper desired operating temperature limit by $R_{\text{DELTA}}$ 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. $R_{\text{LOW}}$ sets the lower threshold and $R_{\text{DELTA}}$ sets the “width” of the “window”. Adjusting the $R_{\text{DELTA}}$ 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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