Procedure

Lab 7: Output Filtering and Project Integration

ECE 327: *Electronic Devices and Circuits Laboratory I*

Abstract

In this lab, you complete the quarter project. First, you verify that the entire project can be powered from a 10 V supply rail (provided by two different LM317 type regulators). Next, you build and test the filter that smooths the demodulated signal. Penultimately, you build and test the high current output stage that drives an 8 Ω speaker. Finally, you integrate all components together. In this document, we present some basic topics and procedures to help you complete (and understand) the lab.

Contents

Note: Lab Report Data

The main content of this laboratory is found in Section 3 and Section 4. Thus, the only data that needs to be saved for the lab report is found in those sections. The other sections, Section 1 and Section 2, make sure your circuit is ready for integration. While you do not need to save data from those sections, you still may need to complete them in order to proceed to the rest of the lab.

1 Single-Rail LSA

In the analog-to-digital (ADC) lab, you built a level-shifter amplifier (LSA) that takes a signal that fits within a $2\,\mathrm{V_{pp}}$ envelope centered at $0\,\mathrm{V_{DC}}$ and outputs a signal that spans a $6\,\mathrm{V_{pp}}$ envelope centered at $5\,\mathrm{V_{DC}}$. That LSA must be powered by a single $10\,\mathrm{V}$ supply rail, and so if your LSA depends upon a $-10\,\mathrm{V}$ supply rail as well, then it needs to be modified.

We present two simple single-rail LSA designs here. Both are AC coupled, and so the input signal offset will have no impact on the output[1]; however, we assume the input signal has negligible (i.e., $\sim 0\,\mathrm{V_{DC}}$) offset.

Test your LSA with $2\,\mathrm{V_{pp}}$ sinusoidal inputs at $50\,\mathrm{Hz}$, $1\,\mathrm{kHz}$, and $10\,\mathrm{kHz}$.

Operational Amplifier LSA

A simple LSA can be constructed with an operational amplifier[2], as shown in Figure 1.1.

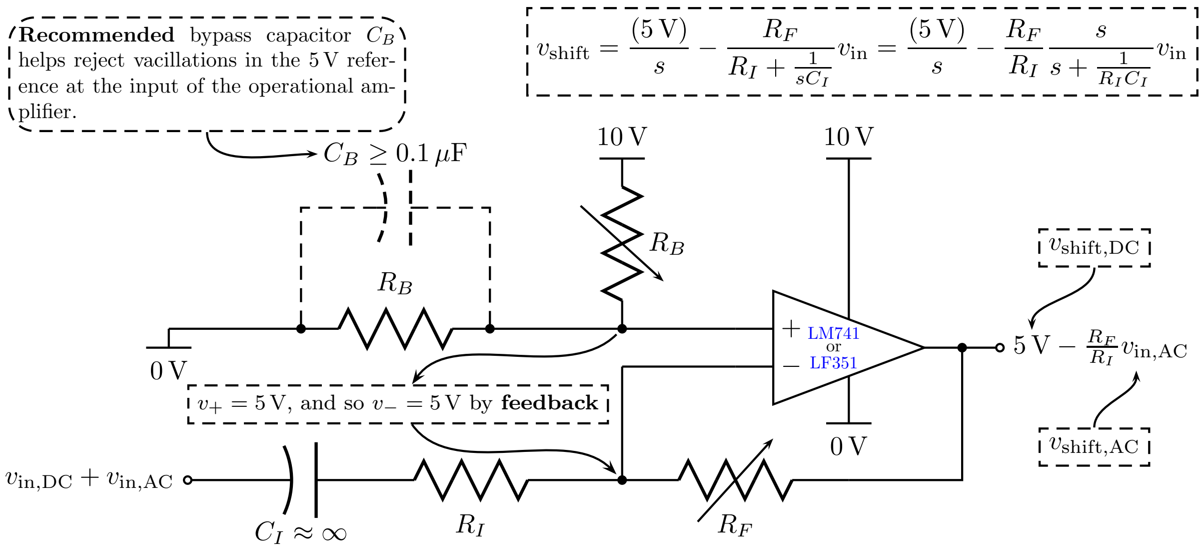


Figure 1.1: Single-ended level-shifter amplifier implemented with operational amplifier.

To provide a gain of 3 to signals of interest, R_F and R_I must be chosen so that

$$\frac{R_F}{R_I} = 3, \qquad \text{which means} \qquad R_F = 3R_I \qquad \text{and} \qquad R_I = 0.25 \times \underbrace{(R_I + R_F)}_{\text{potentiometer total}}. \tag{1.1}$$

Using a potentiometer for the R_I–R_F divider, choose components[3] so that

$$C_I \leq 2\,\mu\mathrm{F} \qquad \text{and} \qquad \frac{1}{2\pi R_I C_I} \leq 35\,\mathrm{Hz} \qquad \text{and} \qquad 10\,\mathrm{k\Omega} \leq R_F \leq 50\,\mathrm{k\Omega}. \tag{1.2}$$

Use the R_I–R_F potentiometer to **tune the gain**. After tuning the gain, **be sure your half-power frequency** is *no higher* than $35\,\mathrm{Hz}$ and increase C_I if needed.

The output DC offset is set with the R_B–R_B divider, which **should be implemented with a potentiometer**. Use the R_B–R_B potentiometer to **tune the offset**. **If you can**, ensure that[4]

$$R_B \approx 2R_F \qquad \text{and} \qquad 1\,\mathrm{k\Omega} \leq R_B \leq 500\,\mathrm{k\Omega} \tag{1.3}$$

for good high frequency performance, low current draw, and high robustness to device variations.

[1] If coupling by *electrolytic capacitor*, the cathode must be connected to the signal with the lowest average (i.e., DC offset).
[2] A LM741 is recommended, but a LF351 may also be used.
[3] Try starting with potentiometer *total* $R_I + R_F \geq 20\,\mathrm{k\Omega}$ and $C_I \geq 0.47\,\mu\mathrm{F}$.
[4] Try starting with potentiometer total $R_B + R_B \geq 50\,\mathrm{k\Omega}$ (and $C_B \geq 0.1\,\mu\mathrm{F}$).

NPN Common-Emitter Amplifier LSA

Alternatively, the common-emitter amplifier LSA shown in Figure 1.2 can be used.

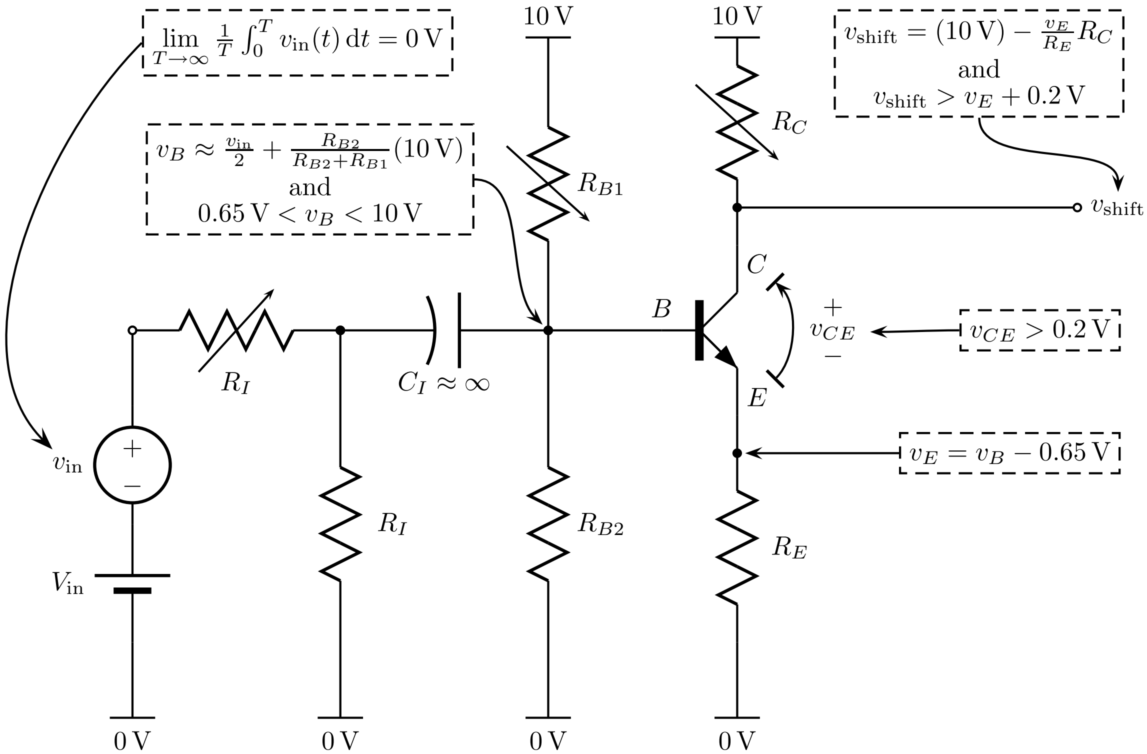


Figure 1.2: Level-shifter amplifier implemented with attenuated-input common-emitter NPN configuration.

For this LSA to work well, components *must* be chosen so that

$$0 \ll R_I \| R_I \ll R_{B1} \| R_{B2} \ll \beta R_E \tag{1.4}$$

where $\beta \approx 100$. Assuming Equation (1.4) holds, it must also be that[5]

$$C_I \leq 1\,\mu\text{F} \qquad \text{and} \qquad \frac{1}{2\pi\,(R_{B1}\|R_{B2})\,C_I} \leq 35\,\text{Hz} \qquad \text{and} \qquad 10\,\text{k}\Omega \leq (R_{B1}\|R_{B2}) \leq 50\,\text{k}\Omega. \tag{1.5}$$

For an LSA gain of 3, R_C and R_E should be chosen so that

$$\frac{R_C}{R_E} = 6. \tag{1.6}$$

For tuning, use a variable resistor for R_C. The R_I resistors also have an impact on the LSA gain, so one of the R_I resistors may be implemented with a variable resistor or the entire R_I–R_I divider may be implemented with a potentiometer. Tune R_C and R_I for the right gain.

For an output DC offset of 5 V, biasing resistors R_{B1} and R_{B2} should be chosen so that

$$\frac{R_{B2}}{R_{B2} + R_{B1}} = \frac{8.9}{60} = \frac{89}{600} = 0.14833\cdots, \tag{1.7}$$

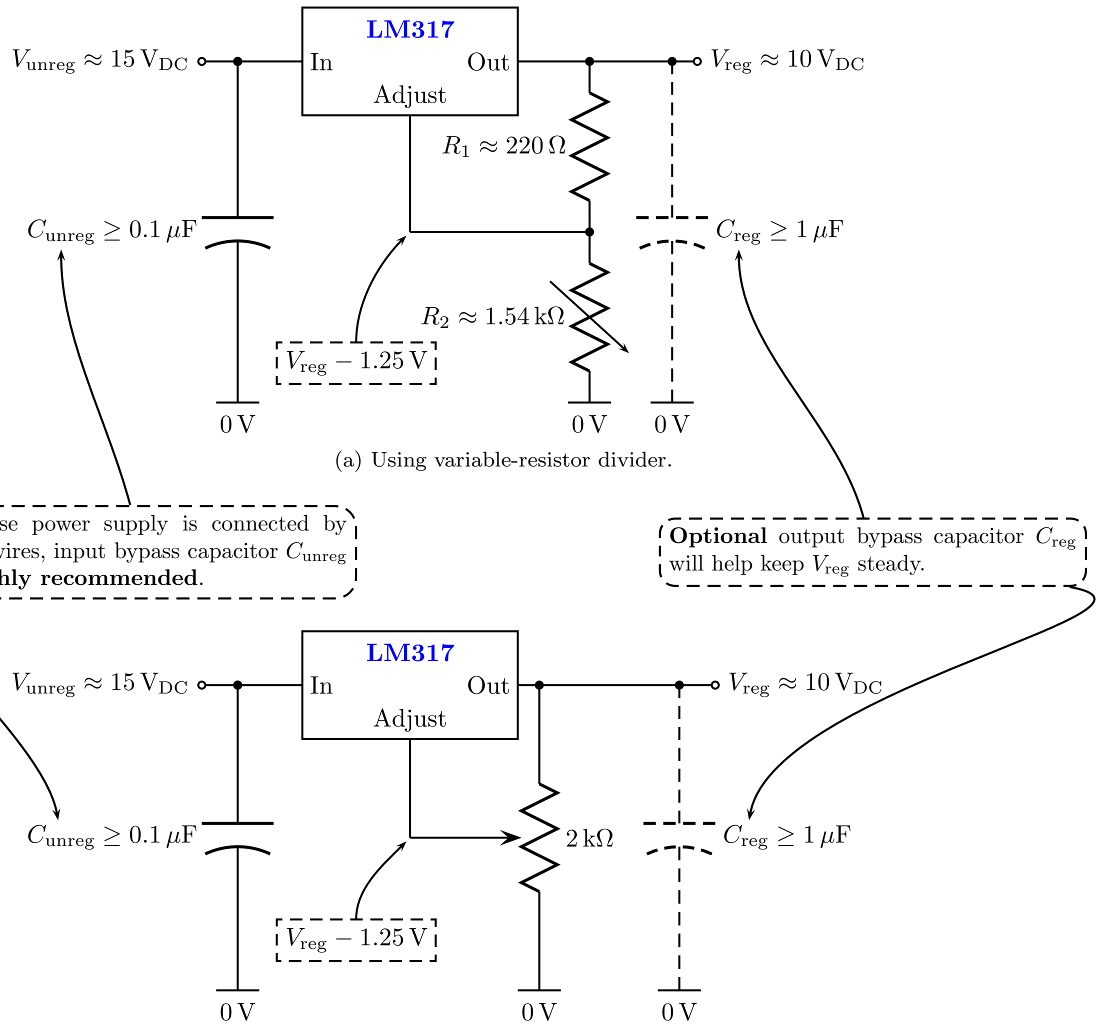
so the transistor base should see a DC average of ~1.483 V. For tuning the output DC offset, implement R_{B1} with a variable resistor, or implement the entire R_{B1}–R_{B2} divider with a potentiometer. As you change $R_{B1}\|R_{B2}$, be sure that Equation (1.4) still holds. After tuning, make sure the half-power frequency is *no higher* than 35 Hz and increase C_I as needed.

[5] Try $R_I = 1\,\text{k}\Omega$, $(R_{B1}\|R_{B2}) \approx 12.75\,\text{k}\Omega$ (i.e., $R_{B1} \approx 85\,\text{k}\Omega$ and $R_{B2} \approx 15\,\text{k}\Omega$), $C_I = 680\,\text{nF}$ (or 470 nF), and $R_E = 10\,\text{k}\Omega$.

2 Using Regulated Supply Rails

Your transmitter and receiver circuits should *each* have their own 10 V regulated supply. If you haven't already, connect each $V-$, V_{EE}, or V_{SS} to 0 V, and connect each $V+$, V_{CC}, or V_{DD} to a *regulated* 10 V supply rail. The transmitter and receiver should *each* have their *own* regulated supply.

The two regulated supply circuits in Figure 2.1 are identical. The LM317 maintains 1.25 V potential difference between its "Out" and "Adjust" pins, and the voltage divider acts as a lever propping V_{reg} above the ground reference. Your references should have $V_{\text{unreg}} \approx 15.0\,\text{V}$ and $V_{\text{reg}} \approx 10.0\,\text{V}$.



(a) Using variable-resistor divider.

(b) Using potentiometer.

Figure 2.1: LM317 adjustable voltage regulator. Use $V_{\text{unreg}} \approx 15\,\text{V}$ and set $V_{\text{reg}} \approx 10.0\,\text{V}$.

In Figure 2.1(a), the (temperature-independent) output is given by

$$V_{\text{reg}} = (1.25\,\text{V}) \times \left(1 + \frac{R_2}{R_1}\right). \tag{2.1}$$

The less conventional circuit in Figure 2.1(b) may also be used (with caution).

In either circuit, input capacitor $C_{\text{unreg}} \approx 0.1\,\mu\text{F}$ should be used to steady the input supply rail. Output capacitor $C_{\text{reg}} \approx 1\,\mu\text{F}$ may be used if artifacts are being coupled into your output via the supply (e.g., clock noise). A $\sim$0.1 μF bypass capacitor to ground can be placed at the 10 V input to each circuit component. Finally, a **large** bypass capacitor (e.g., 1–10 μF) may be placed *near* LM317 from `Adjust` to ground.

3 Smoothing Low-pass Filter

The output of our pulse-width demodulator looks like a sampled (and held) version of our LSA output. The sharp edges in the signal introduce harmonic distortion[6] that will be noticeable to the ears, so we must smooth the signal through a low-pass filter (LPF) to remove these higher harmonics.

Sallen-and-Key Filter

Active filters (e.g., filters that make use of operational amplifiers) provide simple design methods that do not require troublesome components like inductors. The active filter in Figure 3.1 implements a *Sallen-and-Key*[7] *(or Sallen-Key) filter*, which has low-pass, bandpass, or high-pass characteristics based on component values.

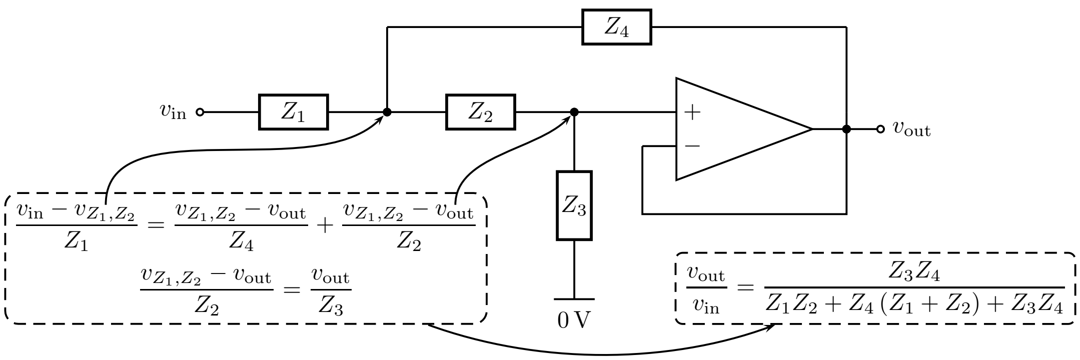


Figure 3.1: Generic unity-gain Sallen-Key filter implemented with operational amplifier.

Recall that the impedance (to ground) looking into the buffer's output is zero. If Z_4 was connected *directly* to ground, the circuit would be a cascade of two voltage dividers. Using a buffered output at Z_4 sharpens the cascade response by bootstrapping output signal back into the input. The unity-gain buffer (which controls the "ground" of Z_4) gives the filter a unity-gain response in its passband. Other buffers (e.g., transistor emitter followers and non-unity gain operational amplifier configurations) can be used as well.

Sallen-and-Key Low-Pass Filter

In our case, we use the unity-gain Sallen-and-Key LPF shown in Figure 3.2.

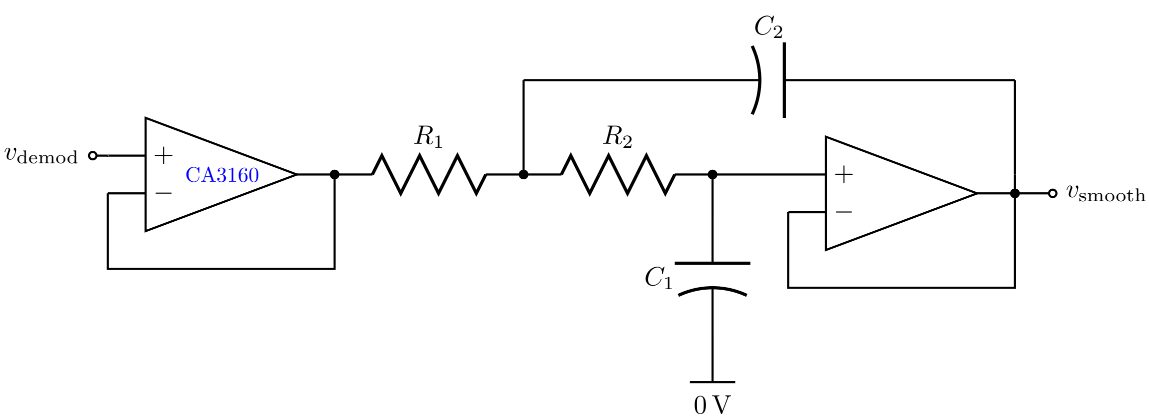


Figure 3.2: Sallen-and-Key low-pass filter ($C_2 \gg C_1$) with input buffer.

This circuit matches Figure 3.1 with $Z_1 = R_1$, $Z_2 = R_2$, $Z_3 = 1/(sC_1)$, $Z_4 = 1/(sC_2)$, and an input buffer.

[6] *Total harmonic distortion (THD)* is the standard measure of this form of distortion. Also see *SINAD* and *ENOB*.
[7] Sallen-Key filters are named after MIT professors R.P. Sallen and E.L. Key who invented it.

If we make $C_2 \gg C_1$, the circuit can be qualitatively analyzed for different input frequencies.

Low: At low frequencies (e.g., DC), neither capacitor passes any current, and so the two resistors pass no current. Therefore, the circuit input and output must match (i.e., unity gain).

Medium-Low: At slightly higher frequencies, the feedback capacitor C_2 *bootstraps*[8] the output to the R_1 resistor, which provides a path for current to flow through the resistor. Because the resistor conducts current, the signal seen by the output buffer will be attenuated. The low-frequency pole comes from the C_2 feedback capacitor.

Medium-High: At even higher frequencies, the C_2 capacitor to ground starts to conduct current, which further attenuates the signal through the R_2 resistor. The high-frequency pole comes from the C_1 capacitor to ground.

High: At very high frequencies, the C_1 capacitor to ground has very low impedance (i.e., it looks like a short circuit to ground). Thus, the high-frequency signal never reaches the input to the buffer.

The capacitors in the circuit setup frequency-dependent attenuation.

Bootstrapping revisited: Connecting C_2 directly to ground makes the circuit two cascaded RC low-pass filters. By bootstrapping the "bottom" of C_2 by the output, we increase attenuation with frequency.

The input buffer: Because the output of our PWM demodulator circuit (i.e., the "DAC") is a simple capacitor, an input buffer is needed. Sourcing the filter current from the capacitor would attenuate the input signal. So a low-input-leakage CA3160 MOSFET (BiMOS) operational amplifier connected as a follower should be used as the input buffer.

Butterworth Filters

Components in Table 3.1 are for a second-order *Butterworth*[9] *filter* with $f_0 = 1/(2\pi\sqrt{R_1 R_2 C_1 C_2})$.

Component	Theoretical Value	Cabinet Value
R_1	$30.2\,\mathrm{k\Omega}$	$33\,\mathrm{k\Omega}$
R_2	$1.68\,\mathrm{k\Omega}$	$1.8\,\mathrm{k\Omega}$
C_1	$470\,\mathrm{pF}$	$470\,\mathrm{pF}$
C_2	$4.7\,\mathrm{nF}$	$4.7\,\mathrm{nF}$
Frequency (f_0):	$15\,\mathrm{kHz}$	$13.8\,\mathrm{kHz}$

Table 3.1: Components for 2-pole Sallen-Key Butterworth LPF with knee f_0.

As you will see, Butterworth filters have maximally flat amplitude response in the passband. Unfortunately, they have two major drawbacks.

Poor Roll-off: Butterworth high-frequency roll-off is shallow, which means that high-frequency signals are not greatly attenuated. To remedy this, our power amplifier has provide an additional high-frequency pole.

Poor Phase Response: Butterworth low-pass filters have extremely poor phase response near the corner frequency. The *non-linear phase* response delays some frequencies more than others. The result is that inputs with significant high-frequency components are distorted on the output because their higher-frequency components are re-combined out of phase.

Together, fast inputs to your circuit may be significantly distorted after smoothing.

[8] *Bootstrapping* refers to using output current to change the operation of circuit internals. Using a capacitor to bootstrap high frequencies into the middle of a circuit is common practice.

[9] Butterworth filters named for engineer Stephen Butterworth. Filter has maximally flat passband response (i.e., $Q = 1/\sqrt{2}$).

Procedure

Set **trigger** Edge for the **input** channel. Adjust **level** knob and **filtering** (see Mode) to stabilize display.

1. Build the input-buffered Sallen-and-Key Butterworth low-pass filter in Figure 3.3 **BELOW**.
   - Use the **REGULATED** 10 V_{DC} RECEIVER supply rail.
   - **DO NOT** connect filter input to demodulator YET.
   - If no CA3160 available, use CA3130 with 45–100 pF (e.g., 68 pF) from pin 1 to pin 8.
   - To prevent instability (e.g., oscillations on output), use an LM741 instead of an LF351.

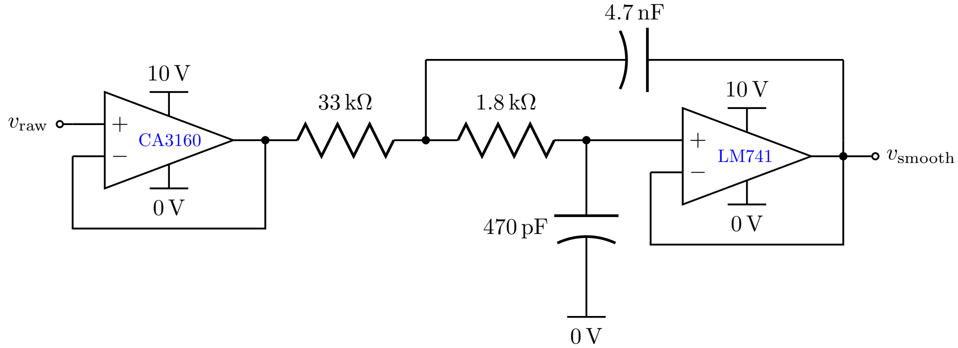


Figure 3.3: Sallen-and-Key Butterworth ∼13.8 kHz low-pass filter with input buffer.

2. Apply 1 kHz 6 V_{pp} sine wave with 5 V_{DC} offset to v_{raw} input.
   - Verify input on scope. May need to set offset on old function generators to ∼2.5 V for 5 V offset.
   - **Bypass capacitors** (e.g., 0.1 µF) at each OA's 10 V supply pin can reduce 30 kHz clock noise.
   - **SAVE A PLOT** of filter input v_{raw} and output v_{smooth}.
3. Output magnitude should decrease with frequency. Find −3 dB "corner" or "breakpoint" frequency.
   - Use Quick Meas to show Peak to Peak of sources 1 and 2. Also show Frequency.
     - Turn on Averaging under Acquire to stabilize measurements. Use ∼1 # samples.
   - **Verify** input signal is 6 V_{pp} with 5 V_{DC} offset (i.e., swings from 2 V to 8 V).
   - **Tune** input signal's frequency until output signal drops to $(6/\sqrt{2})$ V_{pp} (i.e., ∼4.243 V_{pp}). This frequency is the −3 dB or *half-power* frequency.
     - **NOTE** the filter delay at each frequency. When does it become frequency-dependent?
   - At the −3 dB frequency, **SAVE A PLOT** of input v_{raw}, output v_{smooth}, and measurements.
4. Disconnect function generator and connect **PWM demodulator output** to the input v_{raw}.
   - Output is a 100 pF capacitor that is *not* connected to a CA3160 (i.e., only connected to switch).
5. Generate a 1 kHz 2 V_{pp} sine wave with 0 V_{DC} offset at your PWM's *LSA INPUT*.
   - Probe ***LSA* OUTPUT** and **FILTER output** v_{smooth}. Press Single if necessary.
   - Except for a phase shift, the two outputs should be *nearly* identical.
     - To improve match, adjust IR link distance. Be sure input is connected to demodulator output.
6. At the LSA INPUT, place 2 V_{pp} sine waves with 0 V_{DC} offset at 100 Hz, 1 kHz, and 10 kHz.
   - Probe SMOOTHING FILTER **input** v_{raw} and **output** v_{smooth}. Press Single if necessary.
   - Best results at 1 kHz. If 100 Hz results are very bad, then increase size of LSA's *input* capacitor.
   - **SAVE A PLOT** of FILTER's v_{raw} and v_{smooth} for EACH of these **three** frequencies.

   A strange 10 **kHz output** is due to slow sampling & shallow non-linear-phase Butterworth response.

4 Power Amplifier

Now that our signal has been smoothed, it is ready to be delivered to the output speaker. For the speaker output to be audible, we must deliver at least 0.25 W of *average* power to the speaker. For a DC potential v_{DC} driving a DC current i_{DC} through a resistor R, the average power P_{DC} delivered to the resistor load is

$$P_{\text{DC}} = i_{\text{DC}} v_{\text{DC}} = i_{\text{DC}}^2 R = \frac{v_{\text{DC}}^2}{R}.$$

Because our input changes in time, we (arbitrarily) design our amplifier for *sinusoidal* input signals. The average power of a sinusoid is half of the power from a DC signal with the same *peak*[10], and so the power P delivered to resistor R from a sinusoid with peak v driving peak current i is

$$P = \frac{1}{2} i v = \frac{1}{2} i^2 R = \frac{1}{2} \frac{v^2}{R}. \tag{4.1}$$

So by Equation (4.1), for

$$P = 0.25\,\text{W} \quad \text{and} \quad R = 8\,\Omega,$$

it must be that

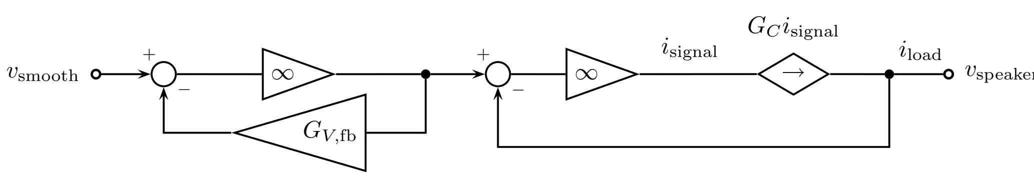
$$v = 2\,\text{V} \quad \text{and} \quad i = 250\,\text{mA}.$$

That is, to deliver .25 W to an 8 Ω load, our prototypical output sinusoid must have a peak 2 V potential and a peak 250 mA current. However, our smoothing filter cannot deliver much more than 20 mA of current (i.e., our operational amplifier is not designed to drive a speaker). Therefore, we need to build a *power amplifier* that has the ability to drive our smoothed signal into our speaker. The power amplifier has two components:

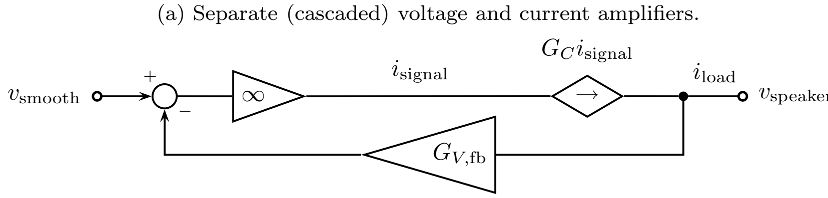
Voltage Amplifier: Increases the amplitude of our signals so that even small components of the signal can be heard through the speaker. Also provides another high-frequency pole for additional smoothing.

Current Amplifier: Ensures that enough current is available to deliver enough power into the 8 Ω speaker load.

To improve the linearity of the current amplifier, we use voltage feedback. The amplifier in Figure 4.1(a) shows how the current amplifier can be integrated into a unity-gain voltage amplifier. The resulting high-current buffer can then be cascaded with the voltage amplifier. However, because the voltage amplifier uses voltage feedback itself, the current amplifier can be incorporated into the voltage amplifier directly, as shown in Figure 4.1(b). Both abstract topologies in Figure 4.1 can be implemented with operational amplifiers.



(a) Separate (cascaded) voltage and current amplifiers.



(b) Integrated voltage and current amplifier.

Figure 4.1: Abstract power amplifier topologies.

[10] This fact is why the *root-mean-square (RMS)* amplitude of sinusoidal signal is the peak value divided by $\sqrt{2}$. The RMS amplitude is the amplitude of a DC signal that delivers the same (average) power to a linear resistor.

Voltage Amplifier

The signal v_{smooth} that leaves the smoothing filter may have AC frequency components with 3 V peaks at times, which is more than sufficient for delivering 0.25 V to the 8 Ω speaker. However, most signals will have smaller peaks, so at the risk of *overdriving*[11] our output, we build a voltage amplifier with a passband gain of at least 2.

The amplifier we use is shown in Figure 4.2.

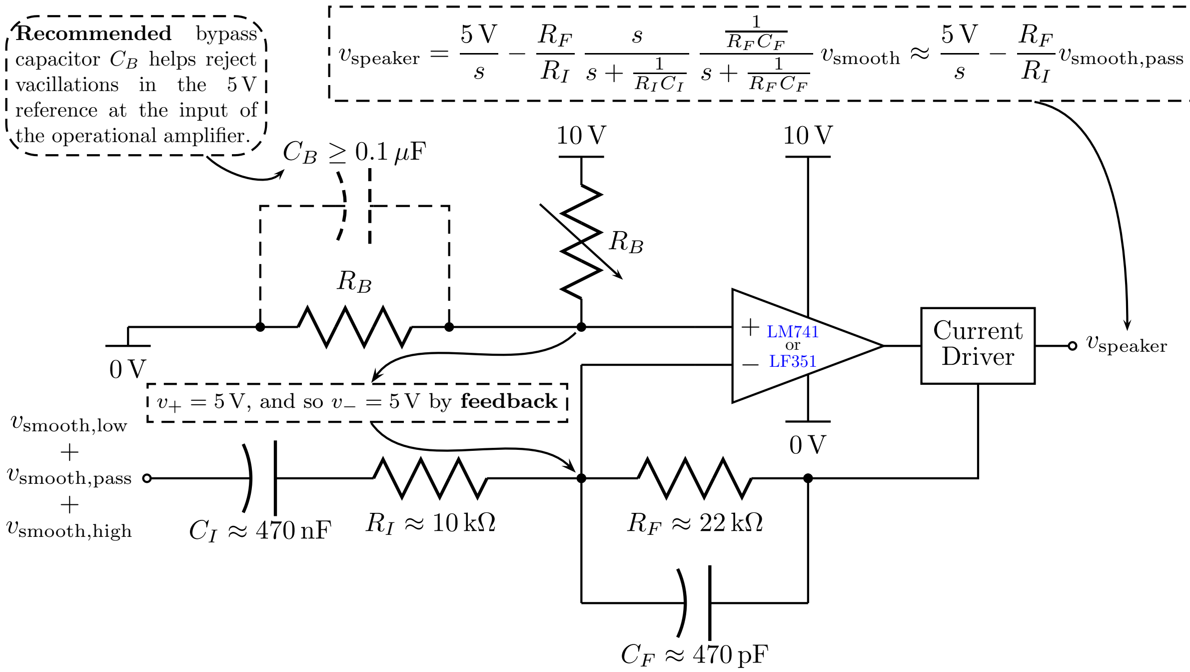


Figure 4.2: Single-ended voltage amplifier with high-frequency pole.

The *Current Driver* will be introduced later. **For now (and for testing your voltage amplifier), assume it is just a wire junction. That is, assume it shorts its three terminals together.**

Compare this amplifier with the level-shifter amplifier in Figure 1.1; they are identical except that this amplifier adds a feedback capacitor C_F to make it a *bandpass filter*. This AC coupled topology makes it easy to apply high-frequency gain while maintaining the same DC shift. In particular, ideally

(i) the *zero* at DC (from C_I) removes $v_{\text{smooth,low}}$ (i.e., signals up to 35 Hz)

(ii) the *pole* from $R_I C_I$ passes mid-frequency signals in $v_{\text{smooth,pass}}$ and amplifies them by $-R_F/R_I$

(iii) the *pole* from $R_F C_F$ removes high frequency signals in $v_{\text{smooth,high}}$ (i.e., signals above 15 kHz

(iv) a 5 V DC offset is added by the R_B–R_B divider[12]

So components should be chosen so that

$$\frac{R_F}{R_I} \approx 2 \quad \text{and} \quad \frac{1}{2\pi R_I C_I} \approx 35\,\text{Hz} \quad \text{and} \quad \frac{1}{2\pi R_F C_F} \approx 15\,\text{kHz} \quad \text{and} \quad R_B \approx 2R_F.$$

For example, pick $R_F \approx 22\,\text{k}\Omega$, $R_I \approx 10\,\text{k}\Omega$, $C_F \approx 470\,\text{pF}$, $C_I \approx 470\,\text{nF}$, and $R_B \approx 10$–$50\,\text{k}\Omega$ where the R_B–R_B divider is a potentiometer which should be **tuned** to ensure a $5\,\text{V}_{\text{DC}}$ output offset.

[11] *Overdriving* an amplifier means to drive its output signal to the point of clipping. Musicians will often purposely overdrive their amplifiers (or preamplifiers) as a sort of nonlinear filtering that adds harmonics that can be pleasing to the human ear.
[12] Use the *superposition theorem* to verify this statement.

Current Driver

When the voltage amplifier is unloaded (e.g., **for testing**), the *current driver* in Figure 4.2 is not needed. That is, the current driver can be implemented with a short circuit, as in Figure 4.3.

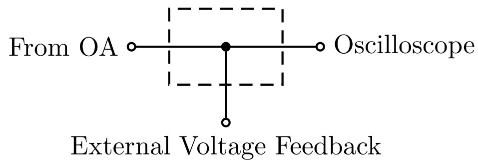


Figure 4.3: Unloaded current driver (**for testing only**).

While this "current driver" can be used when testing the output of the voltage amplifier, it will not be able to drive the 8 Ω speaker load at the required 0.25 W output. Hence, a current amplifier needs to be built into the driver with voltage feedback to ensure good linearity, as in the two following examples.

Current Amplifier and Internal Voltage Feedback

The current amplifier can be put in the feedback path of an operational amplifier buffer, as in Figure 4.4.

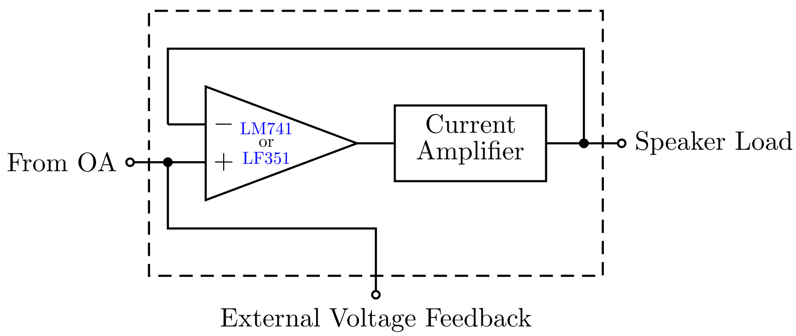


Figure 4.4: Current driver with internal voltage feedback (i.e., cascaded amplifier case).

To simplify testing, construction, and amplifier trimming, **we will use this configuration in the lab**. Notice that the feedback network *does not load the current amplifier*, so no quiescent current is "wasted."

Current Amplifier and External Voltage Feedback

The external voltage feedback can be taken directly from the output, which removes the need for the buffer circuit. The result in Figure 4.5 puts the current amplifier in the feedback path of the *voltage amplifier*.

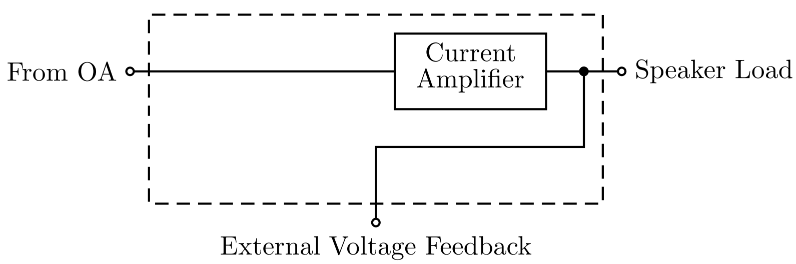


Figure 4.5: Current driver with external voltage feedback (i.e., integrated amplifier case).

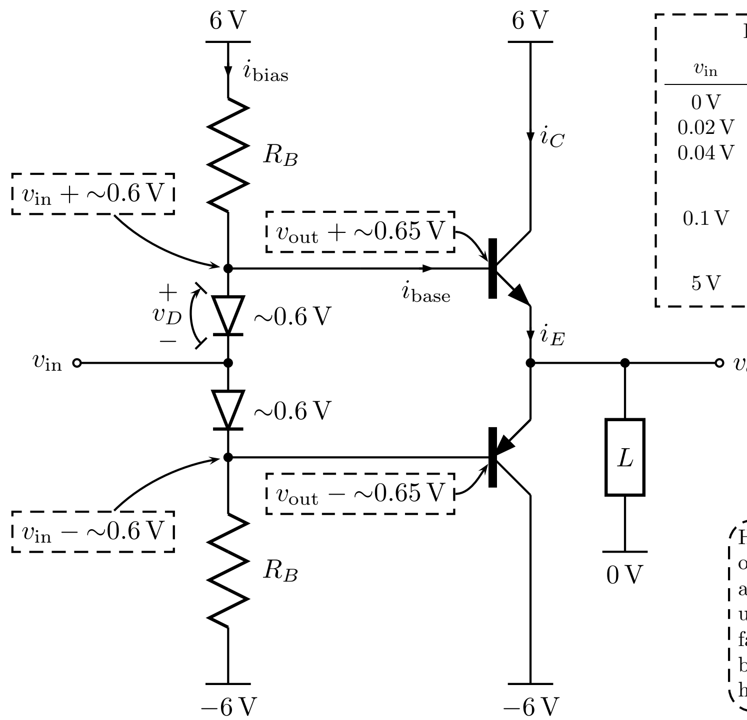
This configuration loads the current amplifier with the feedback network, which complicates trimming and wastes quiescent current (i.e., reduces class-A operating range). **We will not use this configuration in the lab.**

Current Amplifier

Current amplification is needed to drive large currents into the $8\,\Omega$ speaker without loading the voltage amplifier. The approach builds upon a bipolar junction transistor (BJT) emitter follower[13], a buffering circuit that provides roughly unity voltage gain even when driving heavy loads (i.e., low load impedance).

Class-B Push–Pull Amplifier Revisited

The *biased push–pull amplifier* from the BJT lab is shown in Figure 4.6 (with double-ended supply rails).



For $R_B \approx 5.4\,\text{k}\Omega$, $L = 1\,\text{k}\Omega$, and $\beta \approx 100$:

v_{in}	i_{bias}	v_D	i_{base}	i_E	v_{out}
0 V	1 mA	0.6 V	0 mA	0 mA	0 V
0.02 V	1 mA	0.6 V	0 mA	0 mA	0 V
0.04 V	0.99 mA	0.6 V	0 mA	0 mA	0 V
			⋮		
0.1 V	0.98 mA	0.6 V	0.5 μA	0.05 mA	0.05 V
			⋮		
5 V	0.1 mA	0.46 V	48 μA	4.8 mA	4.8 V

Here, we assume that the base–emitter diode drop of the each output transistor is 0.65 V. However, as the transistors dissipate power, they will heat up and require less excitation for conduction. In fact, thermal runaway can be a problem. A typical base–emitter diode drop for a power amplifier (i.e., hot transistors) is 0.5 V.

Figure 4.6: Class-B Biased Push–Pull Amplifier with load L and $|v_{\text{in}}| < {\sim}5.4\,\text{V}$.

All parts of a *class-A amplifier* are conducting at all times. In this circuit, one transistor is in cutoff at any time, which makes it a *class-B amplifier*. By reducing the number of conducting components, the device

(i) reduces power dissipation and increases efficiency but

(ii) *increases* distortion.

When one transistor becomes active, the device behaves as either an NPN emitter follower or a PNP emitter follower[14]. The biasing diodes reduce the amount of input swing away from 0 V needed to bring a transistor out of cutoff. Adding these biasing diodes reduces the distortion introduced by class-B operation; however, they require additional power dissipation. The additional always-on power dissipation places this amplifier's mode of operation in between class-A and class-B on the Pareto optimal power–distortion tradeoff curve.

Because of the negative feedback wrapped around the current amplifier, the impact of the amplifier's distortion can be reduced significantly. However, any additional reduction of distortion in the amplifier will improve the performance of the negative feedback. Additionally, because our aim is to prevent severe loading of the voltage amplifier, increases in current gain and input impedance are welcome.

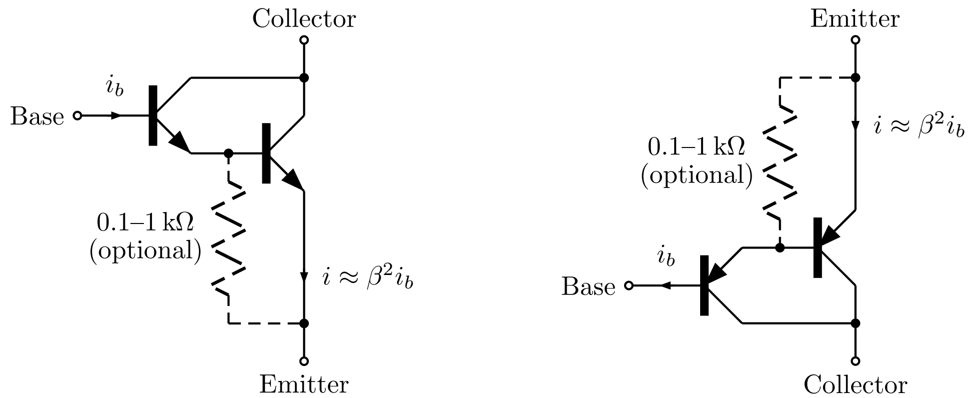
[13] An emitter follower is also known as a *common collector* circuit.

[14] Remember that an **NPN** BJT is **N**ot **P**ointing i**N** and a **PNP** BJT **P**oints i**N** **P**roudly.

Increasing Transistor Current Gain: Darlington and Sziklai Pairs

In our transistor models, for a base current i_b, we assume that collector current is βi_b, where the current gain $\beta \gg 0$. Having a high current gain increases the input impedance of the transistor, so little input current is needed for large output currents. It is beneficial for us to increase our effective β as far as possible so that we can ensure large output currents will still require negligible base currents.

The Darlington Pair: Consider the *Darlington*[15] *transistor pairs* shown in Figure 4.7.



(a) *npn* configuration. (b) *pnp* configuration.

Figure 4.7: Darlington transistor pairs. *Optional* small resistor increases bandwidth.

In Figure 4.7(a) the leftmost transistor drives current into the base of the rightmost transistor. Therefore, the current gains of the two transistors multiply. The pair of transistors combine to make a device that is nearly identical to a single *npn* transistor except that the current gain is β^2 instead of β. In other words,

- the device in Figure 4.7(a) acts like a "big" *npn* transistor, and
- the device in Figure 4.7(b) acts like a "big" *pnp* transistor.

Because $\beta \geq 100$ for most signal transistors, the β^2 gain can greatly improve the performance of our amplifier. Unfortunately, the Darlington pair has several drawbacks:

Slower response: There can be a delay between the deactivation of the "master" (i.e., leftmost) transistor and the deactivation of the "slave" (i.e., rightmost) transistor. To improve turn-off time and, a *small* resistor can be placed across the "slave" transistor's base and emitter. The resistor must be small so that the leakage current from the "master" transistor does not activate the "slave" transistor. Leakage currents are small (e.g., 20 nA), and so resistors as large as 1 kΩ are safe to use (i.e., $20\,\text{nA} \times 1\,\text{k}\Omega \ll 0.6\,\text{V}$).

Increased drop: The "base"-"emitter" drop of this new device is the sum of the base–emitter diode drops of the two component transistors. Thus, depending on temperature, a typical Darlington pair will have 1.0–1.4 V between its "base" and "emitter" when active.

Thermal runaway: While it provides high current gain, it does nothing to prevent thermal runaway problems associated with high-current applications.

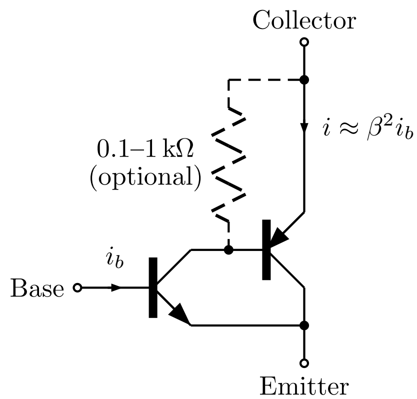
Saturation: The device saturates at a whole diode drop (i.e., ∼0.7 V) instead of the usual 0.2 V.

Matching: Matching *npn* transistors to *pnp* transistors is difficult even when designing integrated circuits. However, matching of *complementary* devices is a necessity to reduce distortion in push–pull stages. These mismatches are amplified when using Darlington pairs.

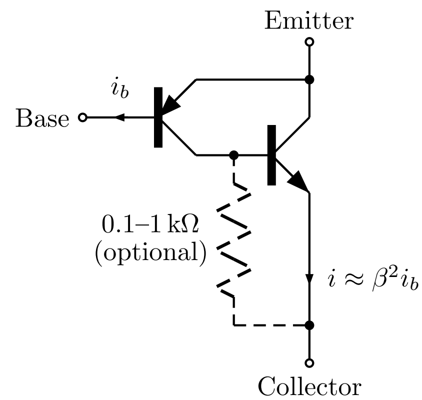
The impact of these problems can be reduced with a slight rearrangement of the pairs, as we will show next.

[15] The *Darlington pair* is named for its inventor Sidney Darlington, who showed how transistors in silicon could share collectors.

The Sziklai Pair: Compare the Darlington pairs with the *Sziklai*[16] *pairs* shown in Figure 4.8.



(a) *npn* configuration.



(b) *pnp* configuration.

Figure 4.8: Sziklai transistor pairs. *Optional* small resistor increases bandwidth.

Like the Darlington configuration, the Sziklai configuration is a compound transistor device that provides an approximate β^2 current gain.

Originally, the Sziklai *pnp* configuration was preferred over the Darlington *pnp* configuration because it used one fewer *pnp* transistor. Good *pnp* transistors were costly and scarce, and so there was a premium on reducing their numbers. Now that *pnp* transistors are readily available, Sziklai pairs are preferred for both the *npn* and *pnp* configurations for several reasons:

Normal drop: When active, the "base"-"emitter" drop of the Sziklai device is equal to a single transistor (e.g., 0.5–0.7 V depending on temperature).

Thermal runaway: In both the Sziklai and Darlington configurations, most current conducts through the "slave" transistor. In the Darlington configuration, thermal compression of the slave's base–emitter drop leads to additional current and additional thermal compression. However, in the Sziklai configuration, the thermal compression of the slave transistor has less impact on the combination as a whole. The Sziklai combination is more immune to thermal runaway.

Matching: Matching *npn* transistors to *pnp* transistors is difficult even when designing integrated circuits. However, because both Sziklai configurations use an *npn* and a *pnp* transistor, Sziklai push–pull stages are more immune to *npn*–*pnp* mismatches.

However, they have some of the same drawbacks as the Darlington:

Slower response: There can still be a delay between the deactivation of the "master" (i.e., leftmost) transistor and the deactivation of the "slave" (i.e., rightmost) transistor. Again, a *small* resistor (e.g., no larger than 1 kΩ) can be placed across the "slave" transistor's base and emitter.

Saturation: The Sziklai device also saturates at a whole diode drop (i.e., ∼0.7 V) instead of the usual 0.2 V.

In our push–pull stage, we

(i) **replace the top *npn* transistor with the configuration in Figure 4.8(a) and**

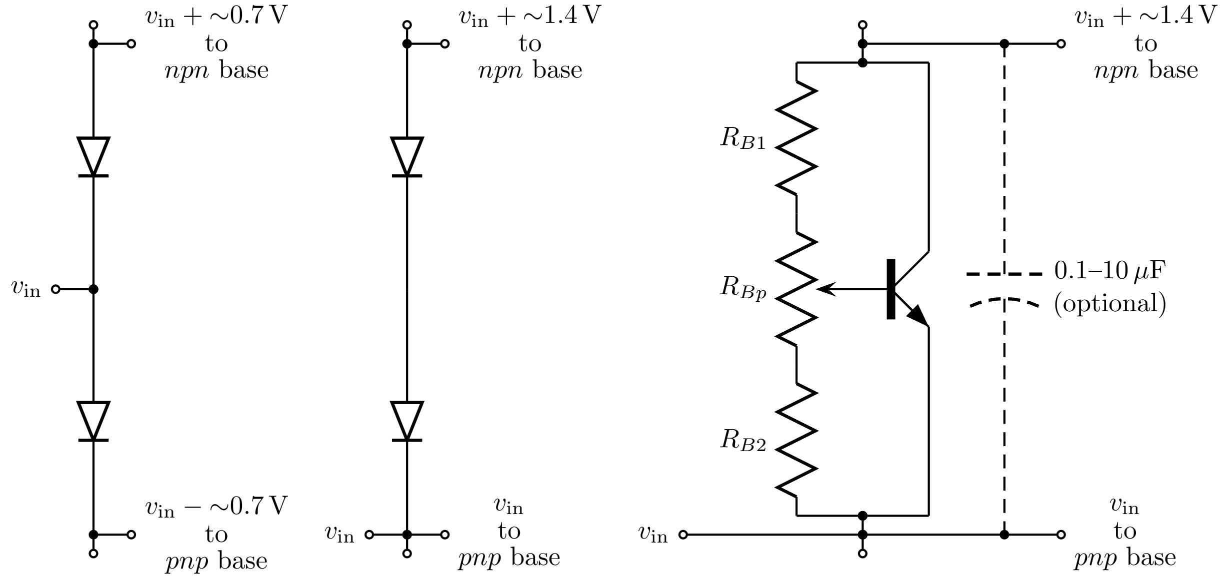
(ii) **replace the bottom *pnp* transistor with the configuration in Figure 4.8(b).**

This design provides high current gain, simple biasing, and less risk of thermal instability.

[16] The *Sziklai pair* is named for its inventor George C. Sziklai.

Rubber Diode for Class-AB Operation

The transistors in the push–pull amplifier in Figure 4.6 are biased into operation with two diodes, as in Figure 4.9(a) with ∼0.7 V diode drop, so that the bases are always approximately 1.4 V apart.



(a) Original biasing. (b) Shifted biasing. (c) Rubber diode ($R_B \leq 330\,\Omega$).

Figure 4.9: Evolution of rubber diode (i.e., base–emitter multiplier).

The alternate biasing circuits in Figure 4.9(b) and Figure 4.9(c) also hold the transistor bases apart by ∼1.4 V. However, the output of the push–pull stage will be shifted up by approximately 0.7 V, which will be automatically corrected by the negative feedback from the voltage amplifier.

There are two major problems with using diode biasing:

Non-constant biasing: Without a constant current source in series with the diodes, the current through them will be modulated by the input signal. Because the diodes are not ideal, the change in current will modulate the separation between the two transistors. This modulation appears as a new component of the input signal, and so the output gets slightly distorted. This distortion is reflected in the table that is included with Figure 4.6; the output waveform is not only clipped but is compressed by this effect.

Programmability: Ideally, each diode drop will be equal to the base–emitter diode drops of each of the output transistors. This matching prevents the *crossover distortion* that we explored in the BJT lab with the simple push–pull amplifier (without diode biasing). Unfortunately, the diode drops will not be matched, and we our only avenue to tuning the separation between the two bases is by adding more diodes. Fine-grained control of the base separation is not possible.

To alleviate both problems, a "rubber diode" circuit like Figure 4.9(c) is used. It has its own negative feedback that maintains constant base separation even as the current through the device varies[17]. Thus, it operates like an ideal diode. The optional capacitor ensures both output transistors see AC signals of interest immediately.

By tuning the potentiometer, the transistor base separation can be finely adjusted. We separate them by slightly more than two ***hot*** base–emitter drops (e.g., 2×∼0.5 V) so that they are never in cutoff *simultaneously* (i.e., *class-B* clipping is prevented because *both* transistors are on even for a 0 V output). Because the amplifier stays in *class-A* operation for small current outputs and moves into *class-B* operation for large outputs, the amplifier is a "*class-AB*" device.

[17] Of course, current through the device must be greater than critical threshold $({\sim}0.7\,\text{V})/R_{\text{base-to-emitter}}$ for it to activate.

Components for Single-Ended Biasing

The complete biasing circuit, with single-ended supply rails, is shown in Figure 4.10.

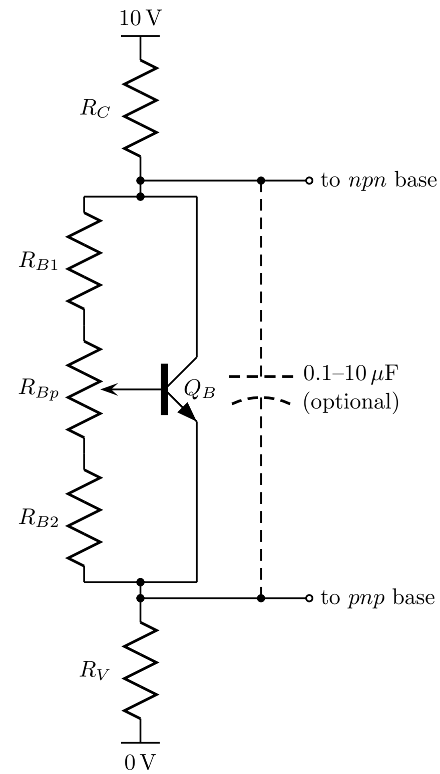


Figure 4.10: Rubber diode with biasing components for single-ended power supply.

Choosing the resistors to populate this circuit is not trivial. The resistor R_C must be chosen small enough to drive both the rubber diode and the *npn* drive current, but it must be chosen large enough to limit the current through the R_V resistor. The rubber diode components, R_{B1}, R_{Bp}, and R_{B2}, must be chosen so that the potentiometer R_{Bp} can adjust the rubber diode potential properly. Additionally, the sum of these three resistances should be chosen large enough to limit the minimum diode current but small enough to prevent too much conduction through the transistor. Here, we describe a few of the details behind these challenging component choices.

Setting quiescent current: First, we set the *quiescent current (QC)* i_Q, which is the *class-A* operating current of the push–pull output stage of the amplifier. Even when there is no load current, at least this much current flows through the output stage. At the point where the load current reaches $2 \times i_Q$, one transistor cuts off and the device enters *class-B* mode. We choose $2\,\text{mA} \leq i_Q \leq 3\,\text{mA}$ and set $1\,\Omega \leq R_E \leq 10\,\Omega$, where R_E is the value of an emitter resistor attached to each transistor in the output stage. We discuss the function of these emitter resistors later. In short, choosing larger values of R_E improves stability, and choosing larger values of i_Q increases class-A range; however, power efficiency and total possible output swing are reduced in both cases. We set i_Q by tuning the rubber diode voltage to between

$$2 \times \overbrace{0.5\,\text{V}}^{\text{Hot } V_{\text{base–emitter}}} + \overbrace{2\,\text{mA}}^{i_{Q,\min}} \times \overbrace{2\,\Omega}^{2R_{E,\min}} = 1.004\,\text{V}$$

and

$$2 \times \overbrace{0.7\,\text{V}}^{\text{Cool } V_{\text{base–emitter}}} + \overbrace{3\,\text{mA}}^{i_{Q,\max}} \times \overbrace{20\,\Omega}^{2R_{E,\max}} = 1.46\,\text{V},$$

which provides a maximum $\pm 3.111\,\text{V}$ sinusoidal swing on the output (i.e., 605 mW through an $8\,\Omega$ speaker).

Sourcing enough current at all times: As the input swings up, the "top" of the "rubber diode" moves closer to the 10 V supply rail, and the current through R_C decreases. Eventually the current in R_C will not be sufficient to drive both the *npn* output transistor and the biasing circuit. So to guarantee undistorted output, we assume a potential drop across R_C of at least 1 V. We will show that the biasing circuit needs to regulate ∼1.46 V across it and the maximum absolute output to the 8 Ω speaker is ∼3.111 V. Hence,

$$\underbrace{\frac{10\,\text{V} - 1.46\,\text{V}}{R_V + R_C}}_{\substack{\text{Minimum available}\\\text{bias current}}} \geq \underbrace{\frac{1\,\text{V}}{R_C}}_{\substack{\text{Minimum}\\R_C\text{ current}}} \geq \underbrace{\frac{1.46\,\text{V}}{R_{B1} + R_{Bp} + R_{B2}}}_{\substack{\text{Minimum}\\\text{rubber diode current}}} + \underbrace{\left(\frac{3.111\,\text{V}}{8\,\Omega}\right)\frac{1}{\beta^2}}_{\substack{\text{Maximum}\\npn\text{ drive current}}} \tag{4.2}$$

where transistor current gain $\beta \approx 100$.

Preventing low-end clipping: As shown below, the input driving the circuit will only be able to *source* current to the R_V resistor (i.e., the output "rests" low and rises with additional R_V current). So the lowest possible output signal is set by the biasing current through the R_V resistor, but R_V cannot be made too small without drawing too much current from the input. As with the top of the rubber diode, we assume the bottom of the rubber diode comes no closer than 1 V from the adjacent supply rail. That is,

$$\underbrace{1\,\text{V}}_{\substack{\text{Minimum}\\\text{signal}}} \geq \left.\underbrace{(10\,\text{V} - 1.004\,\text{V})\frac{R_V}{R_V + R_C}}_{\substack{\text{“Resting” } R_V \text{ potential from biasing}\\\text{(i.e., “no-input” potential)}}} + \underbrace{\left(\frac{3.111\,\text{V}}{8\,\Omega}\right)\frac{1}{\beta^2}R_V}_{\substack{R_V\text{ potential from}\\\text{load sinking}}}\right\} \substack{\text{Largest “no-input”}\\\text{potential drop}\\\text{across } R_V} \tag{4.3}$$

where transistor current-gain $\beta \approx 100$.

Choosing diode components: To set the quiescent current, potentiometer R_{Bp} is adjusted so that the rubber diode maintains a potential difference near $2 \times V_{BE,\text{hot}} + i_Q \times 2R_E$. That is, the rubber diode must be tuned to 1.004–1.46 V. If the minimum and maximum Q_B base–emitter drop is are 0.6 V and 0.7 V,

$$(0.6\,\text{V})\left(1 + \frac{R_{B1} + R_{Bp}}{R_{B2}}\right) \gg \overbrace{1.46\,\text{V}}^{\substack{\text{Cool output}\\\text{with } R_E = 10\,\Omega}} \quad \text{and} \quad (0.7\,\text{V})\left(1 + \frac{R_{B1}}{R_{Bp} + R_{B2}}\right) \ll \overbrace{1.004\,\text{V}}^{\substack{\text{Hot output}\\\text{with } R_E = 1\,\Omega}}. \tag{4.4}$$

Hence, the potentiometer gives plenty of tuning range around the ideal values. Choosing large R_{B1}, R_{Bp}, and R_{B2} allows for the greatest freedom in picking R_V and R_C, but increasing the size of these resistors also increases the conduction and heat dissipation through transistor Q_B.

Component choices: To meet Equations (4.2), (4.3), (4.4), and keep $R_V \geq 45\,\Omega$ (to protect input), set

$$\begin{aligned} R_C &= 820\,\Omega \\ R_{B1} &= 330\,\Omega \text{ (or } \tfrac{1\,\text{k}\Omega}{3}\text{)} \\ R_{Bp} &= 500\,\Omega \\ R_{B2} &= \tfrac{1\,\text{k}\Omega}{2} \\ R_V &= 100\,\Omega \end{aligned} \quad \text{or} \quad \begin{aligned} R_C &= 1\,\text{k}\Omega \\ R_{B1} &= 330\,\Omega \text{ (or } 470\,\Omega\text{)} \\ R_{Bp} &= 1\,\text{k}\Omega \\ R_{B2} &= 680\,\Omega \\ R_V &= 100\,\Omega \text{ (or up to } 120\,\Omega\text{)} \end{aligned} \quad \text{or} \quad \begin{aligned} R_C &= 820\,\Omega \\ R_{B1} &= 330\,\Omega \text{ (or } \tfrac{1\,\text{k}\Omega}{3}\text{)} \\ R_{Bp} &= 500\,\Omega \\ R_{B2} &= 560\,\Omega \\ R_V &= 100\,\Omega \end{aligned} \tag{4.5}$$

based on *component availability*. Component choices assume worst-case combination of a rubber-diode base–emitter drop of 0.6–0.7 V and push–pull output transistor base–emitter drop of 0.5–0.7 V. For the specific case of a hot output stage with 0.5 V base–emitter drop, we can use

$$\begin{aligned} R_C &= 1\,\text{k}\Omega \\ R_{B1} &= 220\,\Omega \\ R_{Bp} &= 500\,\Omega \\ R_{B2} &= 470\,\Omega \\ R_V &= 100\,\Omega \end{aligned} \quad \text{or} \quad \begin{aligned} R_C &= 1\,\text{k}\Omega \\ R_{B1} &= 180\,\Omega \\ R_{Bp} &= 500\,\Omega. \\ R_{B2} &= 680\,\Omega \\ R_V &= 100\,\Omega \end{aligned} \tag{4.6}$$

Emitter Follower for Buffering and Offset

As discussed, the R_V resistor must be small to prevent low-end clipping. So R_V is a significant load on the input. For example, in order for the output to move up 6 V above its "resting" (ie, *quiescent*) state, the input must provide an additional $(6\,\text{V})/R_V$ of current through the R_V resistor. For $R_V = 100\,\Omega$, such a 6 V swing would require 60 mA of additional current, which is far more than any small signal components (e.g., operational amplifiers) can provide. Therefore, we need to buffer the input with an emitter follower, as shown in Figure 4.11.

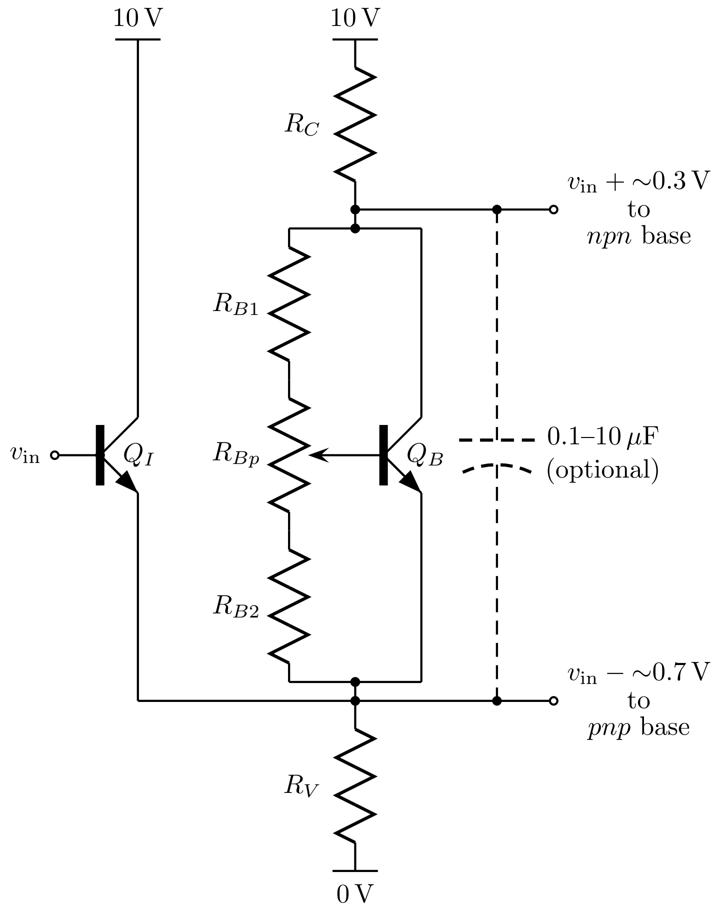


Figure 4.11: Complete biasing circuit. Values are shown in Equations (4.5) and (4.6).

Q_I provides a current gain of $\beta \approx 100$, and so 60 mA of additional R_V current translates to only 0.6 mA of current from the input. The 2N3904 used for Q_I can dissipate no more than 625 mW of power and can carry no more than 200 mA of collector current. The collector current and power dissipation,

$$i_{\text{collector}} \triangleq \frac{V_{\text{emitter}}}{R_V} \qquad \text{and} \qquad P_{Q_I} \triangleq (10\,\text{V} - V_{\text{emitter}})\, i_{\text{collector}} = (10\,\text{V} - V_{\text{emitter}})\,\frac{V_{\text{emitter}}}{R_V},$$

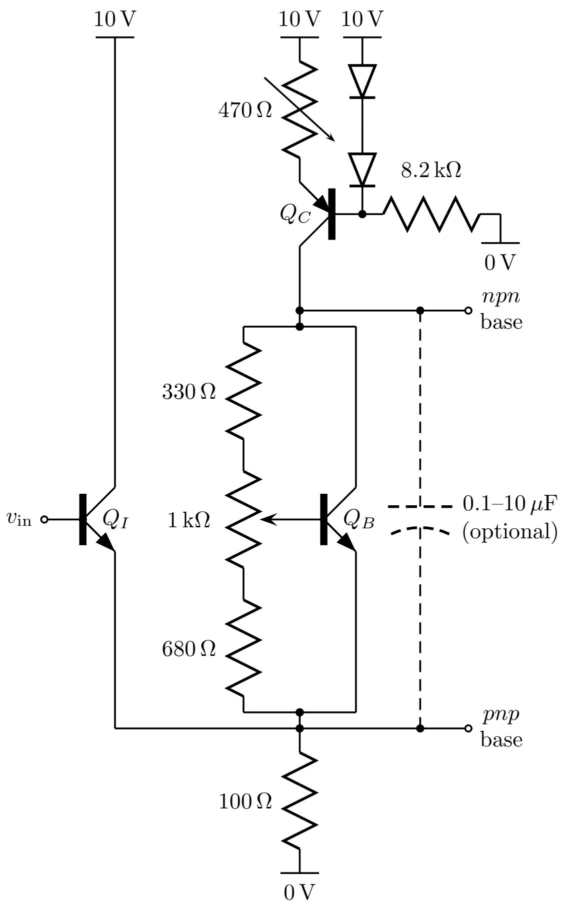
are maximized when $V_{\text{emitter}} \approx 9\,\text{V}$ and $V_{\text{emitter}} = 5\,\text{V}$, respectively. So setting $R_V \geq 45\,\Omega$ protects Q_I.

Output offset: If the rubber diode regulated by Q_B was tuned to ∼1.4 V (i.e., 2 *room-temperature* base–emitter drops), the offset provided by the new emitter follower would almost completely nullify any offset introduced by injecting the signal at the bottom of the rubber diode. However, if the rubber diode must be tuned to ∼1.0 V (i.e., 2 *hot* base–emitter drops), there will still be an offset on the output of −0.2 V. Fortunately, the negative feedback wrapped around this current amplifier will remove any offset. Of course, any residual offset will have little impact on the system as a whole because the output is AC (i.e., capacitively) coupled to the speaker. That is, no DC offsets are transmitted to the speaker.

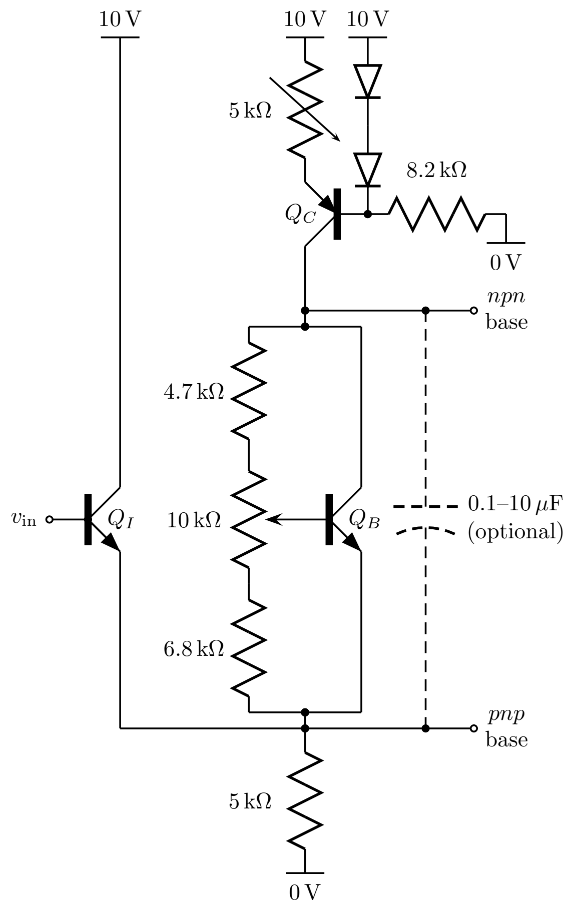
Alternate Biasing Scheme: Current Source

The design in Figure 4.11 provides little freedom for component choice. We can improve the performance and efficiency of our circuit and give more component choice freedom by replacing the R_C resistor with a "smarter" current source that delivers exactly the maximum current needed for the rubber diode and the *npn* output transistor base. Because no extra current is needed, R_V can be increased safely, which increases the impedance of the biasing circuit.

Consider the refined circuits in Figure 4.12.



(a) Current source in place of R_C.



(b) High-impedance biasing.

Figure 4.12: Using a current source for biasing.

The circuit in Figure 4.12(a) uses the same component values as Figure 4.11, but resistor R_C has been replaced with a ~1 mA current source (i.e., a "dynamic resistor" Q_C). The constant current from the source is sufficient to deliver enough base current to the *npn* output transistor while also keeping the rubber diode active at 1.46 V. Because very little current needs to be conducted through Q_B to for the rubber diode regulation, there is far less power dissipated in the biasing for this circuit.

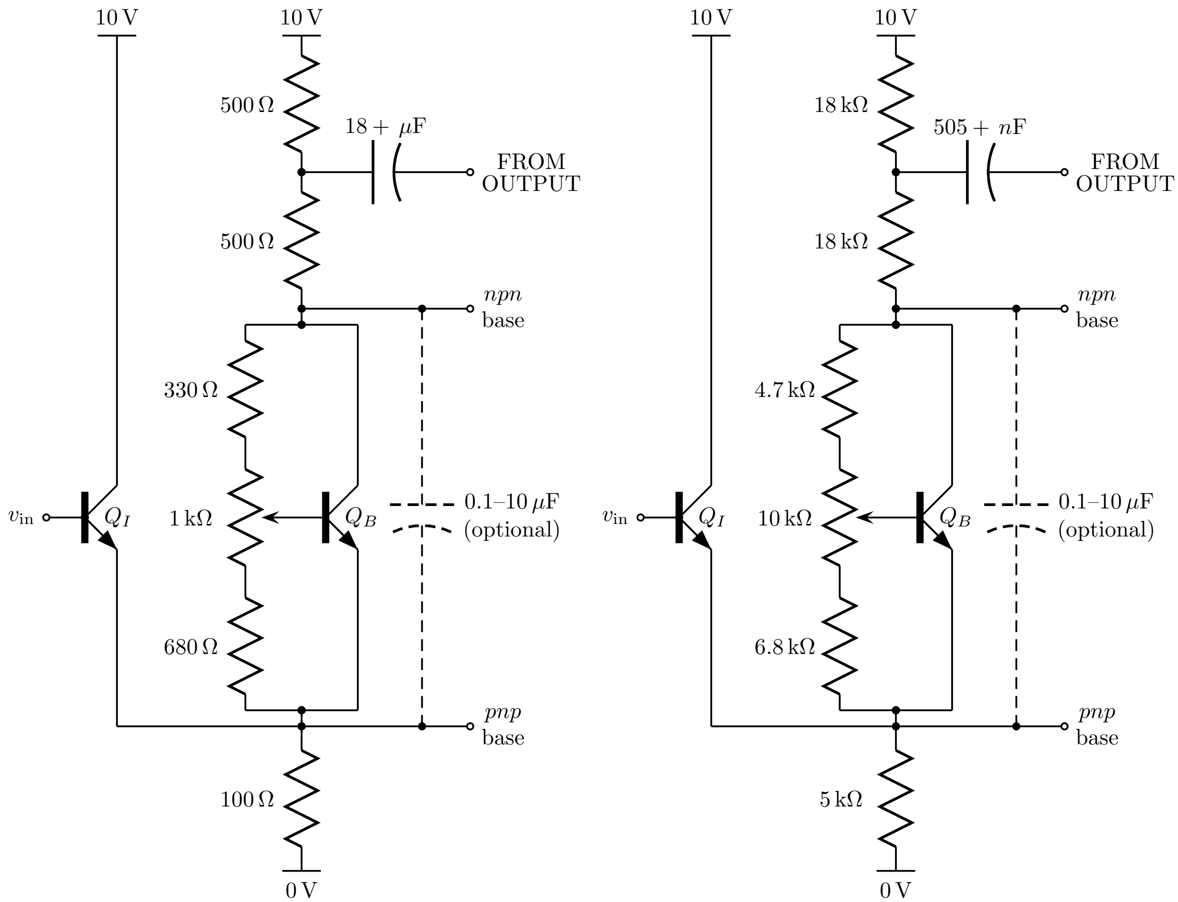
Using a current source gives us more component-choice freedom. In Figure 4.12(b), a ~0.1 mA (i.e., ~0.0001 A) current source is used. The rubber diode resistors have been increased substantially, and so only 0.1 mA provides sufficient drive to the *npn* output transistor and the rubber diode. Because the biasing current has been decreased so much, the R_V resistance can be increased. In this circuit, the emitter follower will never need to source more than 2 mA, which is so small that **the emitter follower can be removed and replaced with a simple diode** (for offset matching)[18].

[18] Recall that an ideal current source has an infinite impedance (i.e., $\Delta v/\Delta i = \Delta v/0 = \infty$). This property is the magic that gives us more control over the input impedance.

Alternate Biasing Scheme: Bootstrapping for Current Source Effect

Another way to provide constant current *at signal frequencies* and greatly increase the input impedance of the biasing circuit is to *bootstrap* the R_C element by a capacitor to the output.

Consider the bootstrapped circuits in Figure 4.13.



(a) Bootstrapped resistor R_C. (b) High-impedance biasing.

Figure 4.13: Bootstrapping R_C as a current source.

The circuit in Figure 4.13(a) uses the same component values as Figure 4.11, but a capacitor from the output is now connected to the "middle" of resistor R_C. By feedback (not shown here), the output has a DC component of 5 V, and so the potential between the two 500 Ω resistors will be between 7.75 V and 7.85 V. Hence, the capacitor will hold a charge of ∼2.75 V. When the input swings up, *both* the output and the *npn* base swing up. As long as the input swing is fast enough, the capacitor will hold a constant 2.15–2.25 V across the bottom 500 Ω resistor. In other words, the 500 Ω resistor *acts* like a ∼4.5 mA current source *to signals of interest* (i.e., to signals of sufficiently high frequency).

The circuit in Figure 4.13(b) is designed to act like the circuit in Figure 4.12(b). That is, the bootstrap capacitor makes the 18 kΩ resistor act like a ∼0.1 mA current source *to signals of interest*. Because the resistances chosen are higher, the minimal size of the capacitor has been reduced.

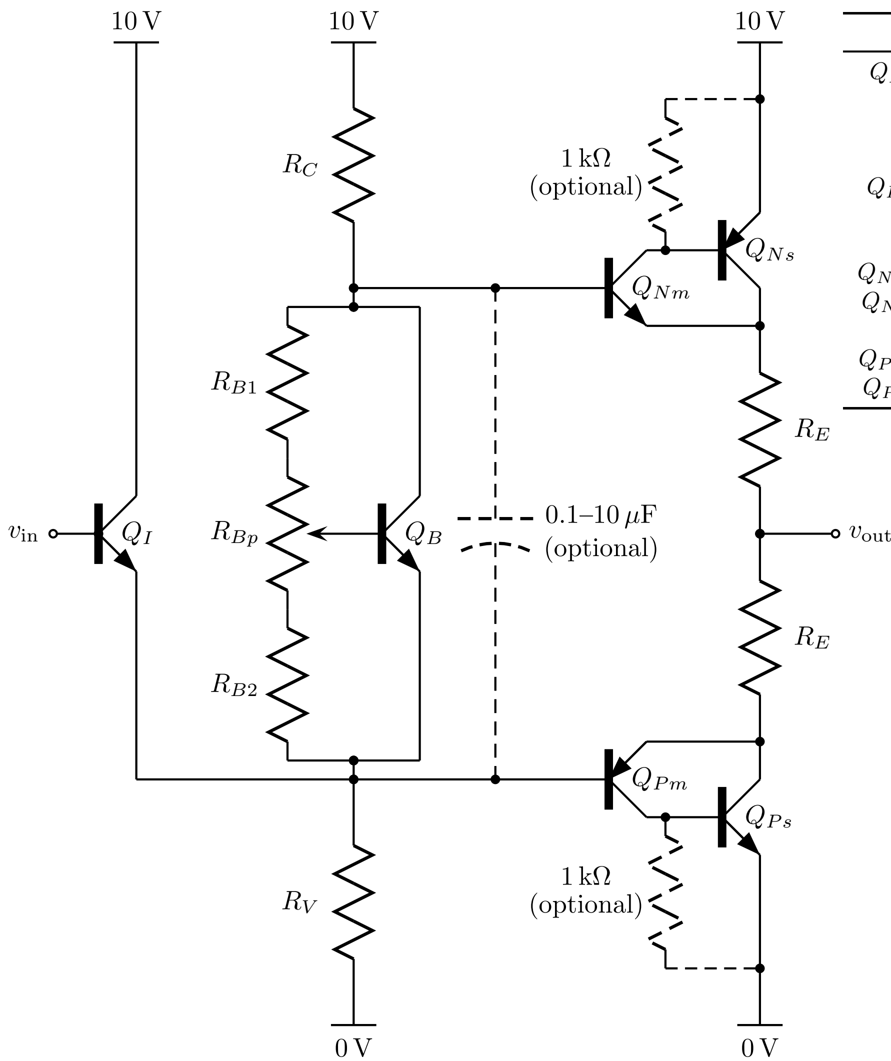
Two important notes:

1. A real current source is almost always preferable over a bootstrapped resistor, especially when low-frequency performance is important.

2. The stored potential in the capacitor can actually cause the top of the capacitor to sit *above* the 10 V rail. Hence, a bootstrapped resistor may allow for larger positive output swings.

The Complete Current Amplifier

The complete current amplifier, using the simplest biasing, is shown in Figure 4.14.



Transistor Legend	
Q_I	Emitter follower for **I**nput buffering and offset adjustment
Q_B	Regulates rubber diode potential drop for **B**iasing
Q_{Nm}	"Master" of *npn* Sziklai pair
Q_{Ns}	"Slave" of *npn* Sziklai pair
Q_{Pm}	"Master" of *pnp* Sziklai pair
Q_{Ps}	"Slave" of *pnp* Sziklai pair

Possible Resistor Choices

Choose a *column*. Find more choices in Equation (4.5) and Equation (4.6).

R_C	820 Ω	820 Ω	1 kΩ
R_{B1}	330 Ω	330 Ω	470 Ω
R_{Bp}	500 Ω	500 Ω	1 kΩ
R_{B2}	$\frac{1\,\text{k}\Omega}{2}$	560 Ω	680 Ω
R_V	100 Ω	100 Ω	100 Ω

R_V can be smaller, but keep $R_V \geq 45\,\Omega$ to protect Q_I.

Figure 4.14: Class-AB current amplifier. **Pick $1\,\Omega \leq R_E \leq 10\,\Omega$ as small as possible.**

Crossover Distortion: To prevent significant crossover distortion, we tune the $500\,\Omega$ potentiometer so that the *unloaded* output stage is always conducting 2–3 mA of *quiescent current*. **For small loads** or small signals that only require 4–6 mA of load current, both Sziklai pairs will conduct simultaneously, just like a *class-A amplifier*. **For large loads** or large signals that require greater than 4–6 mA of load current, one Sziklai pair goes into cutoff mode, just like a *class-B amplifier*. Hence, the amplifier is of *class AB* because it shares characteristics of the other two classes.

Thermal Runaway: The R_E emitter resistors slow the rate of *thermal runaway* by dissipating power from the 2–3 mA of *quiescent current* and providing some voltage feedback. Unfortunately, the voltage dropped across them limits the maximum output swing of the load. To deliver sufficient power to the load, set

$$1\,\Omega \leq R_E \leq 10\,\Omega$$

so that R_E is *as small as possible*. A hot output stage with $R_E = 1\,\Omega$ has a 3.111 V peak output (i.e., 605 mW) while a cool stage with $R_E = 10\,\Omega$ has a 1.467 V peak output (i.e., 135 mW).

Speaker Load and Lag Compensation

THE INSTRUCTOR WILL BRING YOU THE SPEAKER AND COUPLING CAPACITOR. DO NOT CONNECT THE SPEAKER YOURSELF! CHECK WITH THE INSTRUCTOR FIRST!

Because the output of the current amplifier is offset by $5\,\mathrm{V_{DC}}$, it must be AC coupled to the speaker. Applying a $5\,\mathrm{V_{DC}}$ signal to the $8\,\Omega$ speaker will produce no acoustic output, will draw significant current from the amplifier, and can damage the speaker. So the DC signal needs to be removed while all AC signals are transmitted. So a large capacitor is inserted between the current amplifier and the speaker, as shown in Figure 4.15.

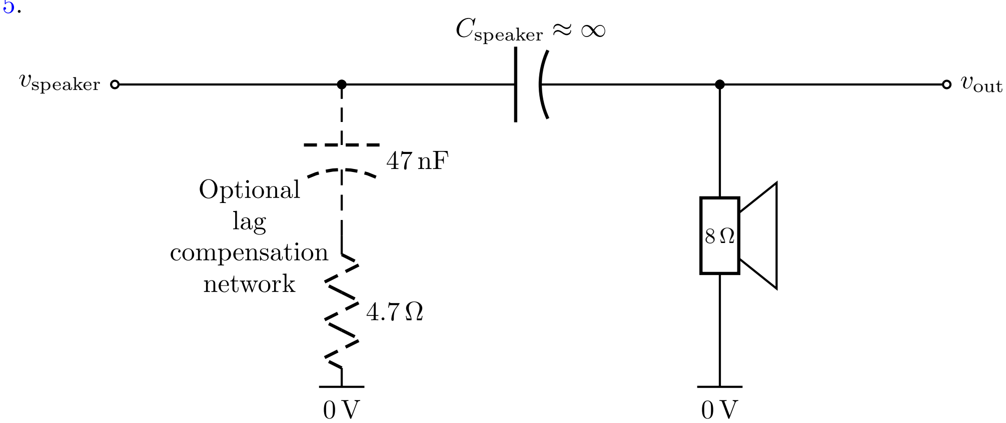


Figure 4.15: Speaker output and optional lag compensation.

The capacitor must be large enough to couple 35 Hz signals into the $8\,\Omega$ speaker without significant attenuation. That is,

$$C_{\mathrm{speaker}} \geq \frac{1}{2\pi(8\,\Omega)(35\,\mathrm{Hz})} \approx 568\,\mu\mathrm{F}.$$

Additionally, the capacitor must be rated for at least 5 V of peak potential difference. A smaller capacitor will reduce the amplitude of low frequency signals reaching the speaker. Because capacitors of this size are scarce in the lab, **YOUR INSTRUCTOR** will most likely use a smaller (e.g., 150–220 μF) capacitor.

Lag Compensation: The 4.7 Ω–47 nF series network should be added if *instabilities* occur (e.g., strange shaped oscillations at a frequency that you don't recognize). It may be omitted otherwise.

Negative feedback causes an output to track a reference signal. When the output rises above the reference signal, the output is reduced by the difference. When the output falls below the reference signal, the output is increased by the difference. In reality, the version of the output that is fed back to the device is delayed. For slow signals, the delay has little effect on the feedback. However, if the signal is fast enough, the delayed feedback can result in *positive feedback*. This effect is related to the *phase margin* of the open-loop system.

Consider steering a boat. Due to rudder effects, there is a delay between your steering and the boat's trajectory. For gentle changes in trajectory, you can easily control the boat. However, if the boat must make an unexpected sharp turn, your delayed control will lead you to apply *too much* control and overshoot the mark. The overshoot leads to even more overshoot in the opposite direction, and the boat oscillates endlessly. In other words, your feedback system (i.e., the rudder and your control) is *unstable*.

Lag compensation provides one mechanism to overcome this problem. To stabilize the system, we must make sure that quickly changing controls are attenuated so that they can't build constructively. In other words, a *filter* must be applied to the control. The filter may add *more* delay (i.e., *lag* compensation), but the accompanying attenuation has the dominant effect (at the cost of speed). The *optional* 4.7 Ω–47 nF series network provides lag compensation for the voltage amplifier's feedback loop. If you notice instability (e.g., strange oscillations at a strange frequency), then add the network, which attenuates (and delays) *unwanted* high frequency signals that are added by feedback (and not by the input).

Procedure

Build and test the voltage amplifier.

1. Build the circuit in Figure 4.16 **BELOW**.
   - **DO NOT** connect input v_{clean} to smoothing filter output v_{smooth} YET.
   - Use *RECEIVER'S* REGULATED 10 V supply rail.
   - Implement the R_B–R_B divider with a single **potentiometer** (to tune the offset).
     - Potentiometer should be between 2 kΩ (code: `202`) and 100 kΩ (code: `104`).
   - Capacitor codes:
     470 nF: `474` (i.e., 470000 pF = 470 nF = 0.47 µF)
     470 pF: `471` (i.e., 470 pF = 0.47 nF = 0.00047 µF)
   - **Bypass capacitors** (e.g., 0.1 µF) at OA's 10 V supply pin can reduce 30 kHz clock noise.

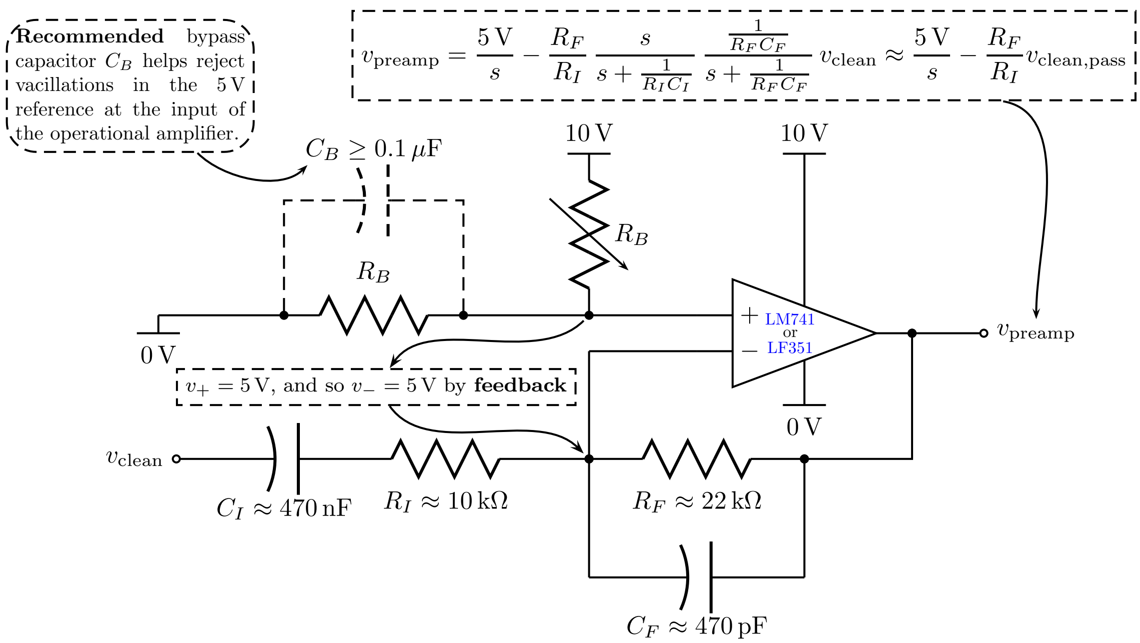


Figure 4.16: Single-ended voltage amplifier with high-frequency pole.

2. Tune your amplifier's offset via R_B–R_B potentiometer.
   - Connect input v_{clean} to 0 V.
   - Measure output v_{preamp} with your digital multimeter (DMM) or oscilloscope.
   - Tune the R_B–R_B divider until the v_{preamp} output is 5 V_{DC}.
   - Disconnect v_{clean} from 0 V.
3. Verify the amplifier has a gain of ∼2 in the passband and attenuates high frequency signals.
   - At input v_{clean}, place 1 V_{pp} sine wave with 0 V_{DC} offset at 100 Hz, 1 kHz, and 15 kHz.
     - May see attenuation at 100 Hz (near lower stopband).
     - No attenuation at 1 kHz (passband).
     - Will see attenuation at 15 kHz (in upper stopband).
   - **SAVE A PLOT** of **input** v_{clean} and **output** v_{preamp} at EACH of the **three** frequencies.
     - Turn averaging **OFF** for these plots.

Build and test the complete audio amplifier. **LEAVE THE AMPLIFIER'S POWER *OFF* AS MUCH AS POSSIBLE** (to prevent *thermal runaway*).

4. Build the *current driver* in Figure 4.17 **BELOW**. Use *RECEIVER'S* REGULATED 10 V supply rail.
   - The LM741 is preferred for the unity-gain buffer because it has a better *phase margin*.
   - The "Current Amplifier" is shown in Figure 4.18, which **slightly DIFFERS from the book!**
     - **BEFORE installing** R_{Bp}, tune to MAXIMIZE Q_B BASE-to-EMITTER resistance.
     - Pick $1\,\Omega \leq R_E \leq 10\,\Omega$ **as small as possible**.
       - Consider **parallel combinations** using 10 Ω, 12 Ω, 15 Ω, and/or 18 Ω.
   - Choices for resistors R_C, R_{B1}, R_{B2}, R_V, and potentiometer R_{Bp} are given in the table.
   - **DO NOT** connect v_{smooth} to v_{signal} YET.
   - **Bypass capacitors** (e.g., 0.1 µF) at **EACH** 10 V supply pin can reduce 30 kHz clock noise.

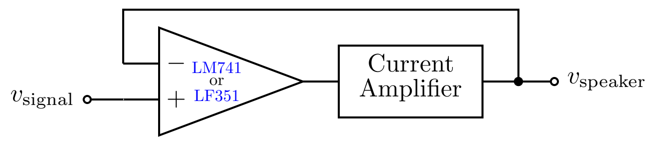
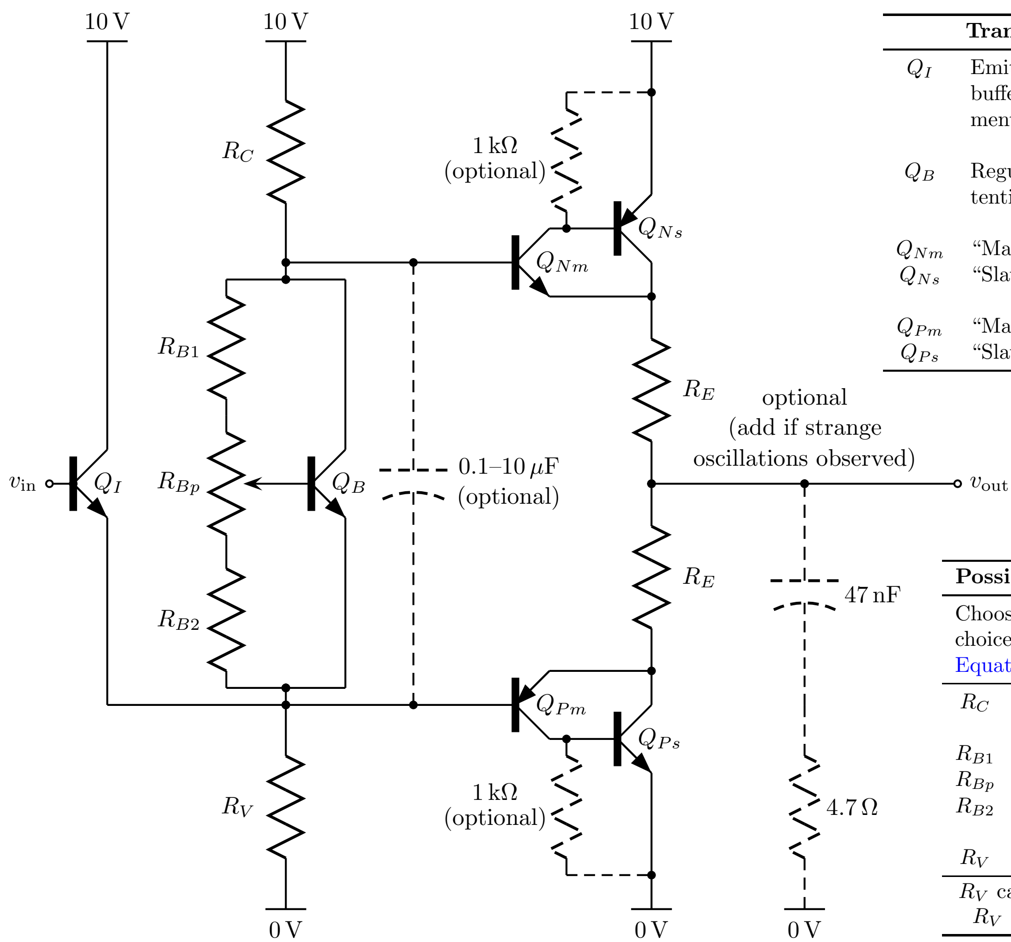


Figure 4.17: Current driver with voltage feedback.



Transistor Legend	
Q_I	Emitter follower for **I**nput buffering and offset adjustment
Q_B	Regulates rubber diode potential drop for **B**iasing
Q_{Nm}	"Master" of *npn* Sziklai pair
Q_{Ns}	"Slave" of *npn* Sziklai pair
Q_{Pm}	"Master" of *pnp* Sziklai pair
Q_{Ps}	"Slave" of *pnp* Sziklai pair

Possible Resistor Choices

Choose a *column*. Find more choices in Equation (4.5) and Equation (4.6).

R_C	820 Ω	820 Ω	1 kΩ
R_{B1}	330 Ω	330 Ω	470 Ω
R_{Bp}	500 Ω	500 Ω	1 kΩ
R_{B2}	$\frac{1\,\text{k}\Omega}{2}$	560 Ω	680 Ω
R_V	100 Ω	100 Ω	100 Ω

R_V can be smaller, but keep $R_V \geq 45\,\Omega$ to protect Q_I.

Figure 4.18: Class-AB current amplifier. **Pick small $1\,\Omega \leq R_E \leq 10\,\Omega$.**

5. For class-AB operation, trim current amplifier (CA) for 2–3 mA of quiescent current (QC).
   - Apply 5 V_{DC} (i.e., half of your 10 V_{DC} supply rail) to the v_{signal} current driver input.
     - If the current amplifier is working properly, output $v_{speaker}$ will match input **EXACTLY**.
     - Manually vary the DC input from 3 V_{DC} to 7 V_{DC}.
       - Use digital multimeter (DMM) or oscilloscope to measure v_{signal} and $v_{speaker}$.
       - For each input, $v_{speaker}$ should match v_{signal} **EXACTLY**. If not, check R_V color codes.
       - Use oscilloscope to view $v_{speaker}$ output, which should be CONSTANT.
         - If output **OSCILLATES** (e.g., blips, ramps, or *big* "noise") for a constant input, add optional RC network (i.e., *lag compensation*) shown in Figure 4.18 from v_{out} to 0 V (if 4.7 Ω is not available, use 10 Ω‖10 Ω‖100 Ω or 12 Ω‖12 Ω or similar).
     - Once you are sure the driver works, set DC input to 5 V_{DC}.
   - **ADJUST THE R_{Bp} POTENTIOMETER** until 2–3 mA flows through the R_E resistors.
     - Measure **voltage** across one of the R_E resistors **with your digital multimeter (DMM)**.
       - Adjust potentiometer until the voltage is between $R_E \times 2$ mA and $R_E \times 3$ mA.
       - **DO *NOT* USE THE OSCILLOSCOPE FOR THIS MEASUREMENT!!**
       - **DO *NOT* use DMM PROBES to make this measurement!**
         - Solidly connect your DMM to your breadboard across one of the R_E resistors.
     - Current amplifier switches from class-A to class-B mode when load draws more than TWICE the QC (i.e., 4–6 mA). In class-A, both Sziklai pairs are active. In class-B, only one pair is.
   - **Turn power OFF & ON.** Use oscilloscope to verify $v_{speaker}$ output is **still** FLAT and THIN.
     - If output is now unstable, add optional RC network shown in Figure 4.18 from v_{out} to 0 V.

6. Connect voltage amplifier v_{preamp} to current amplifier v_{signal}.
   - At voltage amplifier input v_{clean}, generate 1 V_{pp} sine wave with 0 V_{DC} offset and frequency 1 kHz.
   - View v_{clean} and $v_{speaker}$ on the oscilloscope.
     - Set **trigger** `Edge` for v_{clean}, and adjust **level** and **filtering** (see `Mode`) to stabilize display.
     - The output $v_{speaker}$ should be a 2 V_{pp} sine wave at 1 kHz with 5 V_{DC} offset.
   - **SAVE AN UNAVERAGED PLOT** of **input** v_{clean} and **output** $v_{speaker}$.

7. **SEE THE INSTRUCTOR** to attach a speaker to the output (via *polarized* electrolytic capacitor).
   - **KEEP SPEAKER AWAY** from any magnetic media (e.g., floppy disks).
   - Attach the speaker to the current amplifier's output, as shown in Figure 4.19.
     - If using a 4 Ω speaker, add 4–6 Ω (e.g., 12 Ω‖12 Ω‖12 Ω or 10 Ω‖10 Ω) in *series* with the *speaker*.
     - To be practical, use 150 μF ≤ $C_{speaker}$ ≤ 300 μF (e.g., 150 μF or 220 μF are good choices).

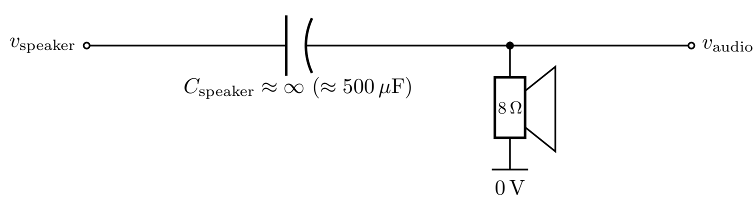


Figure 4.19: Speaker output.

   - Again, apply 1 kHz 1 V_{pp} sinusoid to the input of the *voltage amplifier* — a tone should be *heard.*
   - **SAVE AN *UNAVERAGED* PLOT** of the 1 kHz **input** v_{clean} and **output** $v_{speaker}$.
   - Verify that the magnitude of the signal (i.e., volume) decreases when frequency is increased.
   - **SAVE AN *UNAVERAGED* PLOT** showing a **high-frequency input** and **attenuated output**.
   - Musicians may want to hear 220 Hz, 440 Hz, and 880 Hz through the system.

Integrate the audio amplifier with your **mod**ulator-**dem**odulator (MODEM).

8. Initially, **disconnect the speaker from the output.**
9. Connect the smoothed output v_{smooth} to the voltage amplifier input v_{clean}.
10. Test your complete system.
    - Apply a $1\,\mathrm{V_{pp}}$ sinusoidal wave at $1\,\mathrm{kHz}$ to the **LSA input** (i.e., modulator input).
    - The output of the current amplifier should be an *inverted* and *doubled* version of the LSA *output*.
    - **SAVE AN UNAVERAGED PLOT** showing **LSA output** and **current amplifier output**.
11. **SEE THE INSTRUCTOR** to test your amplifier with real music and a speaker.
    - Connect a music device with a headphone output (e.g., a walkman) to the input of the LSA.
    - Again, attach the speaker and coupling capacitor to your circuit as in Figure 4.19.
    - When power is turned on, you should hear music on the speaker.
    - **SAVE AN *UNAVERAGED* PLOT** showing **LSA output** and **current amplifier output**.
    - For comparison, connect device directly to speaker (capacitor is not needed, but it can be used).

Optional Procedures (BONUS credit)

You can earn up to 10 points for *each* of the following *optional* procedures. Include the optional results *in your report* to receive credit. The amount credit is based on the completeness of your results. **You may complete as many or as few of the following as you wish.**

- *EXPERIMENTALLY* investigate the impact of the optional components shown in Figure 4.14. What happens when you add or remove them? How does the output change? Do your laboratory results match theory? What about when you add the speaker?
- Reconfigure your circuit to use a current-source biasing circuit like the one in Figure 4.12(a). Do you notice any change in your circuit's performance? What about when you add the speaker?
- Reconfigure circuit to use a bootstrapped collector resistor like in Figure 4.13(a). That is,

  (i) replace your single R_C resistor with two $(R_C/2)$ resistors in series.

  (ii) connect a bootstrapped capacitor (i.e., a capacitor from the circuit output) to the junction between the two resistors. The bootstrapped capacitor's capacitance should be at *least*

  $$1/\left(2\pi \frac{R_C}{4} (35\,\mathrm{Hz})\right), \tag{4.7}$$

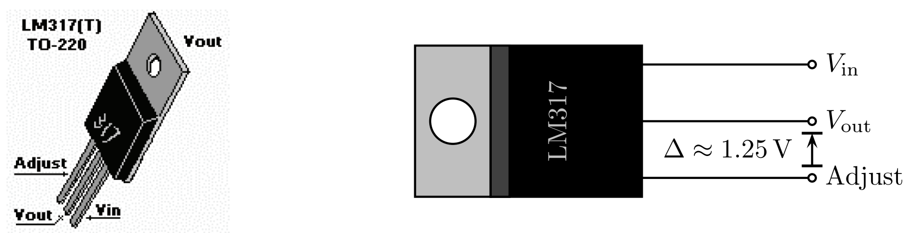
  and the capacitor's cathode (i.e., it's "negative" end) should be connected to the *output*.

  Do you notice any change in your circuit's performance? What about when you add the speaker? What about low frequencies? For additional bonus, derive Equation (4.7) and explain your steps.
- Replace your biasing components with the ones shown in Figure 4.12(b). Do you notice any change in your circuit's performance? Now remove the Q_I transistor and apply your input directly to the top of the $5\,\mathrm{k}\Omega$ resistor at the base of the *pnp* Sziklai pair. Do you notice any difference? What about when you add the speaker?
- Replace your biasing components with the ones shown in Figure 4.13(b). Do you notice any change in your circuit's performance? Now remove the Q_I transistor and apply your input directly to the top of the $5\,\mathrm{k}\Omega$ resistor at the base of the *pnp* Sziklai pair. Do you notice any difference? What about when you add the speaker?

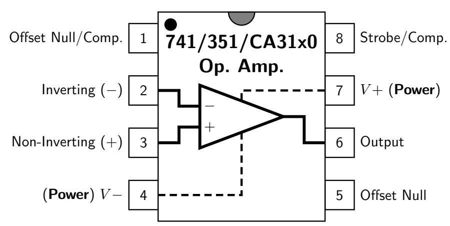
Demolishing Your Circuit

Disassemble your circuit. **Return all parts to their appropriate locations IN THE CABINET. DO NOT place parts in the metal tins at your benches!!** Take your breadboards home.

A Parts

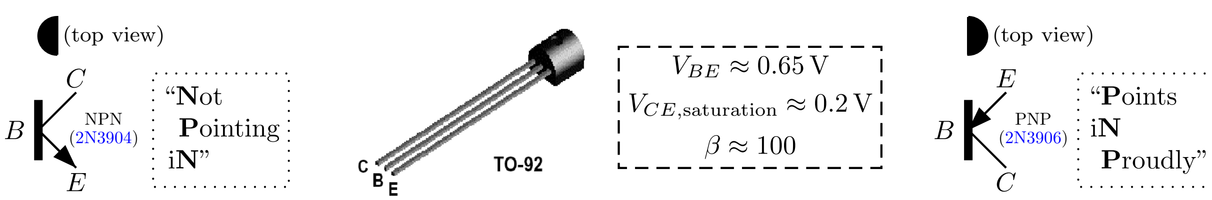


(a) LM317 3-terminal adjustable regulator

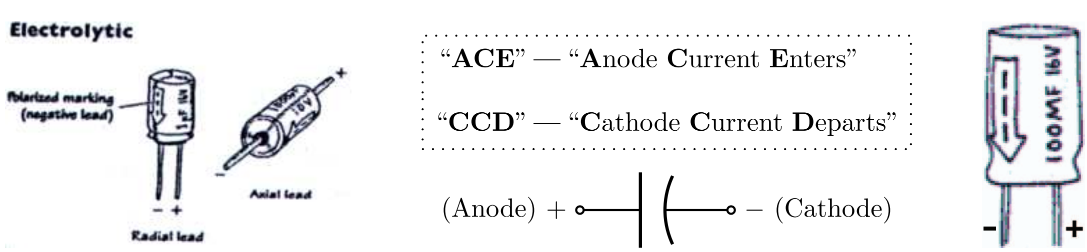


The CA3130 can be used as a replacement for the CA3160. However, capacitance **between** 45 **pF and** 100 **pF** must be added **from pin** 1 **to pin** 8 (for feedback stability compensation).

(b) LM741/LF351/CA3130/CA3160 op. amp.



(c) 2N3904 NPN/2N3906 PNP Bipolar Junction Transistor (BJT)



(d) Electrolytic capacitor

Figure A.1: Part pin-outs.