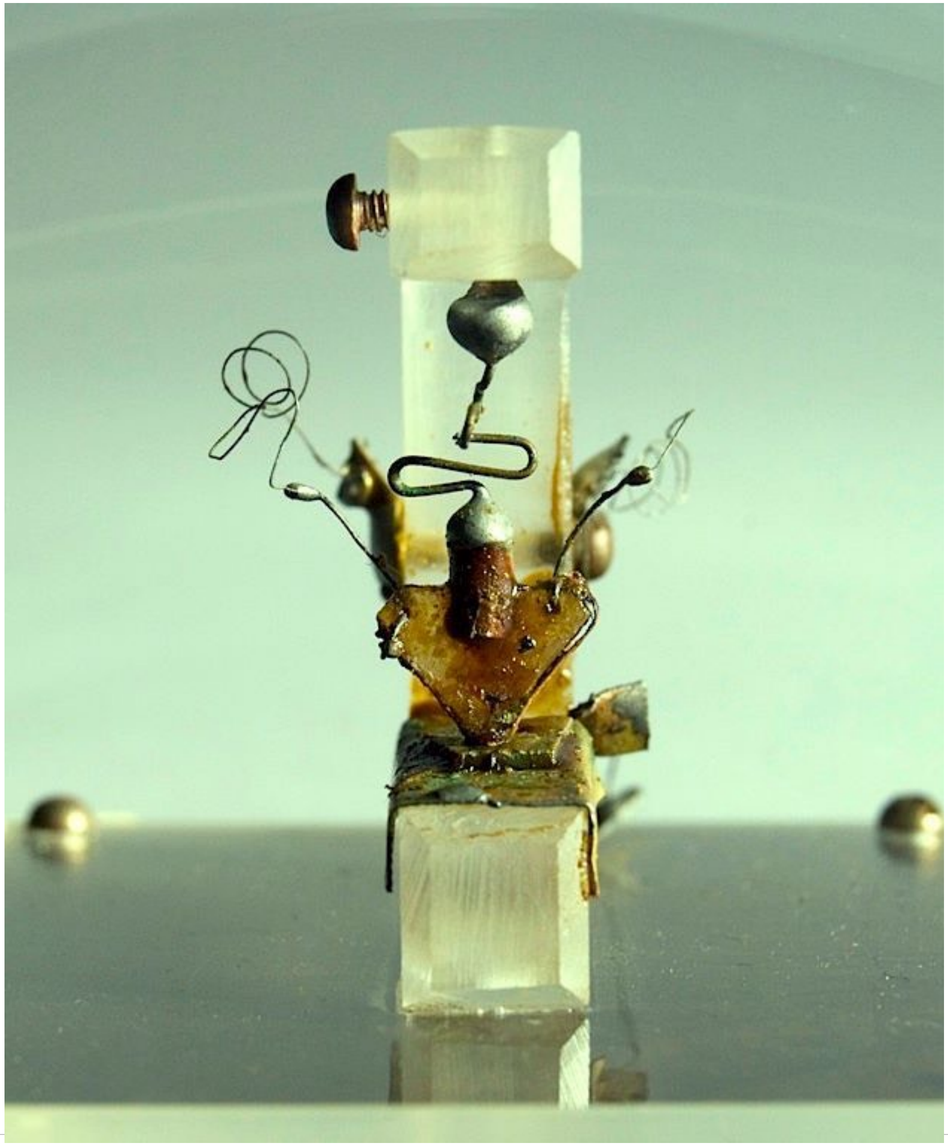
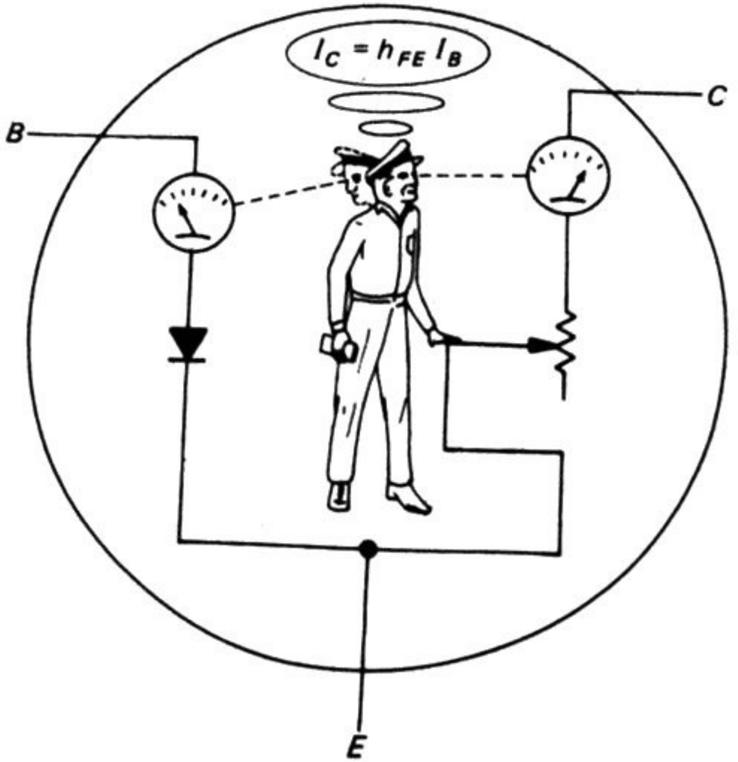
PHYS127AL Lecture 7

David Stuart, UC Santa Barbara

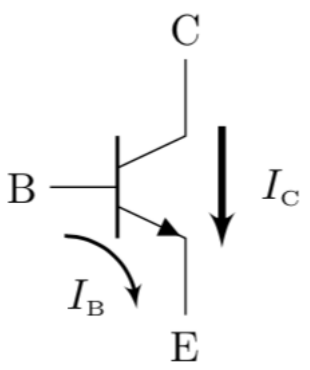
More transistor circuits: current source, PNP, bootstrapping, Ebers-Moll





Review: Transistor rules of operation

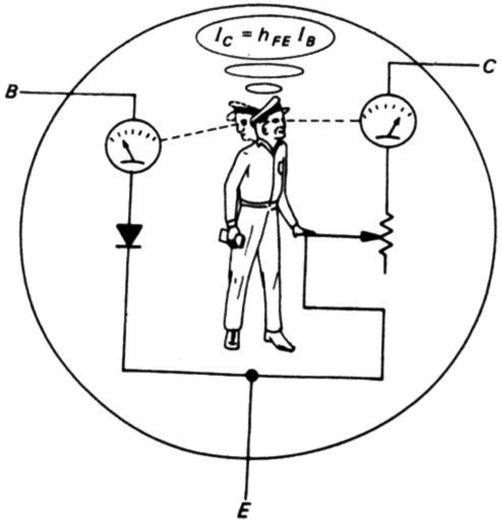
1). $V_{BE} = 0.6$ V or the transistor is off
I.e., $V_B = V_E + 0.6$ V
Once the transistor is on, $\Delta V_B = \Delta V_E$.



2). $I_C = \beta\, I_B$.
And by charge conservation $I_E = I_B + I_C$ so $I_E \cong I_C$

3). $V_{CE} > 0.2$ V

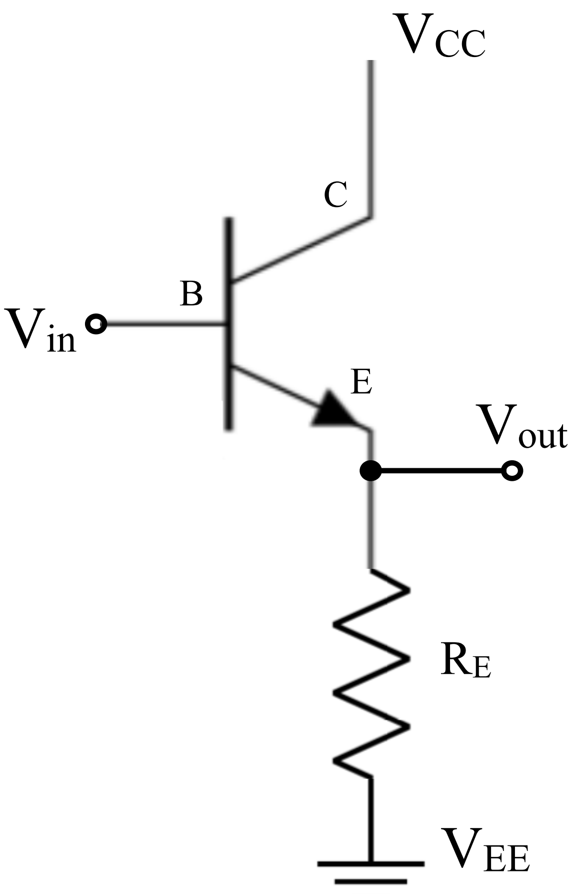
With these simple rules we can analyze most transistor circuits.
We'll add some nuance later today.



Review: Emitter follower

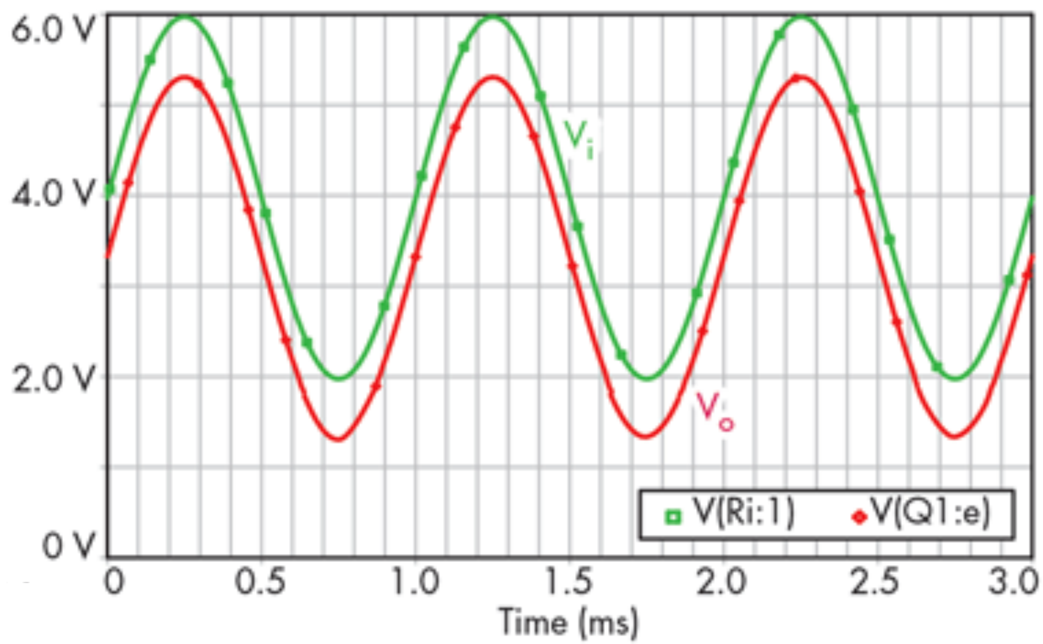
This transistor circuit has the output “follow” the input, with a 0.6 V drop.

$X_{in} = \beta R_E$



$V_{in} = 4 + 2 \sin \omega t$

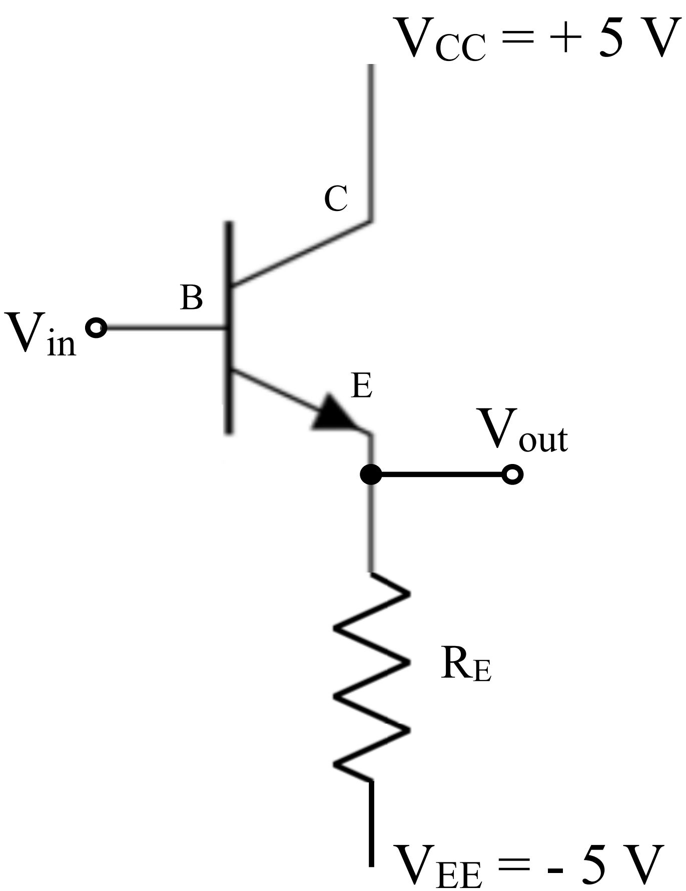
$V_{out} = 3.4 + 2 \sin \omega t$



Review: Emitter follower

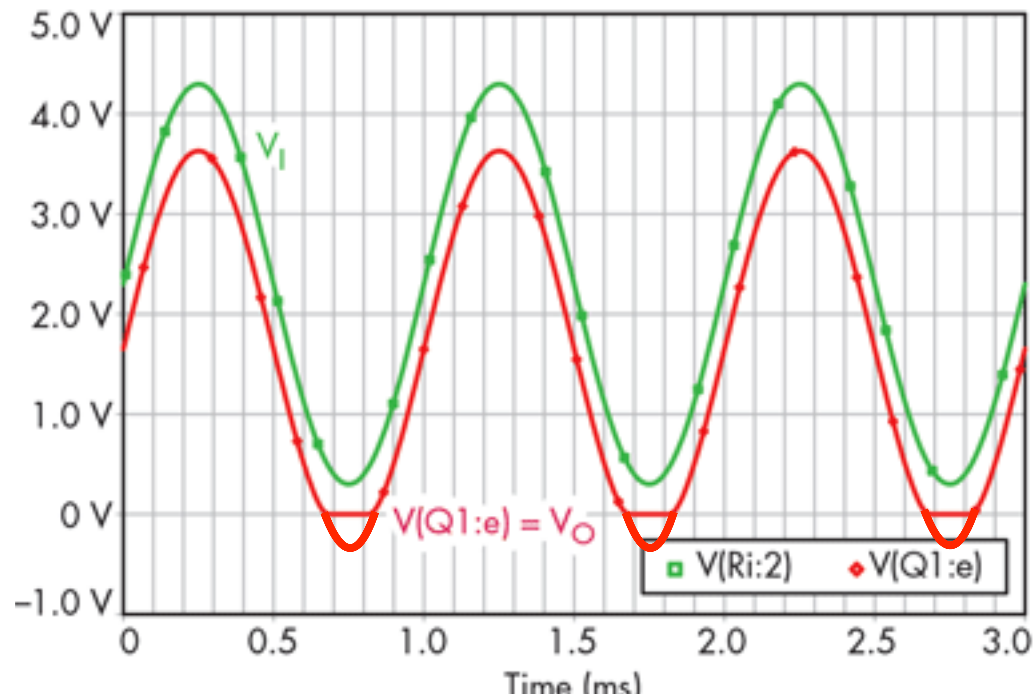
We can remove the clipping at 0 V by setting V_{EE} to a negative supply.

$X_{in} = \beta R_E$



$V_{in} = 2.2 + 2 \sin \omega t$

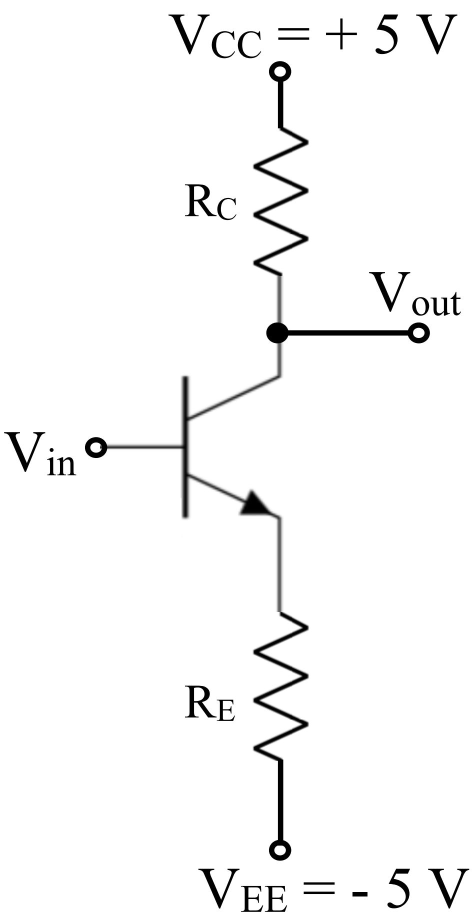
$V_{out} = 1.6 + 2 \sin \omega t$



Output clips at V_{CC} and 0.6 V above V_{EE}.

Review: Common-emitter amplifier

We can use the current amplification of the transistor to get voltage amplification.



$\Delta V_E = \Delta V_B = \Delta V_{in}$

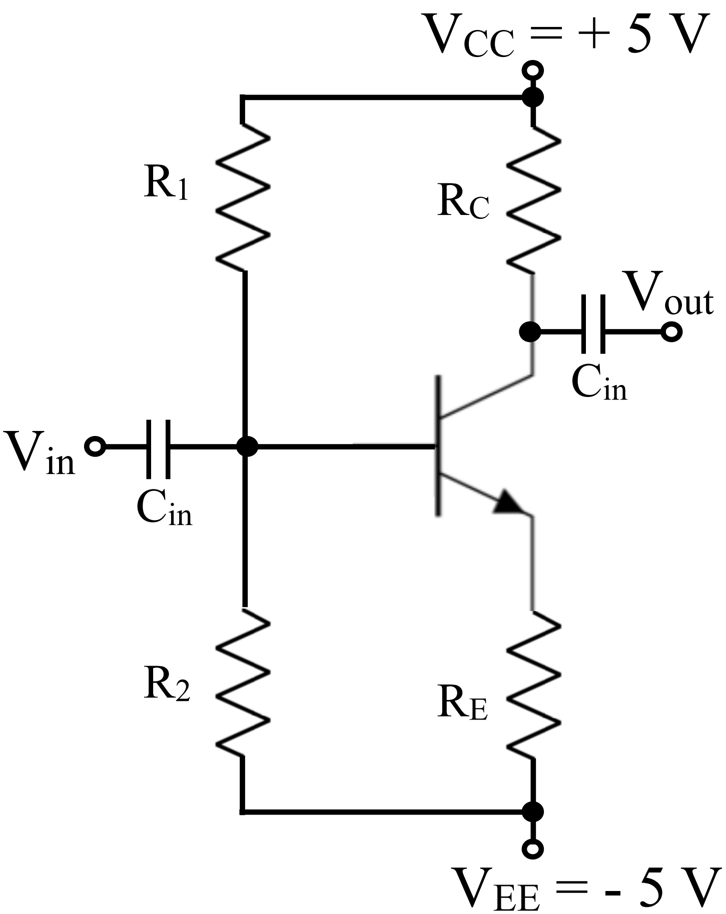
$I_E \cong I_C$

$\Delta V_{out} = -\Delta V_E(R_C/R_E) = -\Delta V_{in}(R_C/R_E)$

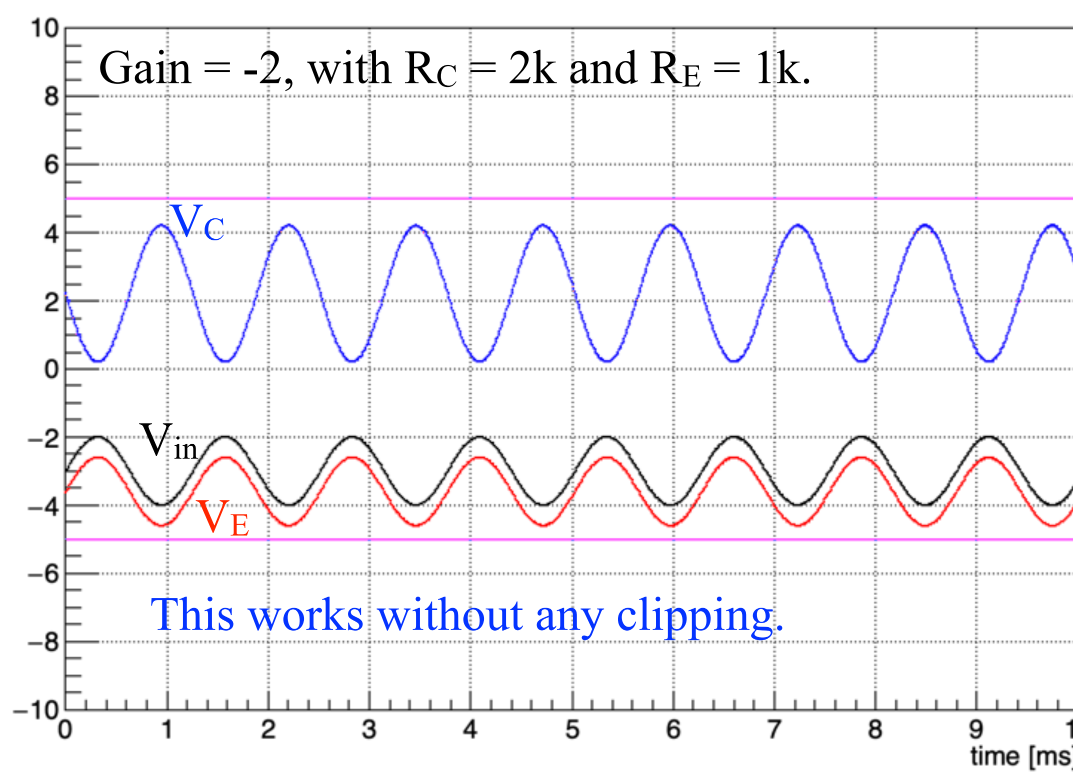
$\text{Gain} = \Delta V_{out} / \Delta V_{in} = -R_C/R_E$

Review: Common-emitter amplifier input biasing

Apply an input bias that puts the emitter close to V_{EE}, within a ΔV that defines the max input swing.



Bias the input to put V_E near V_{EE} and V_{out} half way between V_{CC} and V_{EE}.
Decoupling capacitor on the input and output.



Common-emitter amplifier input biasing

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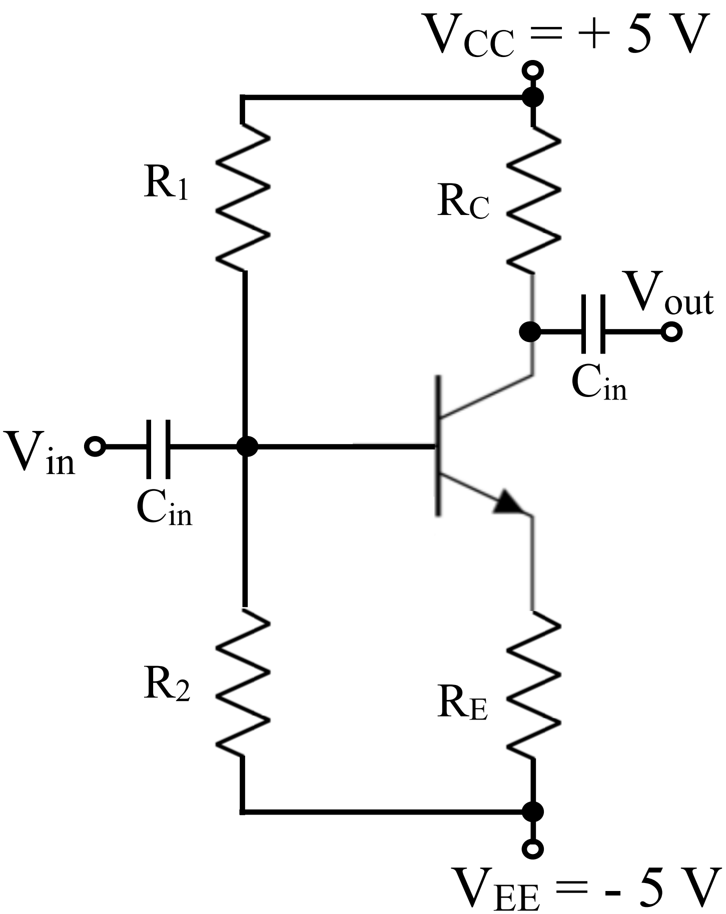


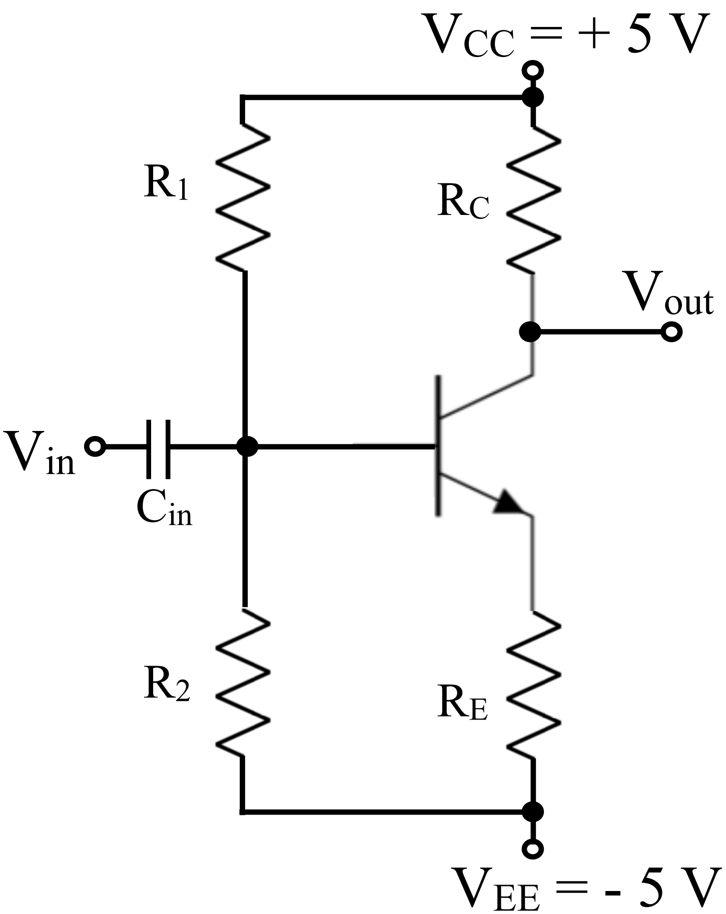
Suppose I want a max input swing of ±0.1 V
Set V_E to vary from -4.8 to -5.0 V, i.e.,
DC set point for V_E is -4.9 V.
DC set point for V_{in} is -4.3 V.
These are called the *quiescent* values, meaning "when quiet, ie without signal".

Choose R_1 and R_2 to be a voltage divider setting V_{in} at -4.3 V.

$V_{in} = V_{EE}+(V_{CC}-V_{EE})*R_2/(R_1+R_2)$
$-4.3 = -5 + 10*1k/(1k+R_2)$
$R_1 = 13k$ and $R_2 = 1k$
Or I could use
$R_1 = 130k$ and $R_2 = 10k$
Which choice is better?

Common-emitter amplifier input biasing

Apply an input bias that puts the emitter close to V_{EE}, within a ΔV that defines the max input swing.



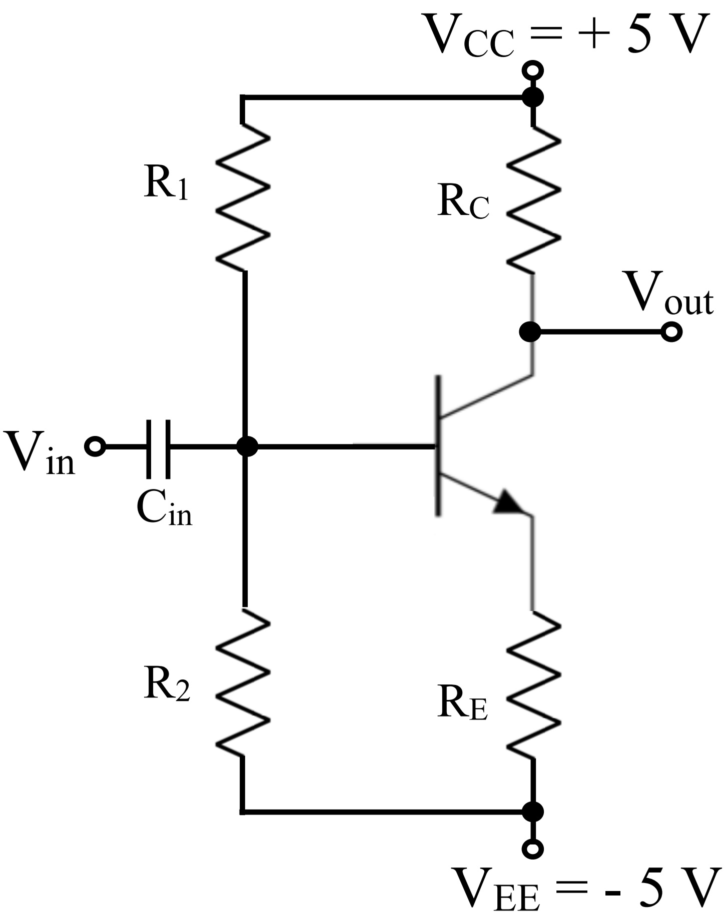
Suppose I want a max input swing of ± 0.1 V
Set quiescent points: V_E=-4.9 V & V_{in}=-4.3 V.
R_1 = 130k and R_2 = 10k

To avoid having this stage yank the output of the previous stage to a different voltage, we *decouple* the input from this "DC bias voltage" with a "decoupling capacitor", C_{in}.

$R_{in}C_{in}$ make a high-pass filter letting the signal through and blocking the DC offsets. What is R_{in}?

Common-emitter amplifier input biasing

Apply an input bias that puts the emitter close to V_{EE}, within a ΔV that defines the max input swing.



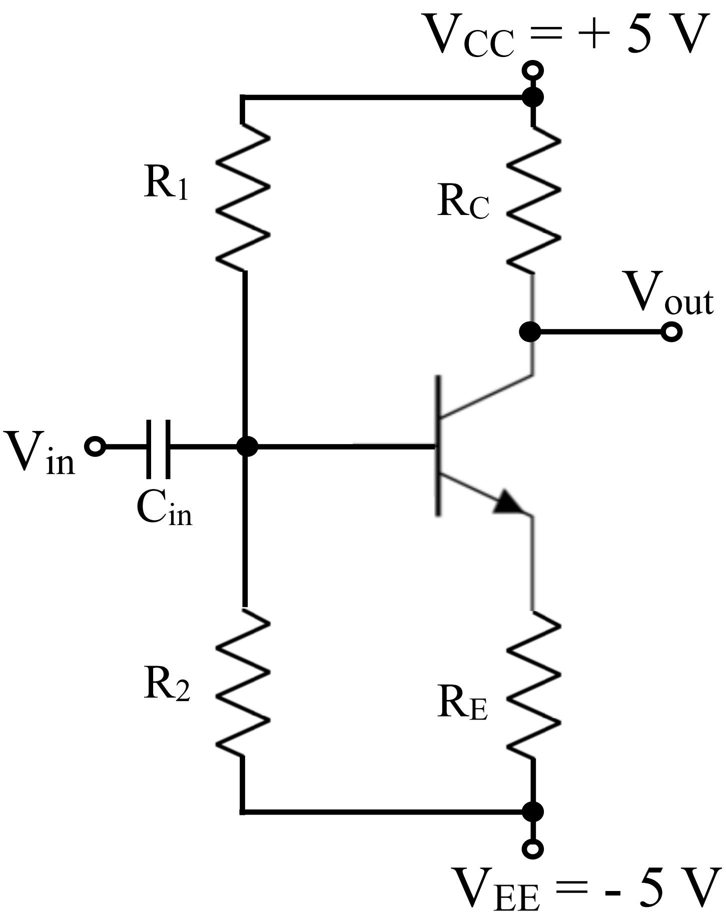
Input impedance is all paths from input to a fixed voltage (V_{CC}, V_{EE}, or Gnd).

$R_{in} = R_1 \parallel R_2 \parallel \beta R_E \cong 130k \parallel 10k \parallel \beta R_E \cong R_2.$

High-pass filter should have f_{3dB}<signal frequency range.
For audio signals, that is 20 Hz, so
$20 = 1/2\pi(10k)C$
$C \cong 1/6*120*10k \cong 1/1k*10k = 0.1\ \mu F$

Common-emitter amplifier input biasing

Now we need to pick R_E and R_C



The ratio of R_E and R_C is set by the desired gain, and avoiding output clipping.

Choose gain = 10, and max $V_{in} = 0.1$ V.
That means V_{out} swings by ±1 V.
Then quiescent point for V_{out} to be at least 1 V away from V_{CC} and V_E.
But,

$$V_{out} = V_{CC} - V_E(R_C/R_E)+V_{EE}(R_C/R_E)$$

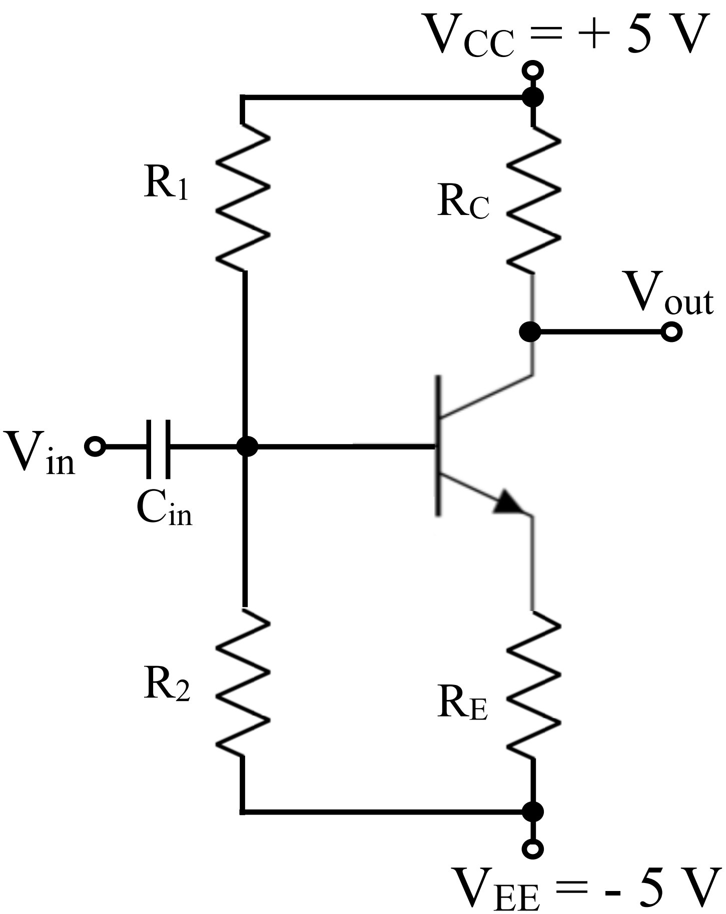
only depends on the gain ratio.

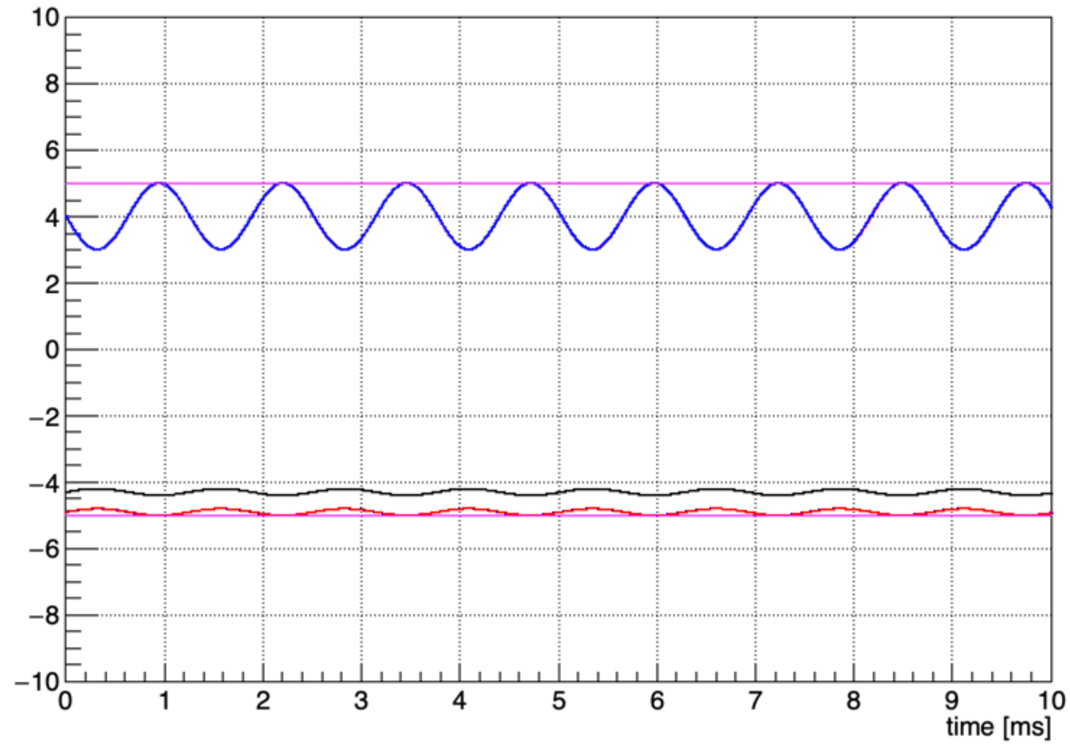
$$V_{out} = 5 - (-4.9*10)-5*10$$
$$= 4$$

That works, but just barely.

Common-emitter amplifier input biasing

Now we need to pick R_E and R_C





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Common-emitter amplifier input biasing

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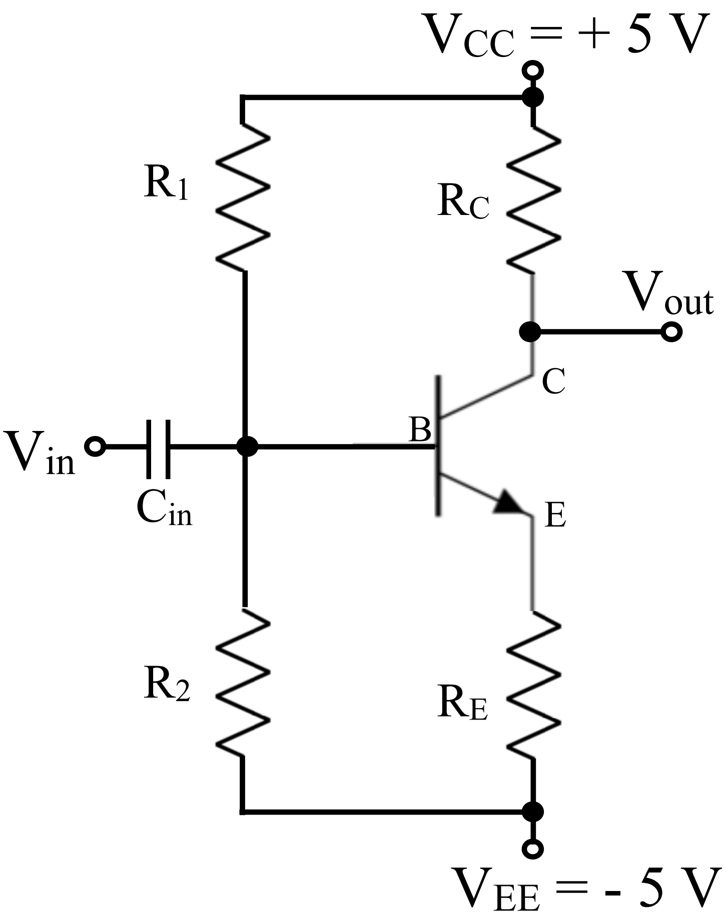
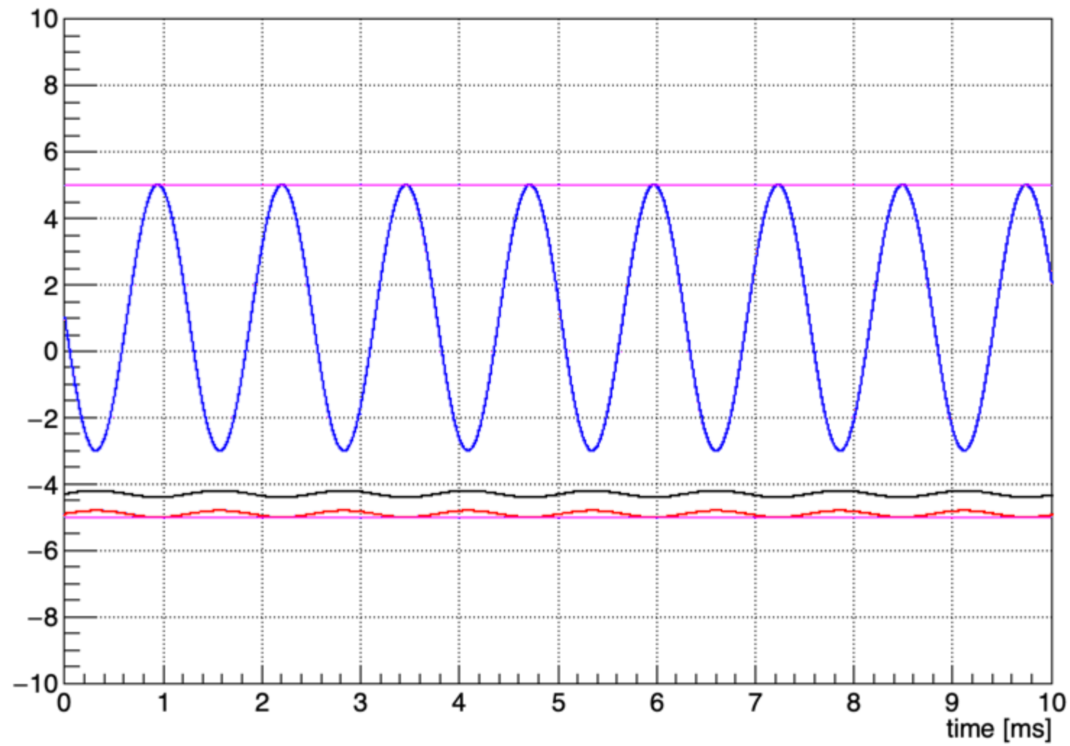
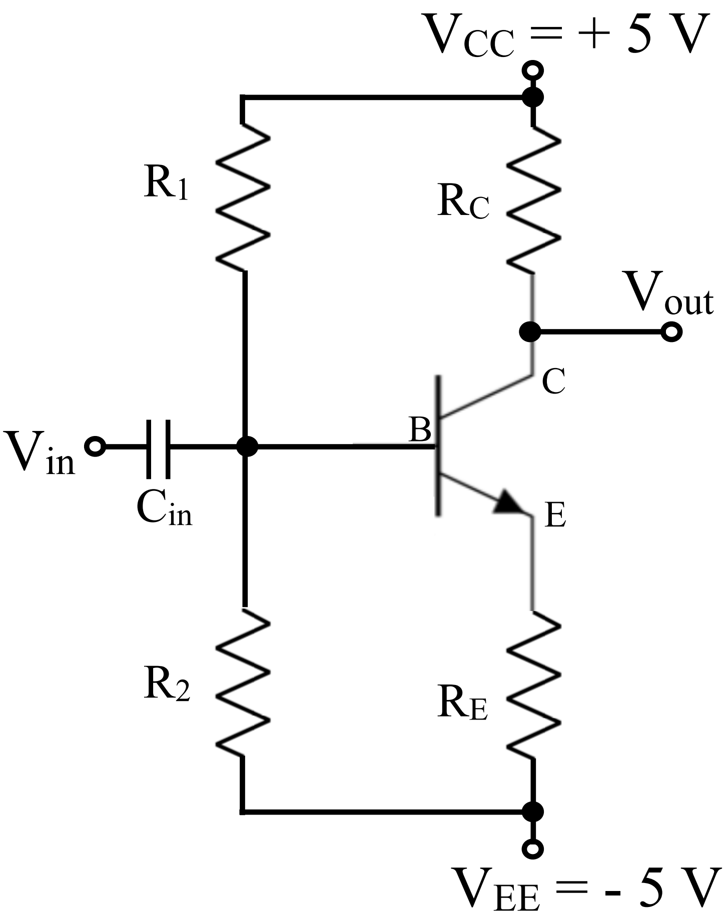




Gain of 40 also works, but only if we change the quiescent V_{out} to 1 V.

Common-emitter amplifier input biasing

The challenge here is that R_E affects both the gain and the quiescent V_{out}.
A small R_E gives big gain but large I_E which affects quiescent V_{out}.



We want a large R_E for setting quiescent voltages and a small R_E for setting gain.

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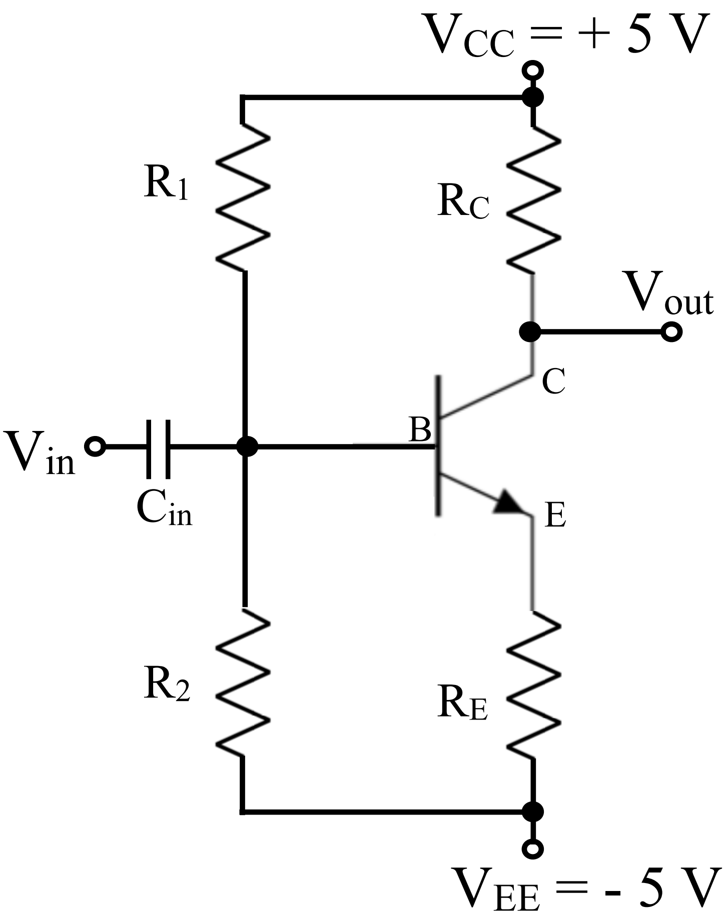

We want a large R_E for setting DC quiescent voltages and a small R_E for setting AC gain.

Common-emitter amplifier input biasing

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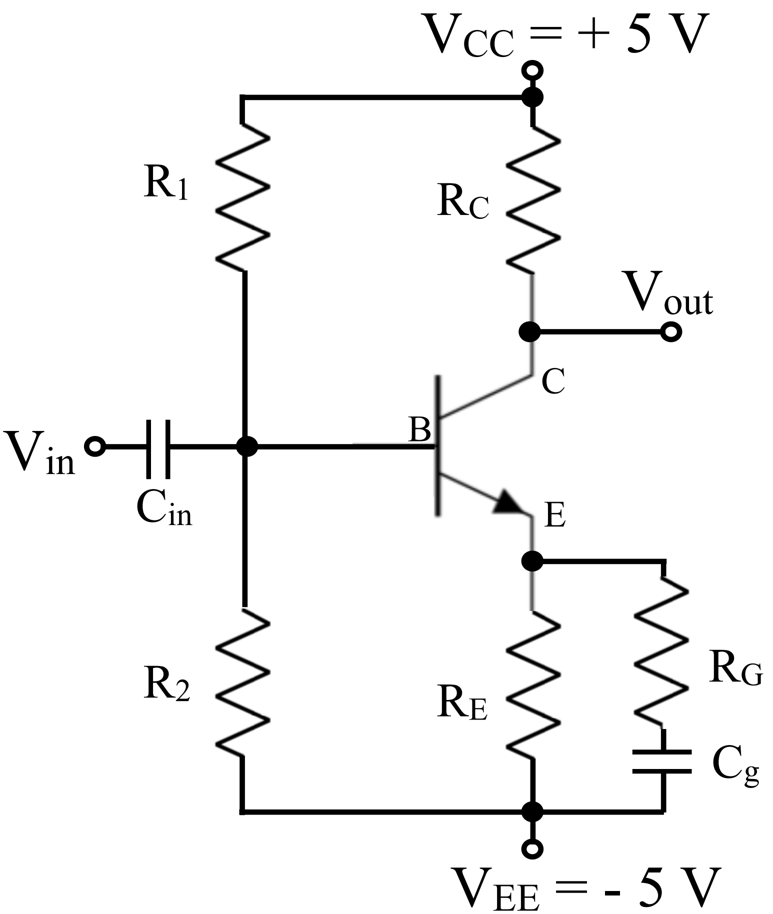
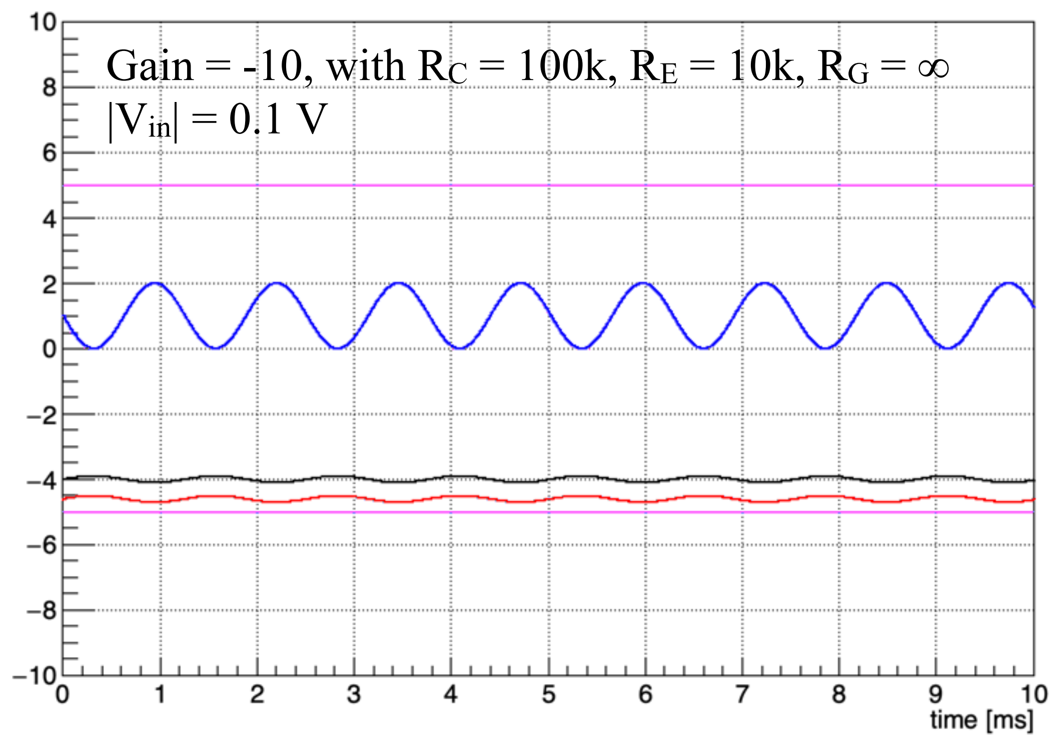


Choose R_1 and R_2 for quiescent V_E = -4 V.
Choose R_E =10k and R_C =100k for quiescent V_{out} = 1 V and base gain of 10.



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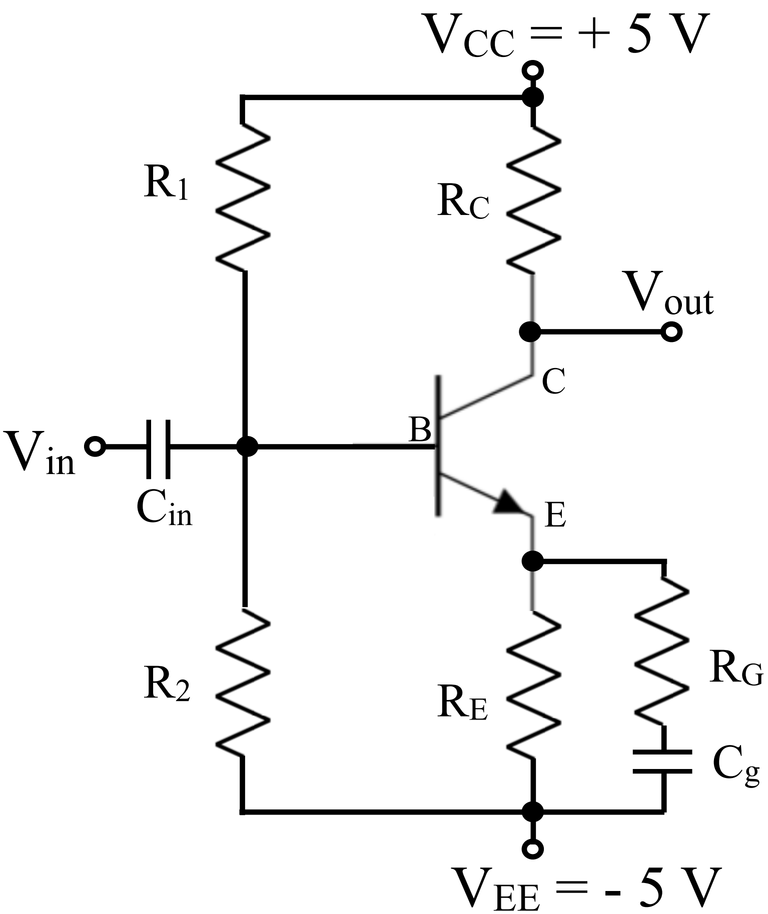

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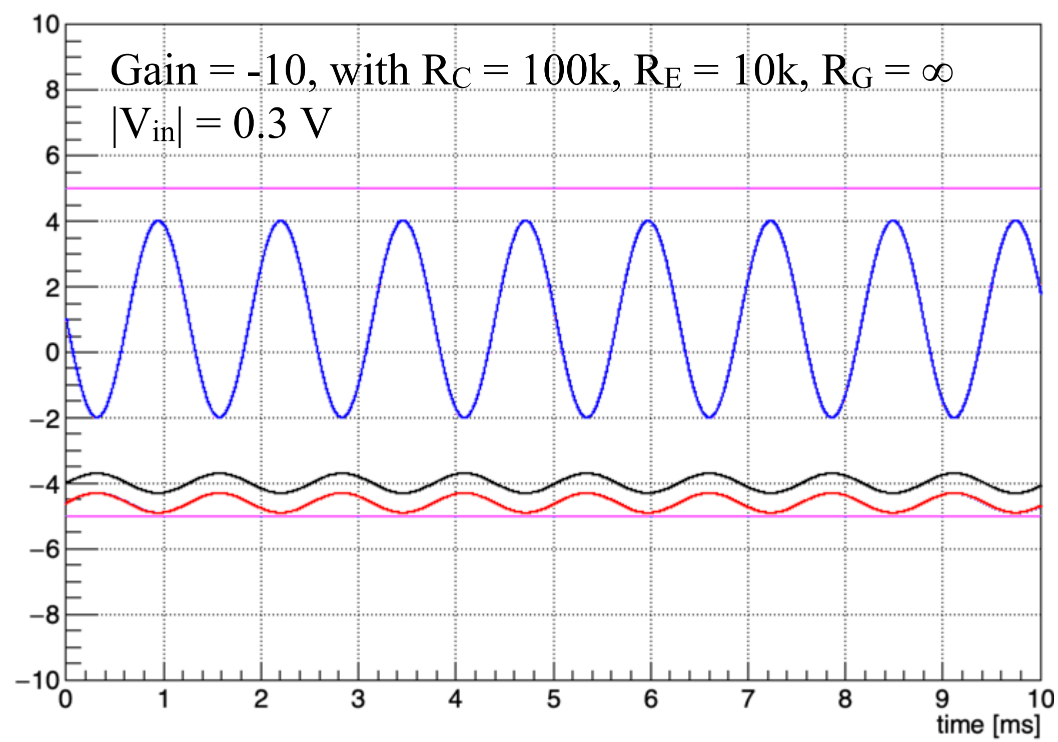

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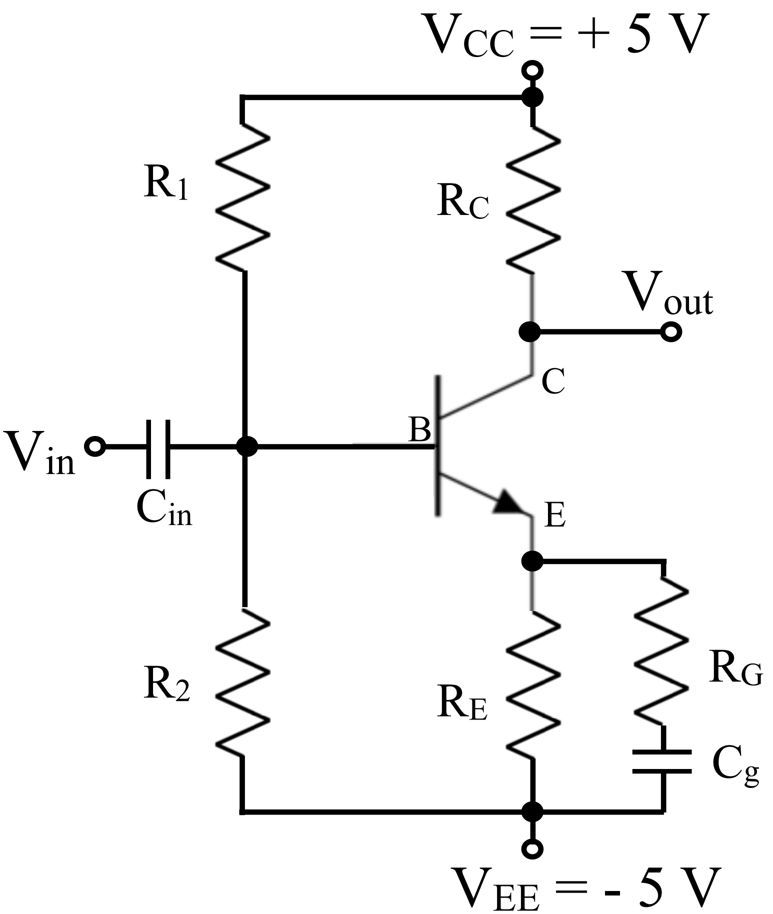

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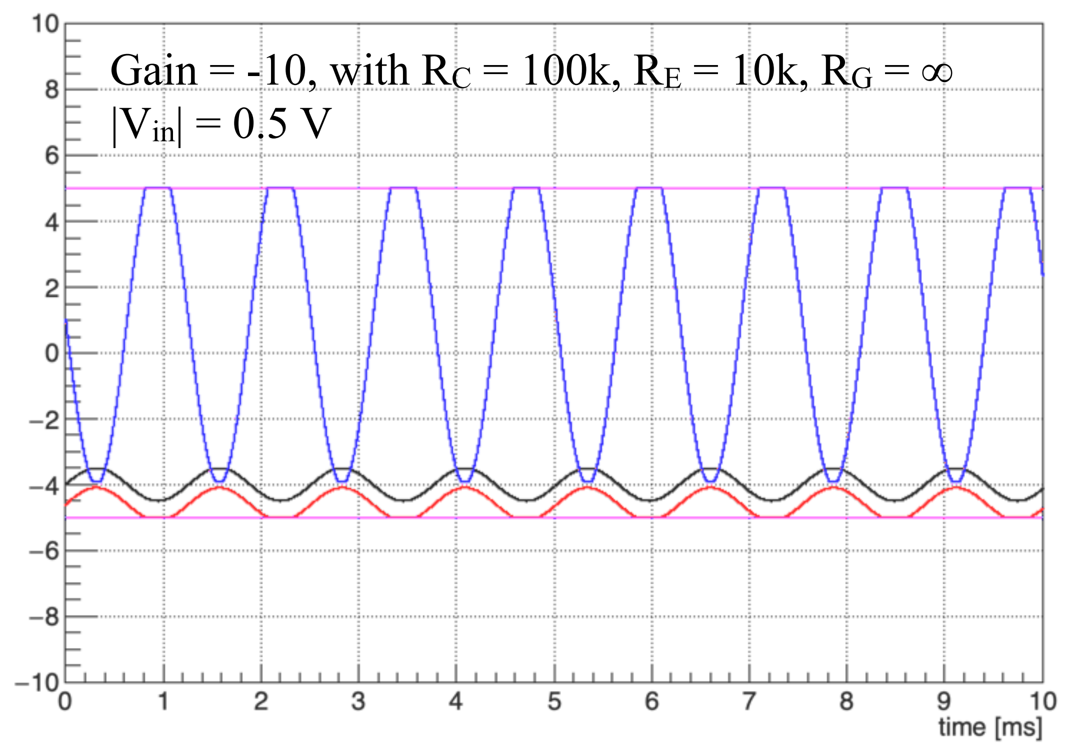

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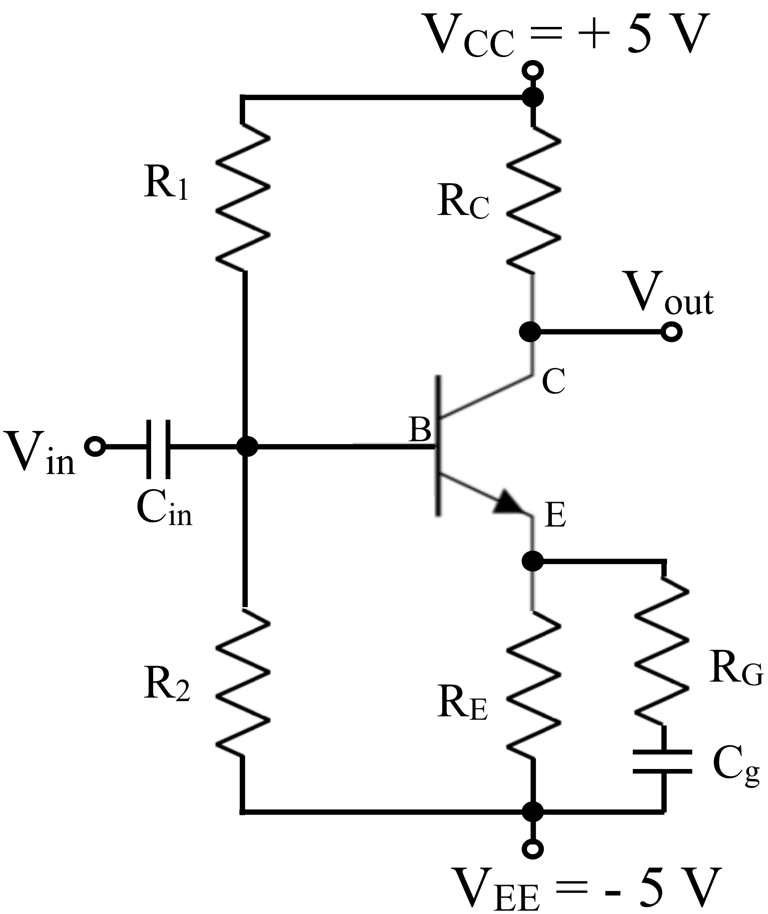

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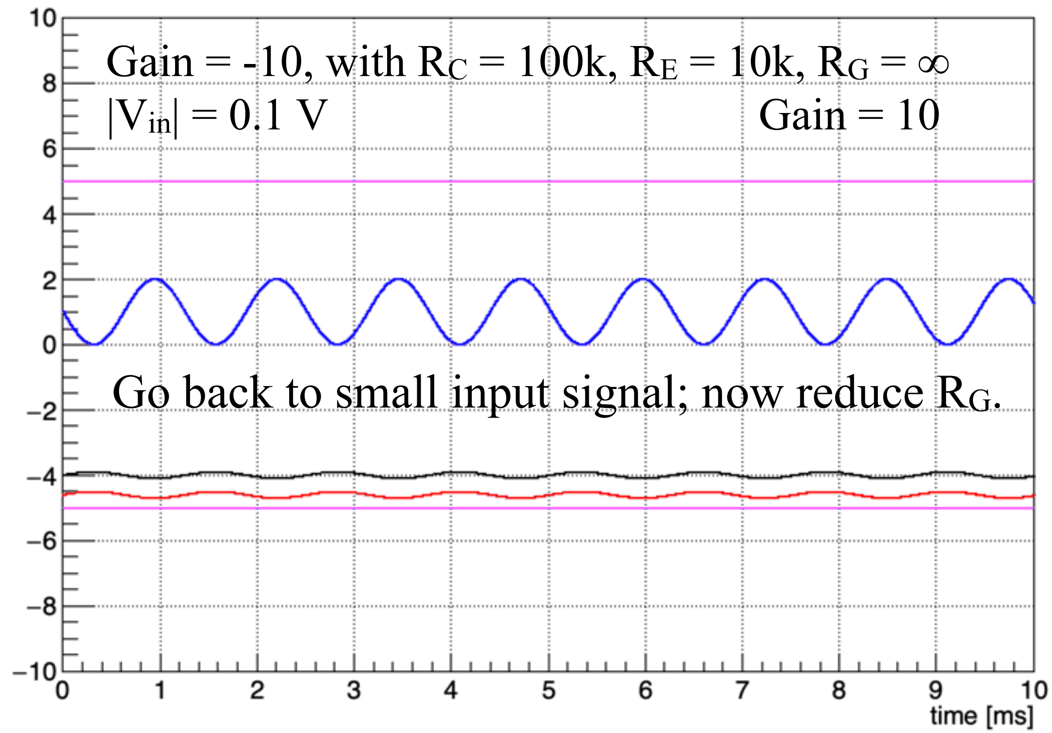

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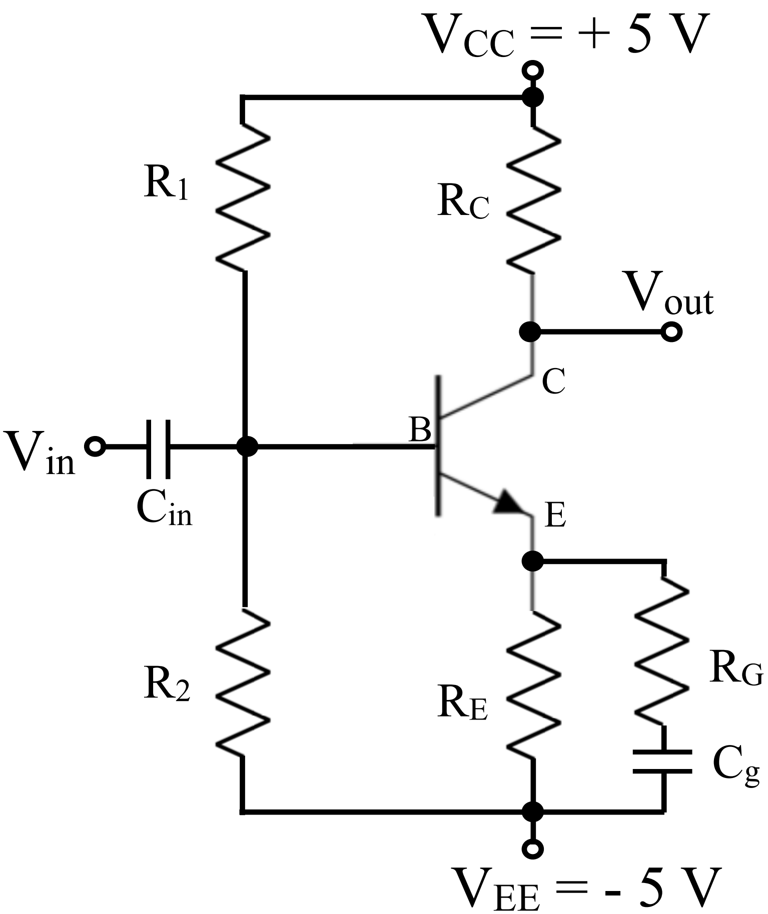

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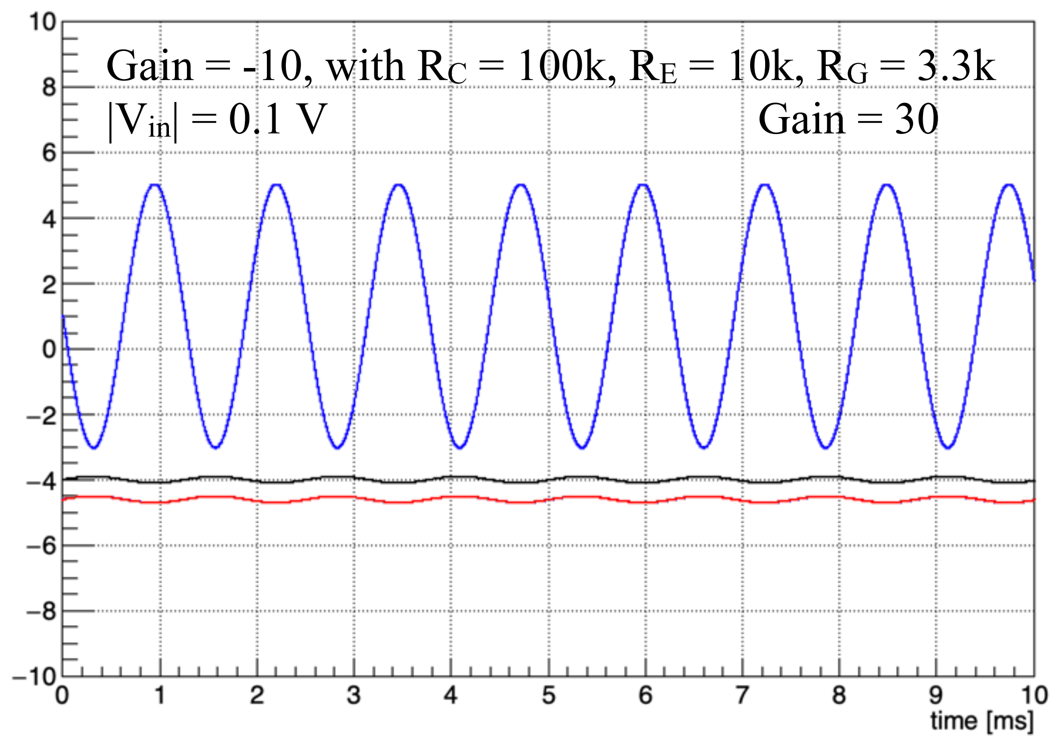

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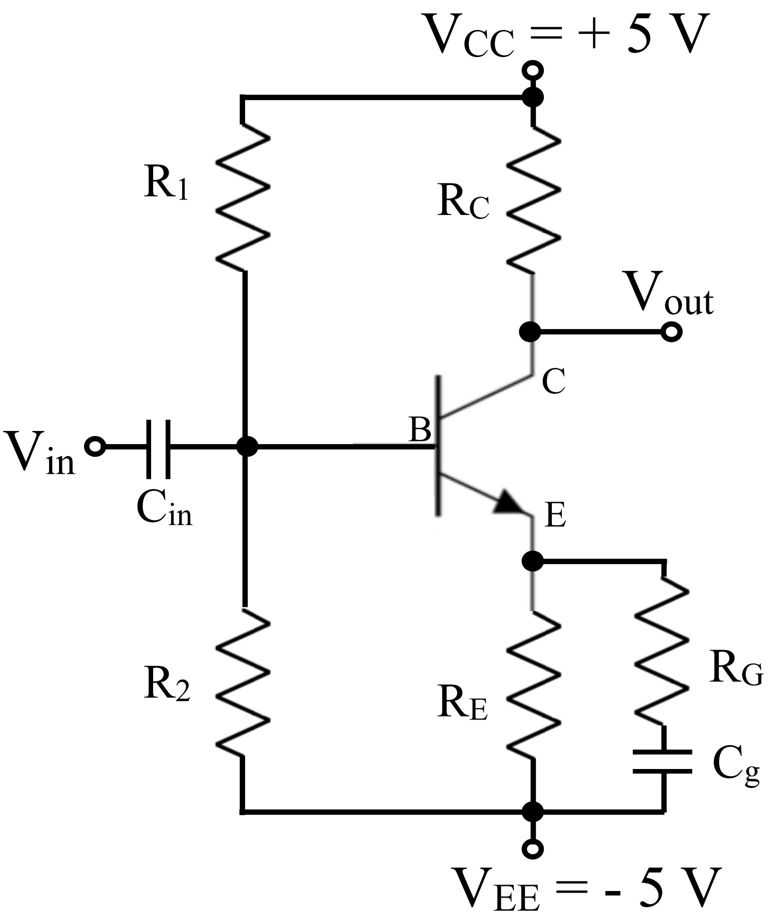

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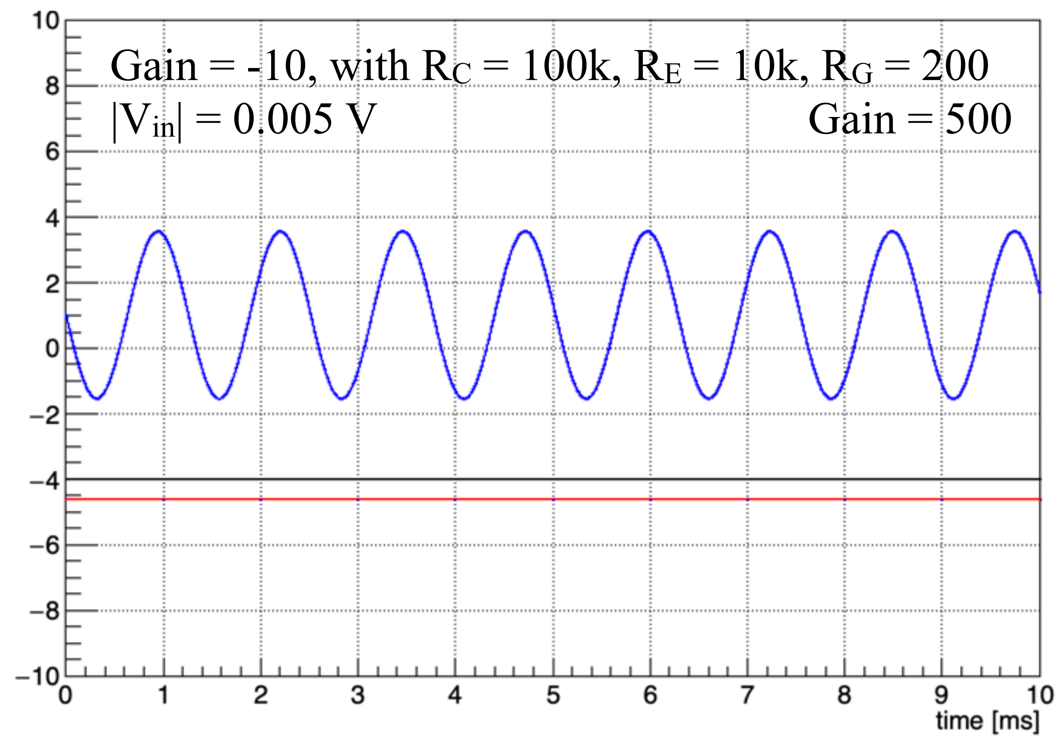
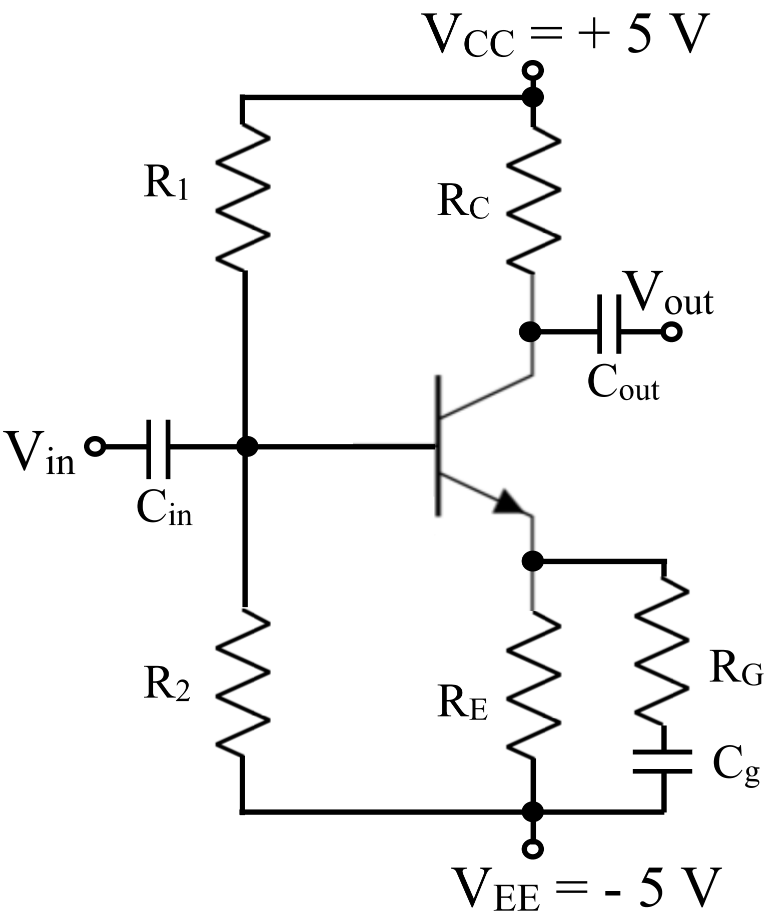
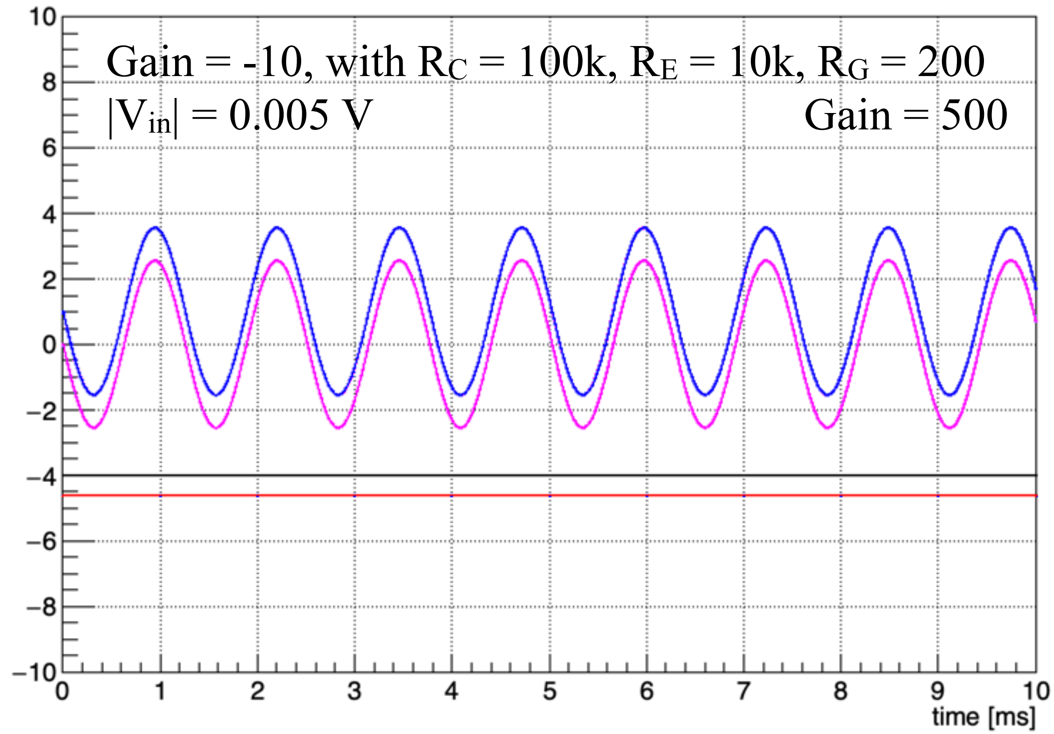


Common-emitter amplifier input biasing

Finally, what can we do about the 1 V quiescent offset on V_{out}?

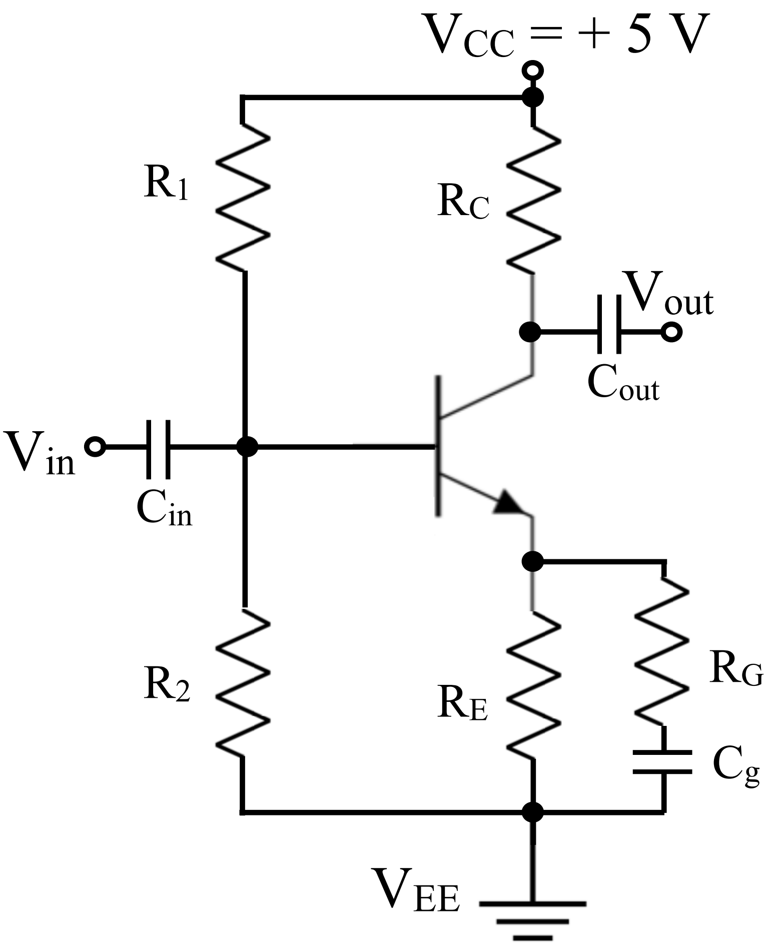


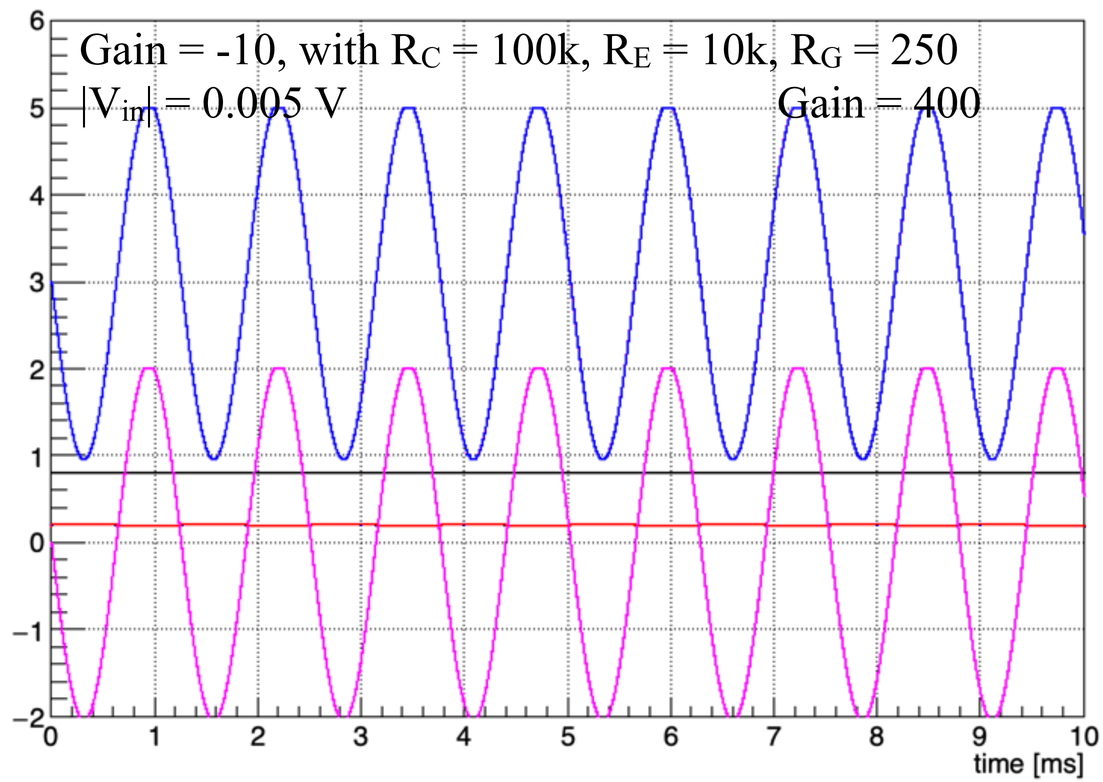
Remove it with a decoupling capacitor.



Common-emitter amplifier input biasing

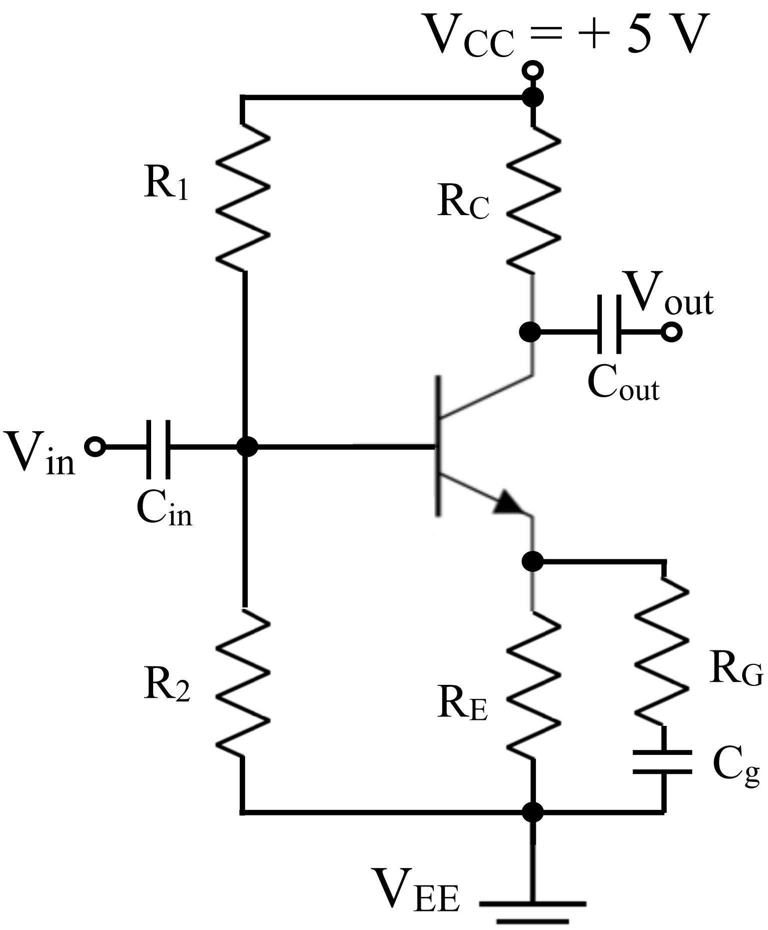
This also works if V_{EE} is ground. We just choose quiescent points.
In fact with V_{EE}=Gnd, we *must* have input biasing.





Common-emitter amplifier input biasing

Some checks of understanding.



Without DC biasing, what would limit the signal?

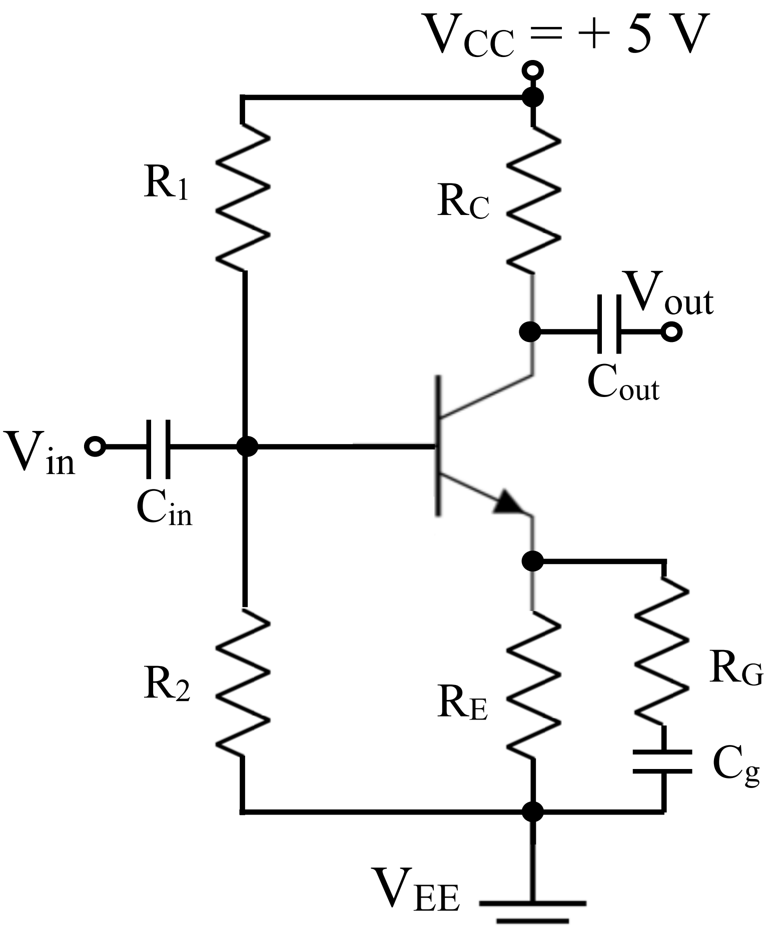
What is the output impedance of this circuit?

With V_{EE} = Gnd, about where should you put the quiescent V_{out}? Where is the quiescent V_{in}?
In general, how do you maximize the *dynamic range*?

What would happen if you set $R_G = 0$?

Common-emitter amplifier operation

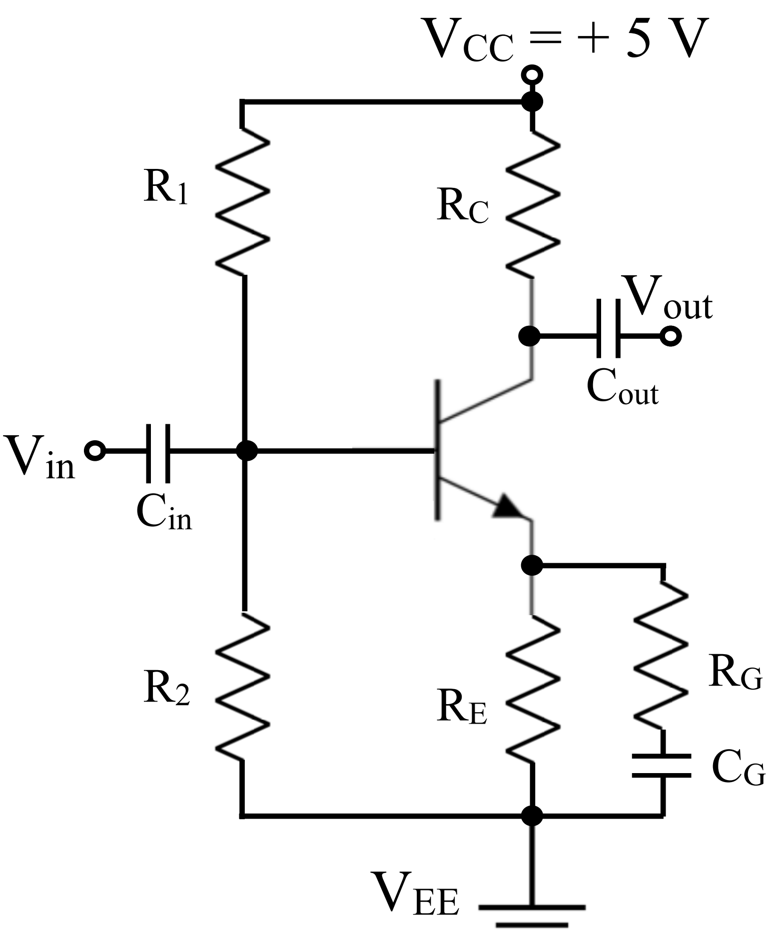
The transistor is changing the voltage dropped across it to satisfy the rules of operation.



Increase in V_{in} causes increase in V_E
That causes an increase in I_E
That causes a decrease in V_C
The voltage across the transistor, V_{CE}, goes down to compensate.

Review: Common-emitter amp AC gain control

We separated the determination of quiescent points from determination of gain with a gain resistor that only matters for signal because of C_G.

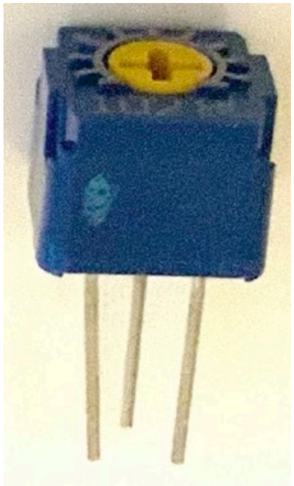


Gain = $\Delta V_{out} / \Delta V_{in} = - R_C/(R_E \parallel (R_G+C_G))$

Gain $\cong - R_C/R_G$

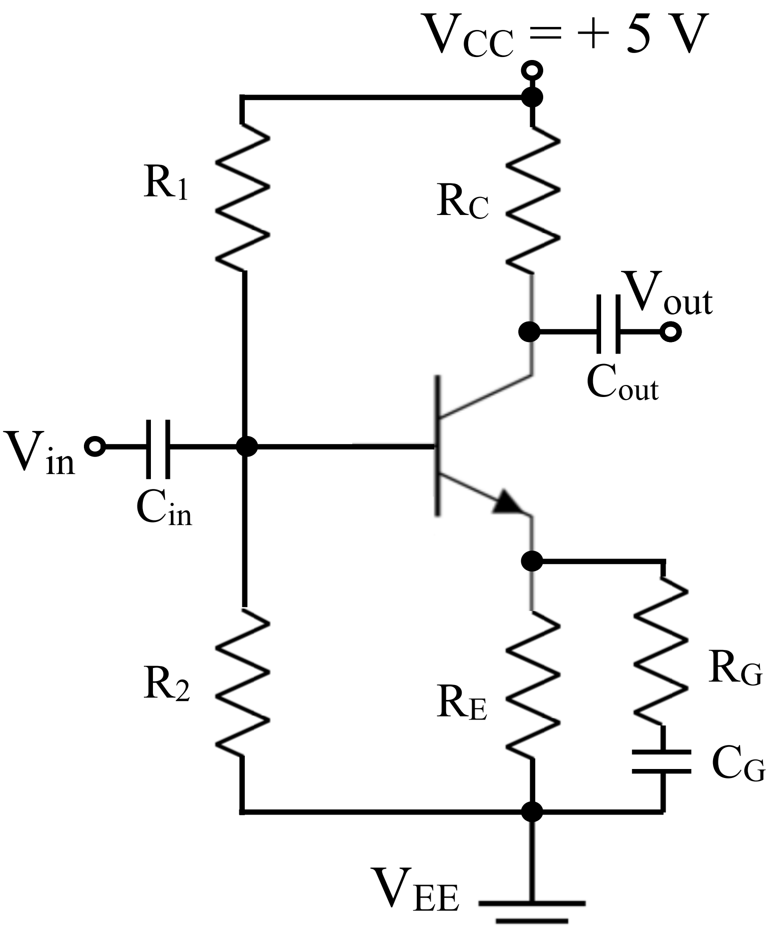
R_E can be chosen to set V_{out} quiescent point.
It can be large, and so can R_C.

R_G then is chosen to set the gain,
it can vary without altering quiescent points,
and it needn't be too small.



Common-emitter amp input impedance

What is the input impedance of the common-emitter amp?



Follow all paths to fixed voltages.

$$
\begin{aligned}
X_{in} &= C_{in} + \{R_1 \parallel R_2 \parallel \beta(R_E \parallel [R_G + C_G])\} \\
&\cong 0 + \{R_1 \parallel R_2 \parallel \beta(R_E \parallel [R_G + 0])\} \\
&\cong 0 + \{R_1 \parallel R_2 \parallel \beta R_G\} \\
&\cong R_2 \parallel \beta R_G
\end{aligned}
$$

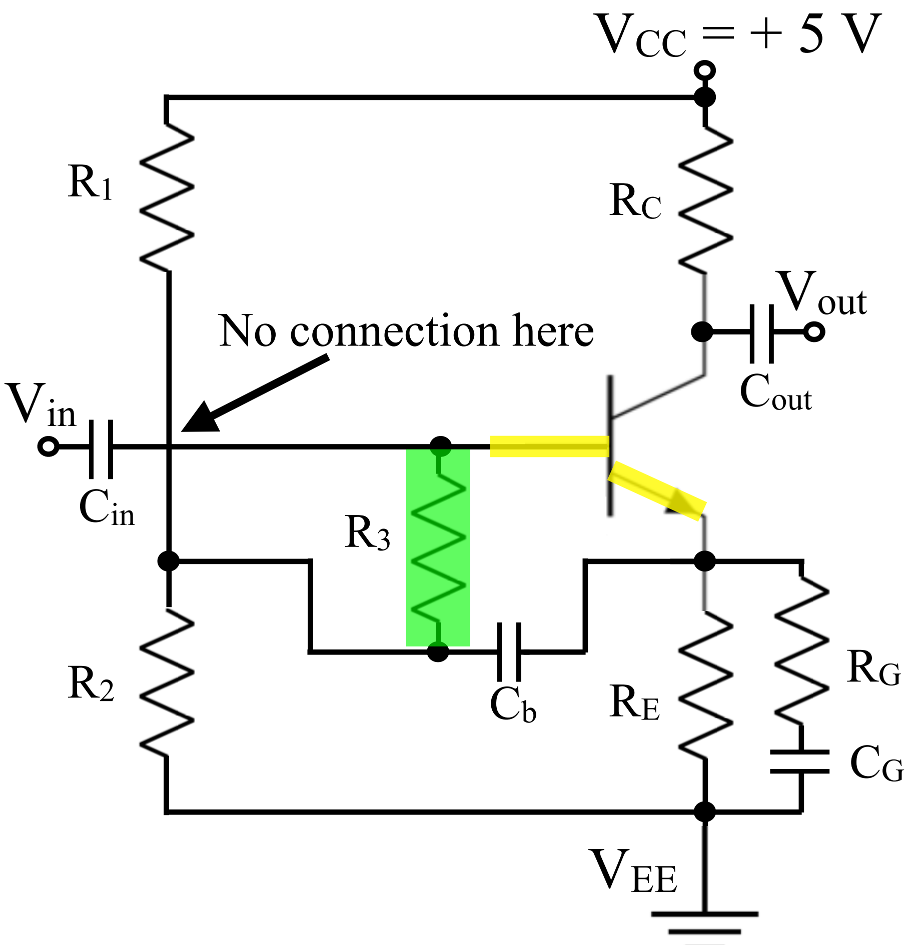
($R_2 << R_1$ because we want to bias V_E close to V_{EE}.)

Can make βR_G reasonably large.
What about R_2?

We can make the bias network have very large impedance with a trick called bootstrapping. It uses the same signal specific impedance trick as C_G does.

Bootstrapping

Add capacitive feedback from V_E to V_B.



Now the input impedance is

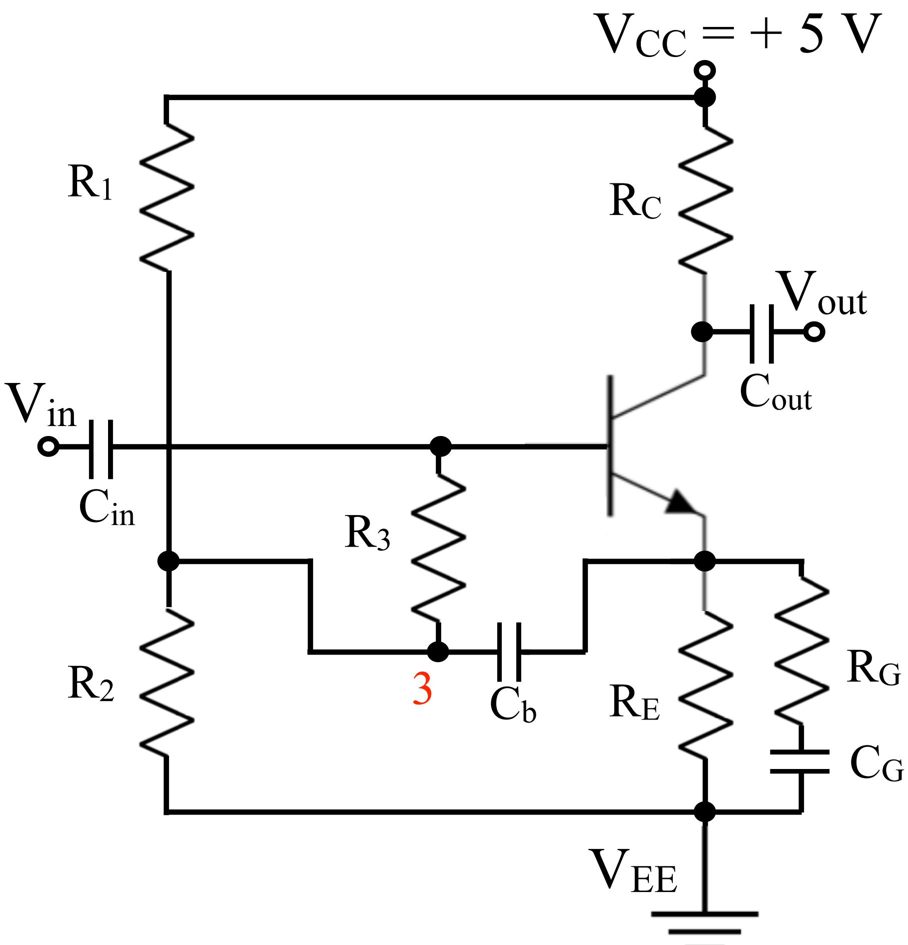
$X_{in} = C_{in} + \beta(R_E \parallel [R_G + C_G] \parallel \{C_b + R_1 \parallel R_2\}) \parallel$
$[R_3 + (R_1 \parallel R_2 \parallel \{C_B + R_E \parallel [R_G + C_G]\})]$
$\cong \beta(R_E \parallel R_G \parallel R_1 \parallel R_2) \parallel$
$[R_3 + (R_1 \parallel R_2 \parallel R_E \parallel R_G)]$
$\cong R_3 + (R_1 \parallel R_2 \parallel R_E \parallel R_G)$

The bootstrapping trick is to make R_3 go to infinity <u>for signal</u>.

This works because $X_{C_b}=0$ <u>for signal</u> and $\Delta V_B = \Delta V_E$ <u>for signal</u>.

Bootstrapping

Add capacitive feedback from V_E to V_B.



The input impedance is

$$X_{in} \cong R_3 + (R_1 \parallel R_2 \parallel R_E \parallel R_G)$$

Imagine a small signal $v_{in} = \Delta V_{in}$.
It passes through C_{in} and also through the transistor because $\Delta V_B = \Delta V_E$.

$v_{in} = v_B = v_E = v_3$

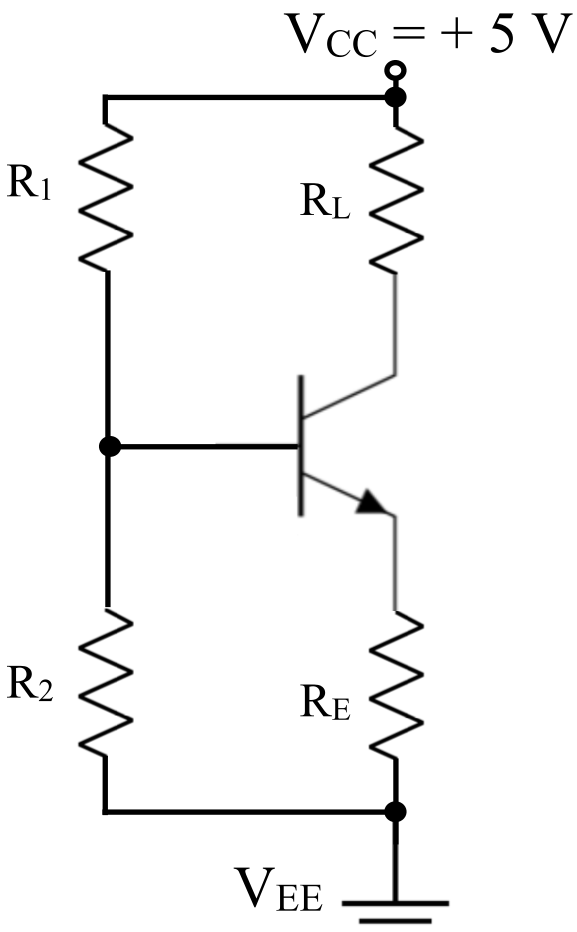
This means a small signal moves the top and bottom of R_3 the same, $v_B = v_3$.

The ΔV across R_3 is 0. So no signal current has to flow through R_3 into R_1 and R_2. The impedance of $R_3 \rightarrow \infty$ making R_1 and R_2 unimportant for X_{in}.

Separate DC and AC response.

Constant current source

We can use a transistor to pull a *constant* specified current through a load.



To get a constant 1mA flow through R_L, even as R_L changes, we can set R_E to 1k and V_E to 1 V.

That sets the value of I_E, which is equal to I_C, regardless of R_L.

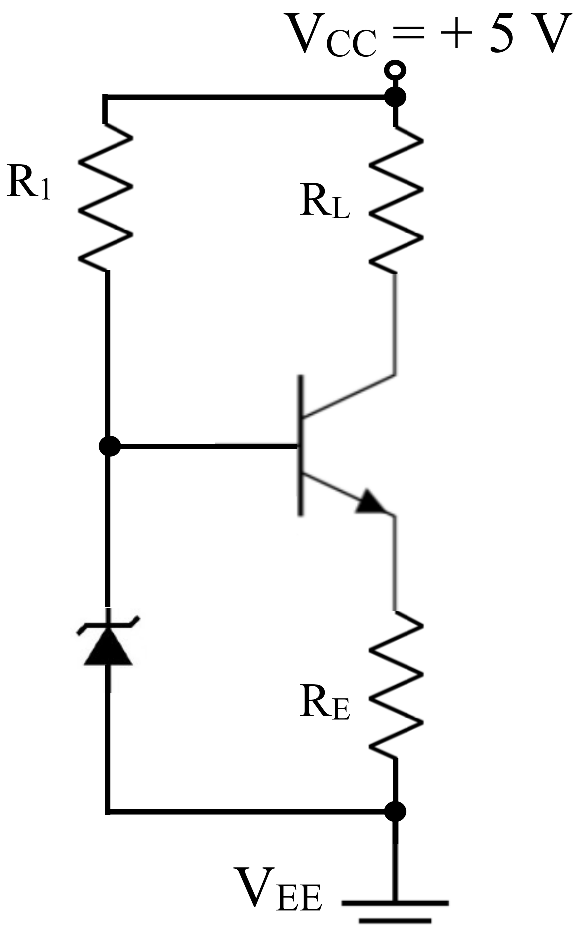
Choose R_1 and R_2 to make $V_B = 1.6$ V.
Then $V_E = 1.0$ V.
$I_E = 1$ mA.
$I_C = 1$ mA, regardless of R_L.

This works until $V_C < V_E + 0.2$

Note that there is no input signal here.

Constant current source

We can use this to pull a specified current through a load.

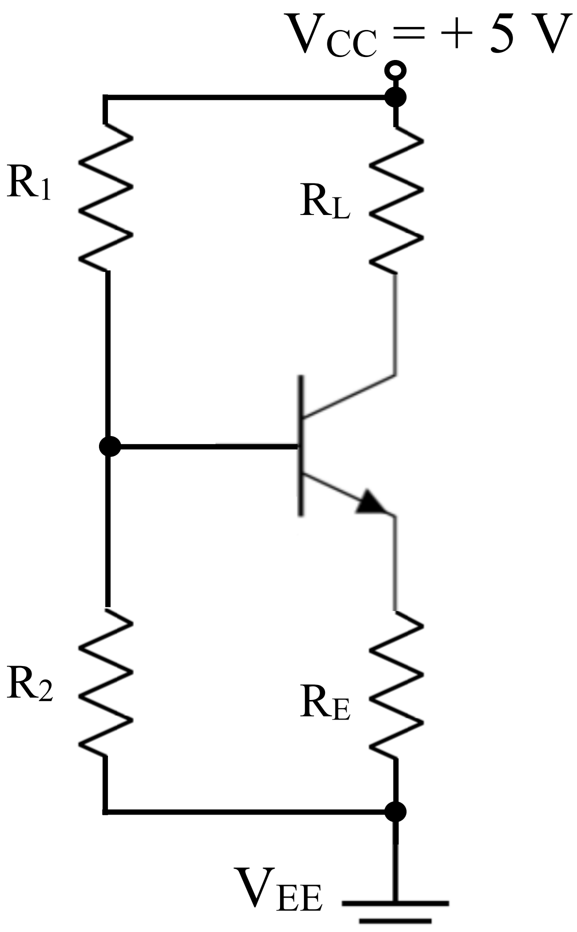


To get a constant 1mA flow through R_L, even as R_L changes, we can set R_E to 1k and V_E to 1 V.
That sets I_E which is equal to I_C, regardless of R_L.

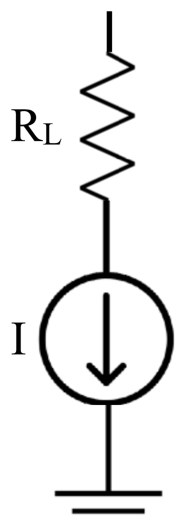
Choose the zener diode to make $V_B = 1.6$ V.
The zener reduces sensitivity to V_{CC} variations.

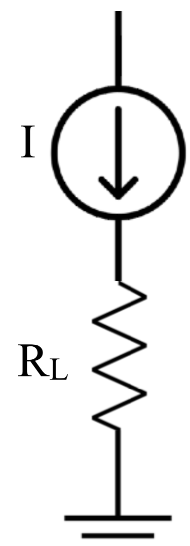
Constant current source

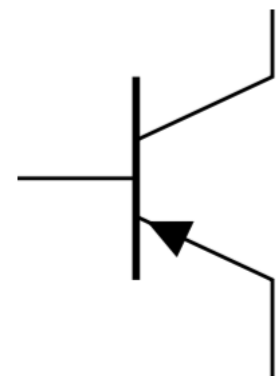
We can use a transistor to *pull* a *constant* specified current through a load. This is actually called a *current sink* since it pulls current from R_L.



If we wanted to push current through R_L with the bottom of R_L connected to ground, then we need a different polarity transistor. PNP vs NPN.



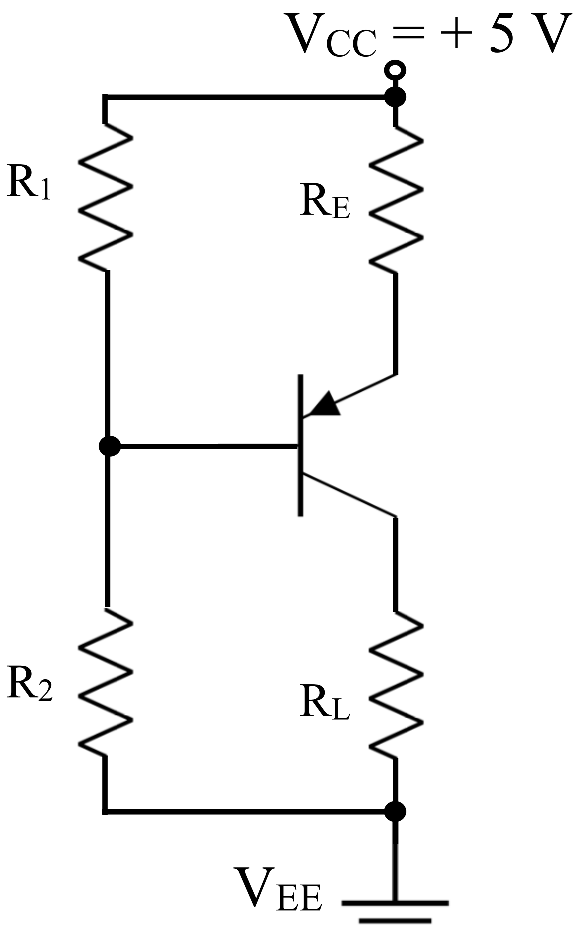




Turns on when $V_E = V_B + 0.6$

Constant current source

We can use a PNP transistor to *push* a *constant* specified current into a load.



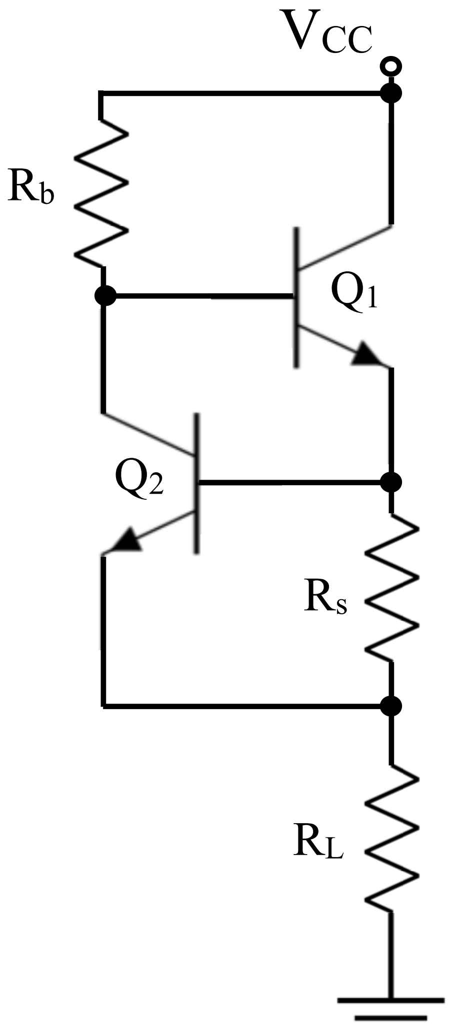
Now we can switch the location of R_L and R_E. The base's bias voltage sets R_E which sets I_E and hence I_C.

For 1 mA we could set $R_E = 1$k and $V_E = 4$ V. That requires $V_B = 3.4$ V which we get from R_1 & R_2 choice.

$3.4 = 5\, R_2/(R_1+R_2)$

Current limiter

Separate from having a constant current, we often want to limit $I<I_{max}$.



If Q_2's $V_{BE}<0.6$ V it turns off, so no current flows through R_b and Q_1 has a high V_b and Q_1 is on.

If enough current flows to cause the voltage drop across R_s to go above 0.6 V, Q_2 turns on and current flows through R_b. That reduces the base voltage of Q_1, lowering the current through Q_1 and hence the current through R_s to turn off Q_2. This rapid on/off leads to an equilibrium at the max current of $0.6/R_s$.

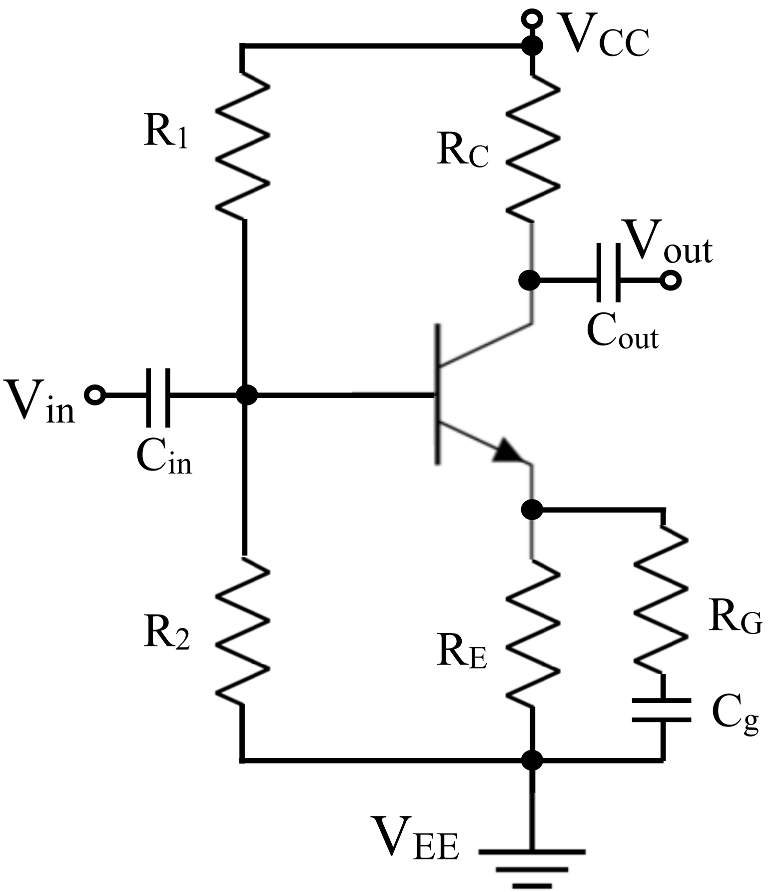
I.e., attempts to increase the load current beyond $I_L = 0.6/R_s$ (either by higher V_{CC} or lower R_L) will lead to a max current of $0.6/R_s$.

E.g., $R_S = 0.6\Omega$ limits load current to 1 A.

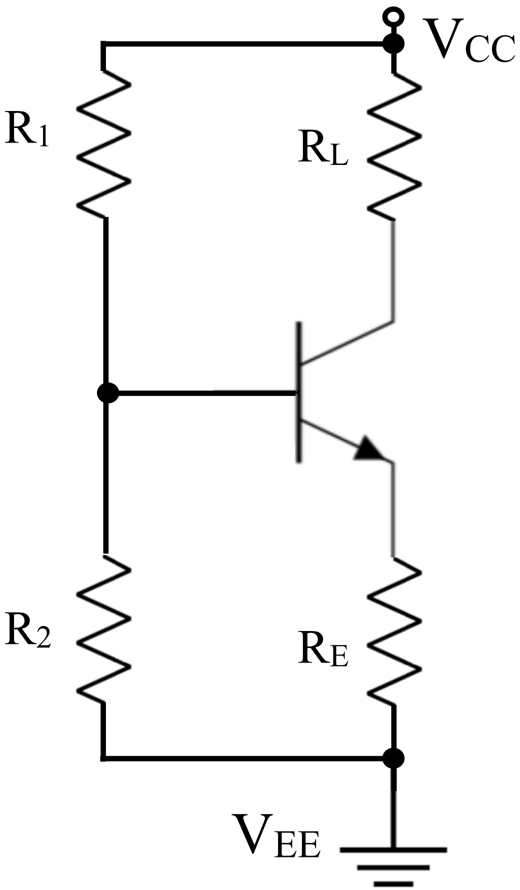
Ebers-Moll model

The simple transistor rules we have been using aren't the full picture.
Two examples of features it misses.

Gain limit with R_G=0.

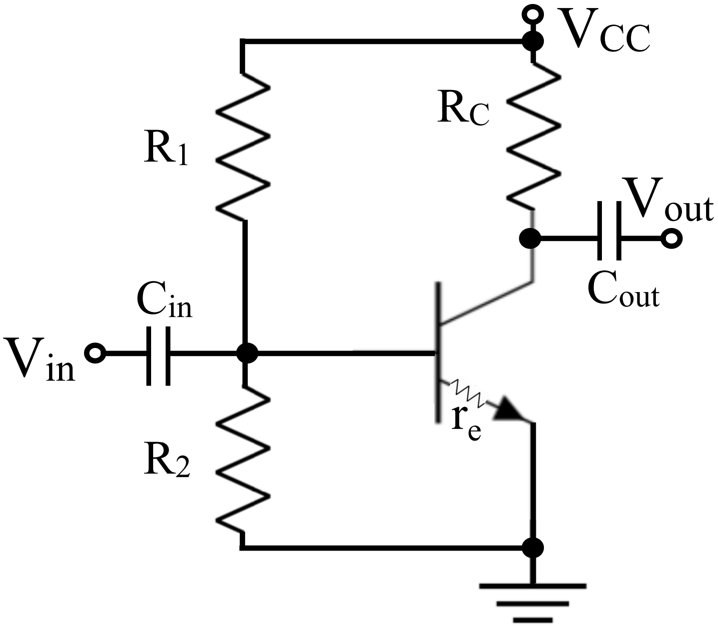


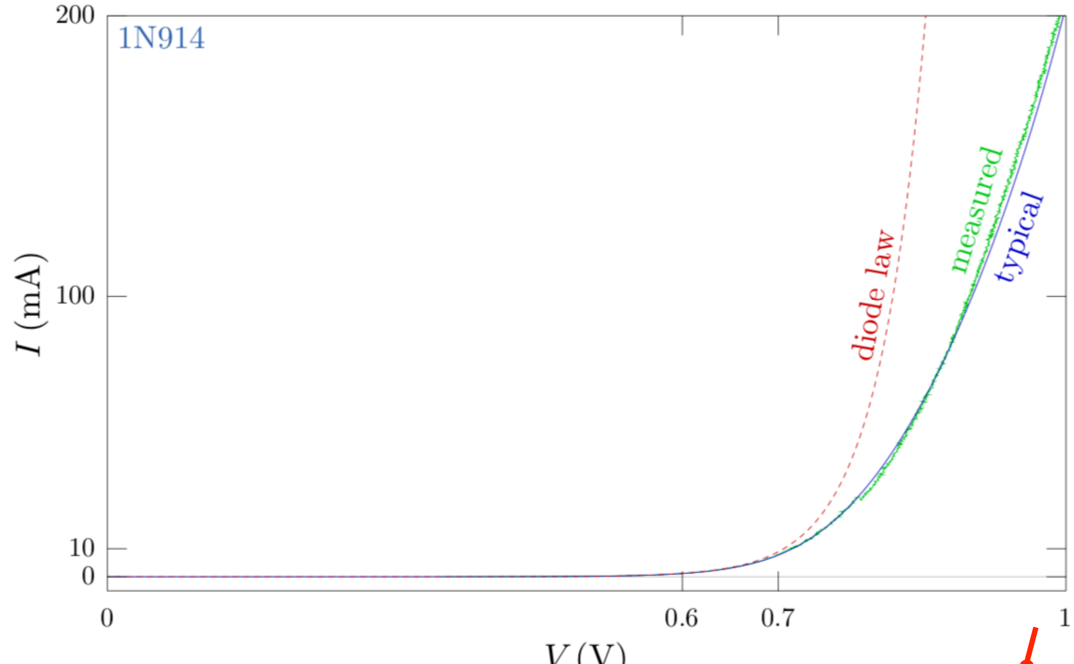
I_L is temperature dependent.



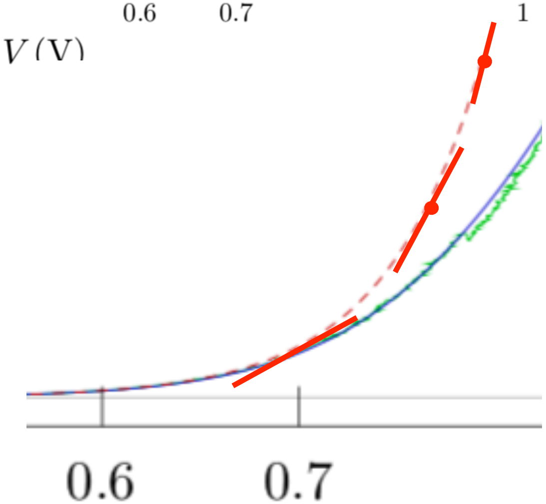
Ebers-Moll model

Gain limit comes from intrinsic resistance in the transistor.





There is a small current-dependent resistance present even when R_E=0. Call it r_e. $r_e \cong 25\Omega$ / I[mA]



Ebers-Moll model

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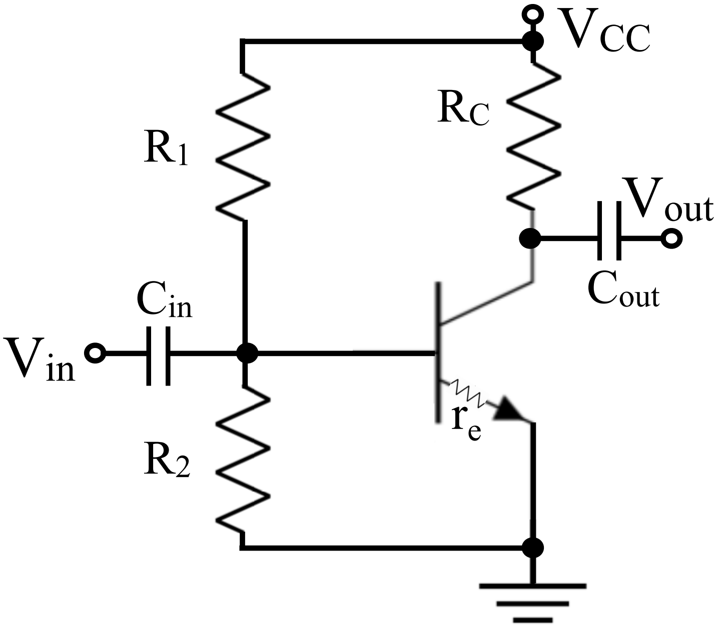
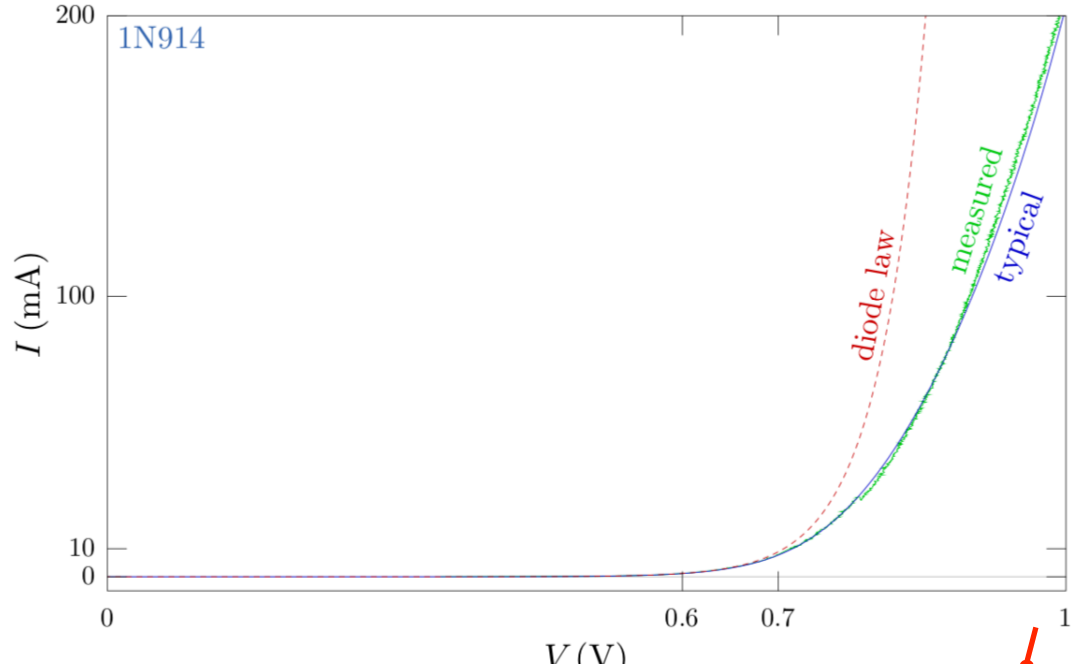
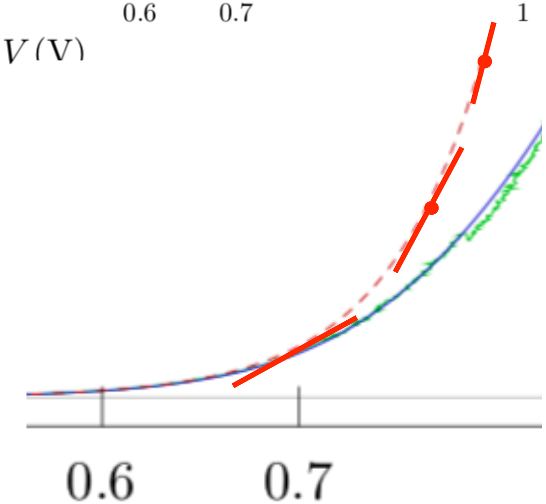




There is a small current-dependent resistance present even when R_E=0. Call it r_e. $r_e \cong 25\Omega$ / I[mA]

And it is temperature dependent.

$$I = I_S\left(e^{V/nV_T} - 1\right) \quad \text{where} \quad V_T = \frac{k_B T}{e}$$



$V_T \cong 25$ mV at room temperature.

I_s is also temperature dependent.

Ebers-Moll model & transconductance

This illustrates something called transconductance,
which is like gain but unit-full.
Gain = $\Delta V_{out} / \Delta V_{in}$ is unitless.
But really, changing V_{in} changes V_{BE}.
That changes I_C through the Ebers-Moll relation.

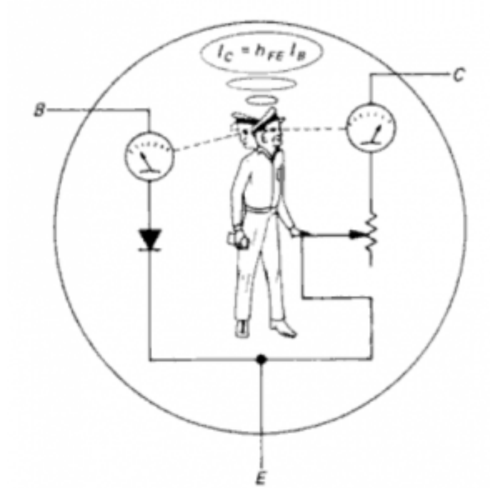
V_{BE}

$$I = I_S \left(e^{V/nV_T} - 1 \right)$$

So the transistor's "gain" is really $g_m = \Delta I_C / \Delta V_{in}$.

This is called transconductance because conductance is 1/resistance,
and the sub-m is short for mho, which is opposite of ohm.

The R_C and R_E used in a common emitter
amplifier convert that ΔI_C back into a ΔV_{out}.



Ebers-Moll model & transconductance

$$I = I_S \left(e^{V/nV_T} - 1\right) \cong I_S\, e^{V_{BE}/nV_T}$$

Strong function of V_{BE}, e.g.,

$\Delta V_{BE} = 18$ mV doubles I_C.

$\Delta V_{BE} = 60$ mV increases I_C by x10.

We can see where r_e comes from by calculating dI/dV.

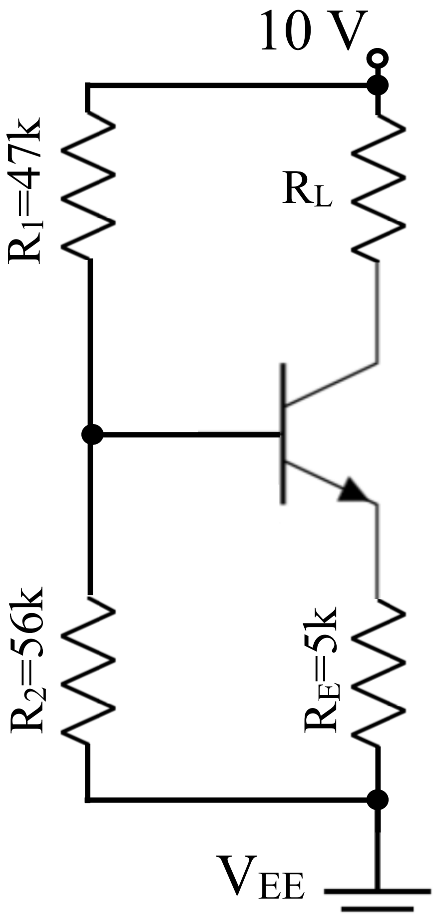
$$1/r_e = dI/dV = (1/nV_T)\, I_s\, e^{V_{BE}/nV_T} = I/nV_T$$

$r_e = nV_T/I = 25\text{ mV}/I = 25\Omega \,/\, I[\text{mA}]$ at room temperature

Ebers-Moll model & temperature effects

$$I = I_S \left(e^{V/nV_T} - 1\right) \cong I_s\, e^{V_{BE}/nV_T} \cong I_s\, e^{V_{BE}q/k_BT}$$

We can see the effect of temperature on our current mirror.



Set $V_B \cong 5.6$, so $V_E = 5.0$ V and $I_E = 1$ mA.

If temperature increases, $I_C = I_E$ reduces.

That reduces V_E

which increases V_{BE}

which increases I_C.

So we have "negative feedback" holding the circuit at an equilibrium behavior, insensitive to temperature.

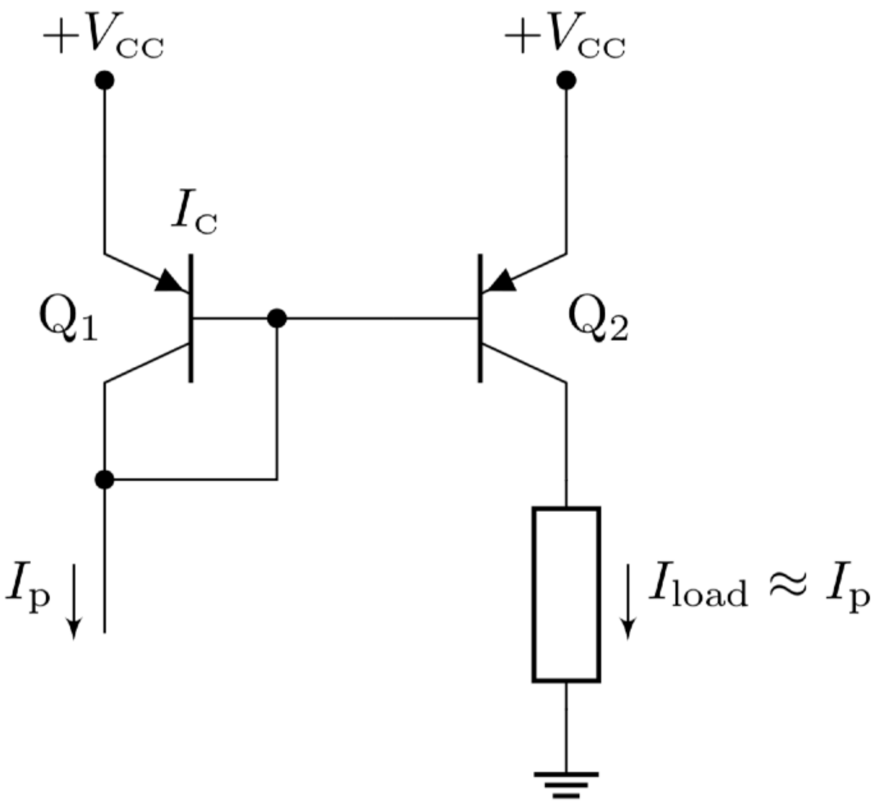
But, I_S is also temperature dependent, with opposite and stronger dependence, ~ 9%/°C.

So we need to build in more negative feedback.

Current mirror for temperature stability

$$I = I_S \left(e^{V/nV_T} - 1 \right) \cong I_s\, e^{V_{BE}/nV_T} \cong I_s\, e^{V_{BE} q/k_B T}$$

Set (program) Ip on left side.



Current mirror for temperature stability

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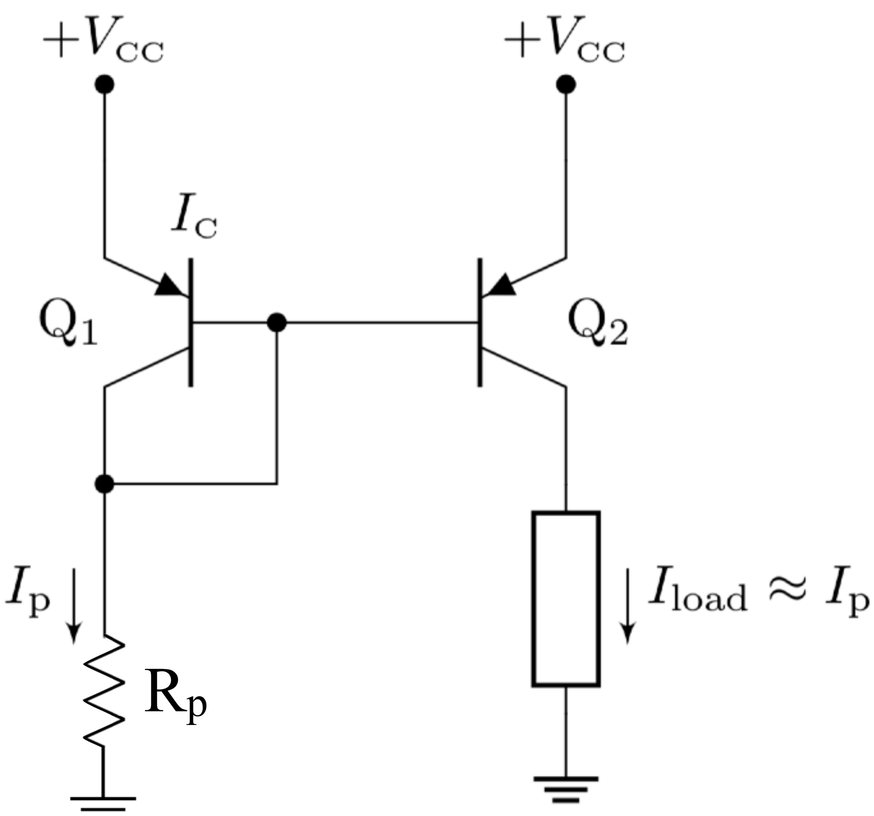


Set (program) Ip on left side.

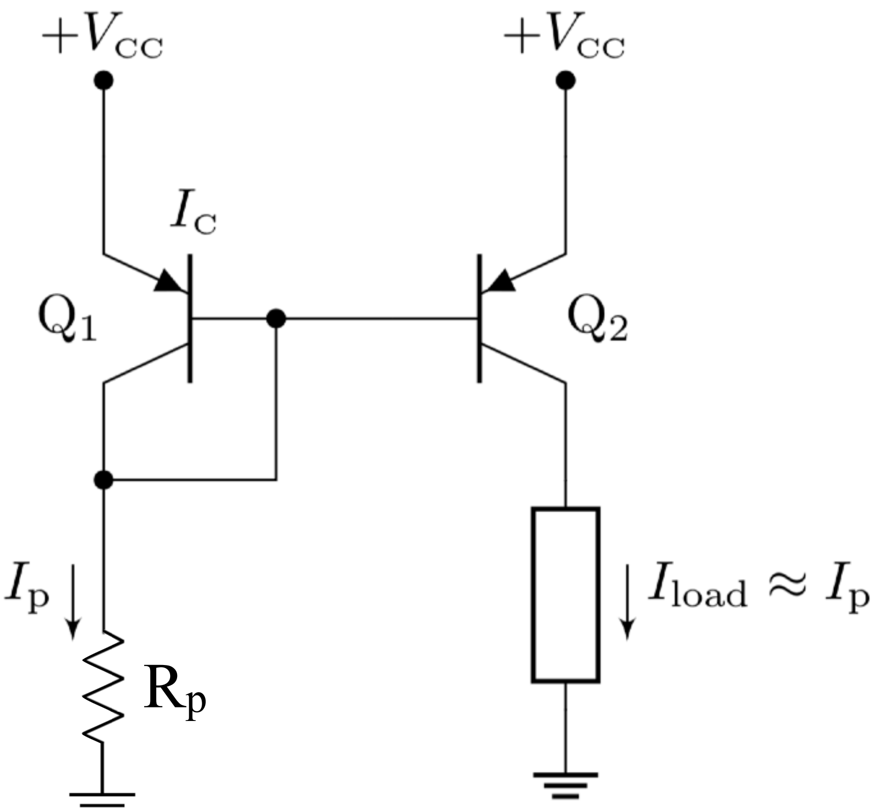
$V_B = V_{CC} - 0.6 = V_C$

$I_p = V_C/R_p$

$I_{load} = I_p$ because V_{BE} for Q_2 is the same as V_{BE} for Q_1.

Current mirror for temperature stability

$$I = I_S\left(e^{V/nV_T} - 1\right) \cong I_s\, e^{V_{BE}/nV_T} \cong I_s\, e^{V_{BE}q/k_BT}$$



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$V_B = V_{CC} - 0.6 = V_C$

$I_p = V_C/R_p$

$I_{load} = I_p$ because V_{BE} for Q_2 is the same as V_{BE} for Q_1.

Use matched transistors:

- same doping concentrations means same I_S, and $I_S(T)$;
- same substrate means same temperature.

Then they also have the same Ebers-Moll relation.

If some ΔT causes I_p to increase, then R_p pushes V_{BE} up to reduce I_p. Same change on both sides.

Differential amplifier

If we try to transmit a signal a long distance, we need to worry about RF pickup because the wires act as an antenna.

We could amplify the signal before transmitting to make it large compared to any pickup. But then it becomes a powerful transmitter causing pickup on other wires nearby.

Differential amplifier

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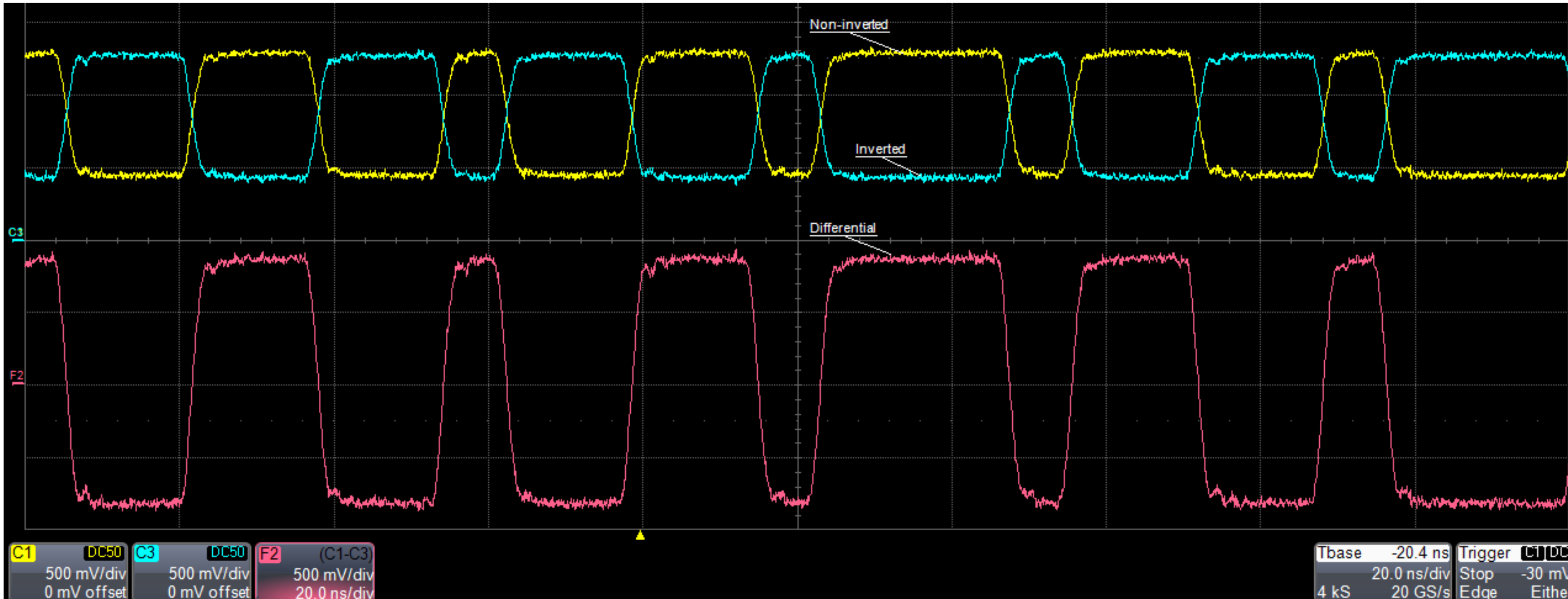
We could amplify the signal before transmitting to make it large compared to any pickup. But then it becomes a powerful transmitter causing pickup on other wires nearby.

Best to transmit signals with small signals that are immune to pickup; use low-voltage differential signals (LVDS) on twisted pairs of wires.

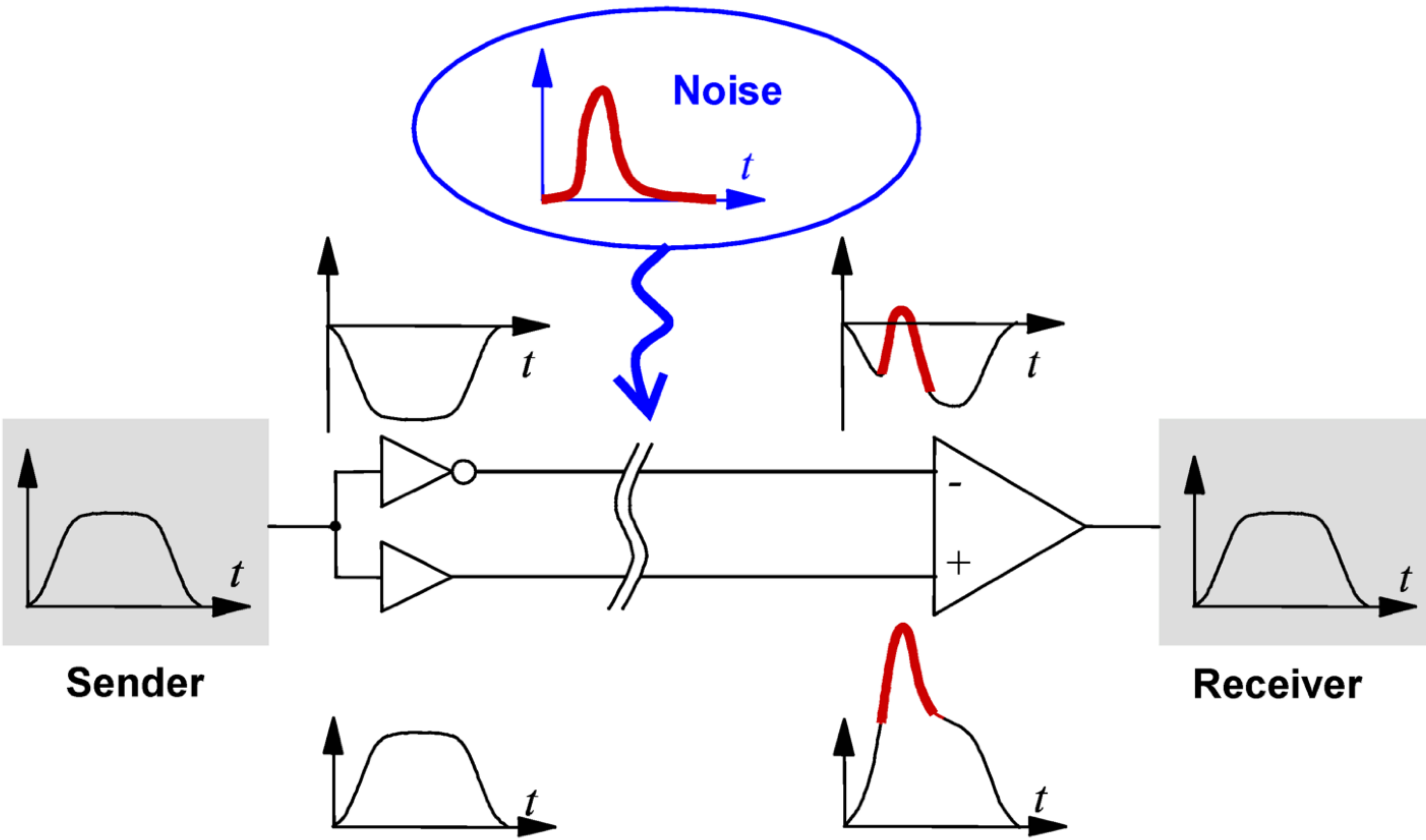
Differential amplifier



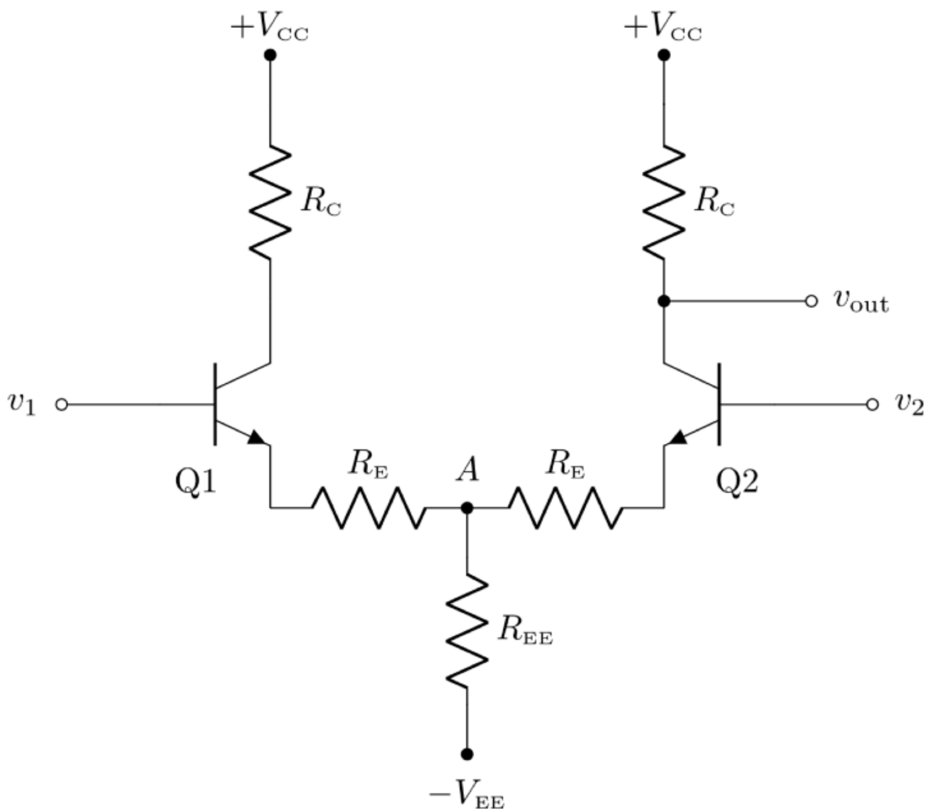
Signal+ = +ΔV/2

Signal- = -ΔV/2

Differential amplifier



Differential amplifier



Analyze this by 1st calculating V_A.

$$V_A = V_{EE} + I_{EE}R_{EE}$$

$$I_{EE} = I_{E1} + I_{E2}$$

$$= (V_{E1}-V_A)/R_E + (V_{E2}-V_A)/R_E$$

$$= (V_{E1}+V_{E2})/R_E - 2V_A/R_E$$

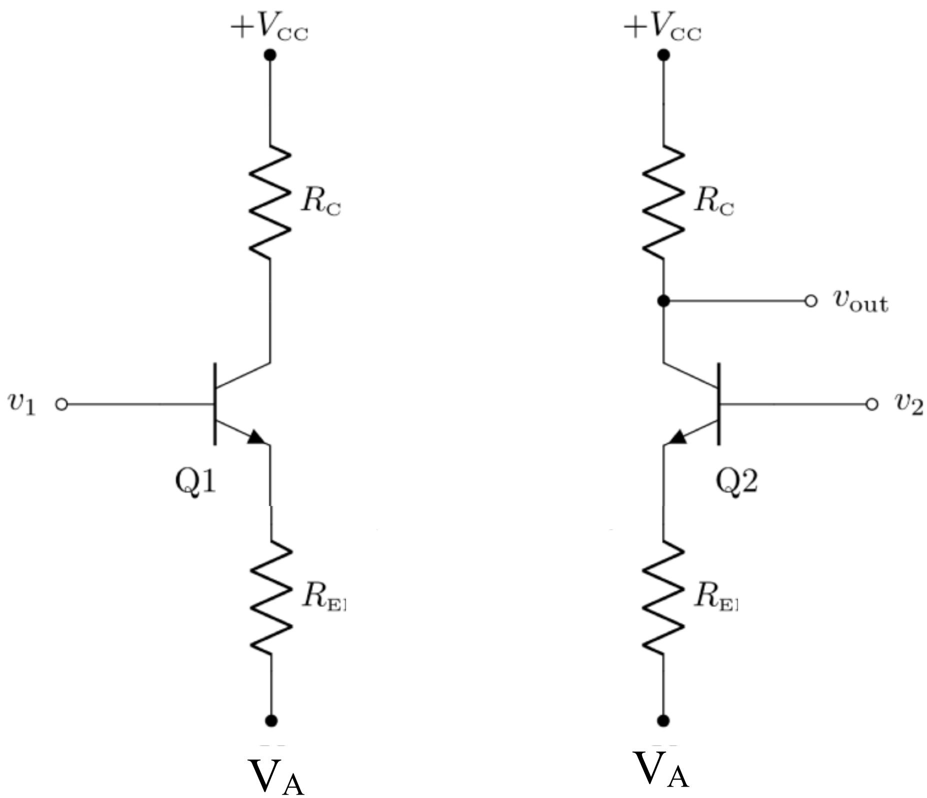
$$V_A = V_{EE} + R_{EE}/R_E\,(V_{E1}+V_{E2})/R_E - 2R_{EE}V_A/R_E$$

$$V_A = \frac{R_E V_{EE} + R_{EE}\,(V_{E1}+V_{E2})}{R_E + 2R_{EE}}$$

$$\Delta V_A = (\Delta V_{E1} + \Delta V_{E2})\,\frac{R_{EE}}{R_E + 2R_{EE}}$$

If $\Delta V_{E1} = - \Delta V_{E2}$ then $\Delta V_A = 0$

Differential amplifier



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$V_A = V_{EE} + I_{EE}R_{EE}$

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$= (V_{E1}-V_A)/R_E + (V_{E2}-V_A)/R_E$

$= (V_{E1}+V_{E2})/R_E - 2V_A/R_E$

$$V_A = V_{EE} + R_{EE}/R_E\,(V_{E1}+V_{E2})/R_E - 2R_{EE}V_A/R_E$$

$$V_A = \frac{R_E V_{EE} + R_{EE}\,(V_{E1}+V_{E2})}{R_E + 2R_{EE}}$$

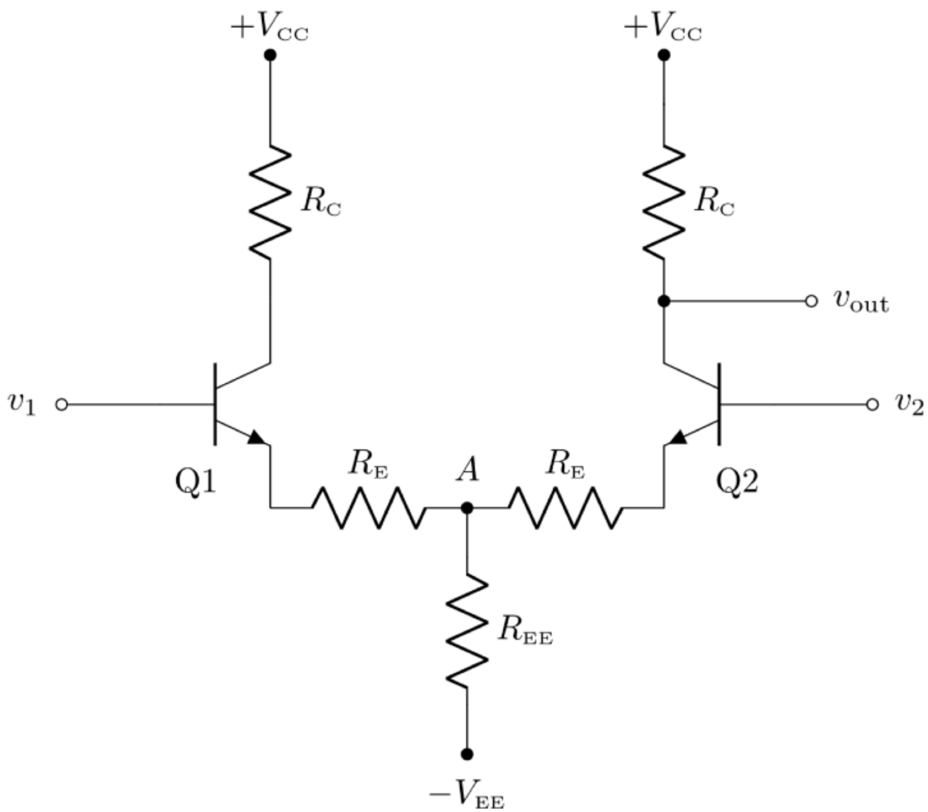
$$\Delta V_A = (\Delta V_{E1} + \Delta V_{E2})\,\frac{R_{EE}}{R_E + 2R_{EE}}$$

If $\Delta V_{E1} = -\Delta V_{E2}$ then $\Delta V_A = 0$

This makes the right side just a common-emitter amp with $v_{out} = (-R_C/R_E)\,v_2$

If $v2 = -\Delta V_{in}/2 = -v_{in}/2$ then $v_{out} = (R_C/R_E)v_{in}$.

Differential amplifier



Now consider the *common mode* signal, where $v_1 = v_2 = \overline{v} = v_{CM}$

That makes $\Delta I_{E1} = \Delta I_{E2}$ & $\Delta I_{EE} = 2\Delta I_{E1}$

Written with "variation notation" its
$i_{E1} = i_{E2}$ and $i_{EE} = 2i_{E1}$

So, $\Delta V_A = v_A = i_{EE}R_{EE} = 2i_{E1}R_{EE}$
Now use Ohm's law to find iE1 as
$i_{E1} = (v_E - v_A)/R_E$
$= (v_{CM} - 2i_{E1}R_{EE})/R_E$
So,
$i_{E1} = v_{CM}/(R_E+2R_{EE})$

$v_{out} = -i_{E1}\ R_C = -\ v_{CM}\,R_C/(R_E+2R_{EE})$

Common mode gain = $-R_C/(R_E+2R_{EE})$
Differential gain = $-R_C/2R_E$