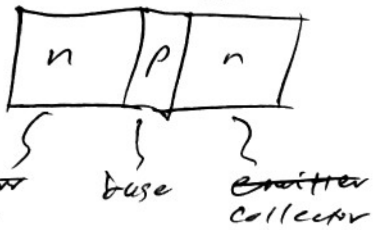
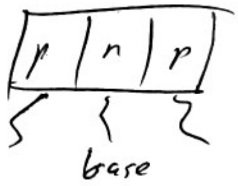
Transistors ("transfer resistors")

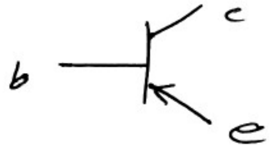
We will largely follow Ch. 2 of Horowitz & Hill, not Eggleston.

For a basic junction transistor, structure is



or





Basic model: $\boxed{I_C \simeq \beta I_B}$

In data sheets, usually written $\boxed{I_C \simeq h_{FE} I_B}$ ← typ. ~100, wide range.

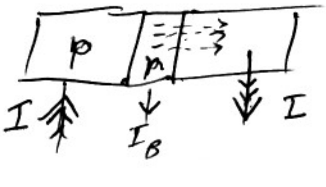
if the following are true (written for npn)

1) c is more (+) than e

2) b-e is conducting, but b-c diode is reverse-biased

3) There are max. values for I_c, I_B, V_{CE}, V_{BC}, ... which ~~can~~ are not ~~be~~ exceeded.

In general terms, e-b conduction injects charge carriers ---



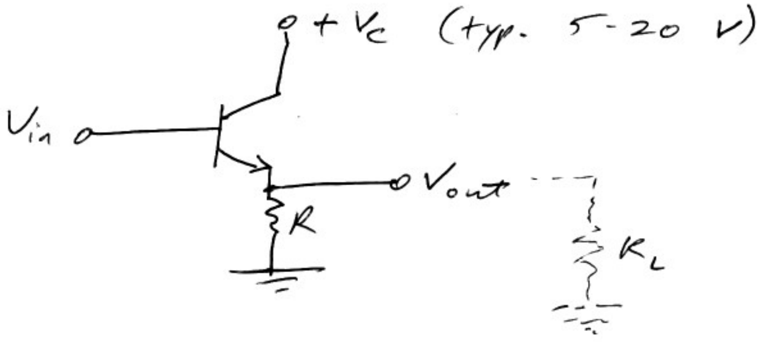
Not same as two diodes --- most of E-B- internal current continues to the collector.

Can read about physics in other texts; most discuss it

Note: can get $V_c - V_E$ as low as 0.1 - 0.2 V
saturation voltage

Emitter follower:

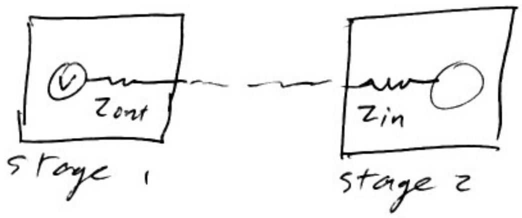
Basic circuit



Once biased into conduction at the B-E junction, the transistor acts to preserve the relation

$$V_{in} - V_{out} = V_B - V_E \simeq 0.6 \text{ V}.$$

Utility: can drive large ~~inp~~ output current with a small input current. No voltage gain, but it's an impedance transformer that can have large power gain:



To simplify design, we normally want $Z_{2,in} >> Z_{1,out}$ (by at least a factor of ten or so; assumes voltage sources.)

To analyze the response to a small change at the input, it's convenient to break up the voltages and currents into two pieces.

Introducing lower-case notation for small changes or ac signals,

$$V_{B(t)} = V_B^{DC} + v_B(t)$$

$$V_{C(t)} = V_C^{DC} + v_C(t)$$

$$V_{E(t)} = V_E^{DC} + v_E(t)$$

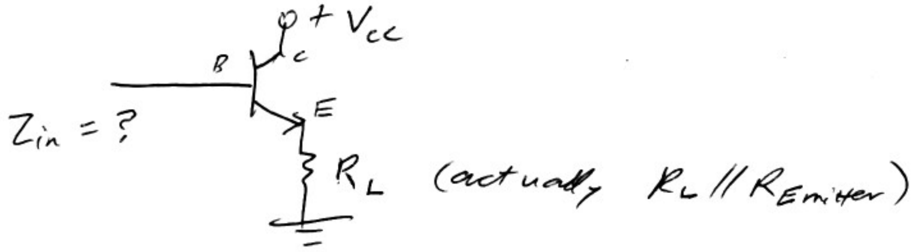
small change = signal, often time-dependent.

steady-state dc value ("quiescent" or "bias" voltage = <u>operating point</u>)

likewise $I_{B(t)} = I_B^{DC} + i_B(t)$, etc.

Can also write as $v_B = \Delta V_B$, etc.

Now back to the emitter follower, with our super-simplified model:



① $I_E = I_B + I_C \simeq I_B(1+\beta)$

② Find $Z_{in} = \frac{v_B}{i_B}$ for small changes v_B.

Since $V_E \simeq V_B - 0.6\,V$,

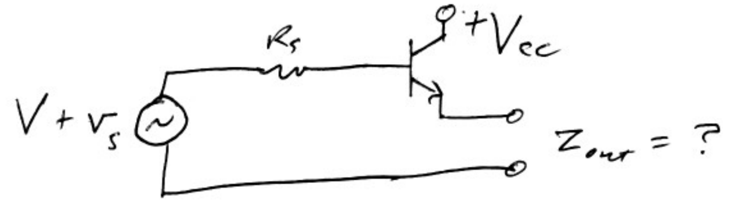
$\Delta V_E = v_E \simeq v_B$.

Also, $i_E = \frac{v_E}{R_L}$ and from ①, $i_E \simeq i_B(1+\beta)$

So $Z_{in} = \frac{v_B}{i_B} \cong \frac{v_E}{i_E/(1+\beta)} = \frac{v_E(1+\beta)}{v_E/Z_L}$

$\boxed{Z_{in} \cong R_L(1+\beta)}$ for emitter follower (increased by up to ~100x!)

There's a similar beneficial effect as viewed from the output:



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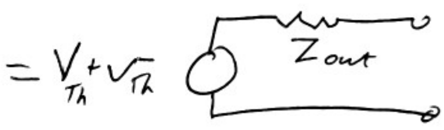


For small changes (ac response), $Z_{out} \equiv \frac{v_{Th}}{i_{\text{"short-circuit"}}}$ (i.e., Thévenin equivalent Z)

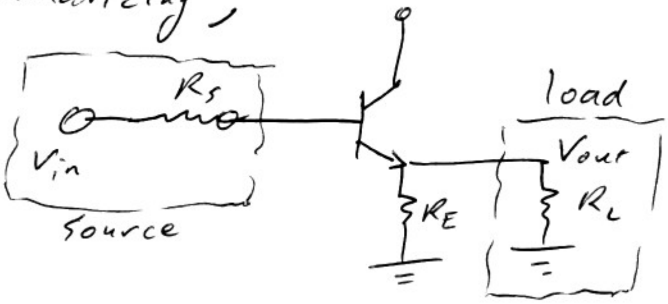
where $v_{Th} = v_E \cong v_s$

and $i_{\text{short-circuit}}$ is actually i_E with V_E fixed at V_E^{DC}. (i.e., ac short-circuit)

Then $i_{s-c} \cong i_B(1+\beta) = \frac{v_s}{R_s}(1+\beta)$, and

$Z_{out} = \frac{v_s}{\frac{v_s}{R_s}(1+\beta)}$; $\boxed{Z_{out} = \frac{R_s}{1+\beta}}$

Summarizing,

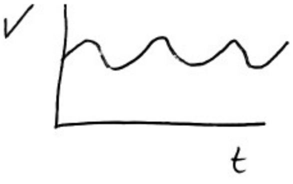


1) $V_{out} \simeq V_{in} - 0.6\,V$

2) $Z_{out} \simeq \frac{R_S}{1+\beta} \left(// R_E, \underset{\text{usually irrelevant}}{} \right)$

3) $Z_{in} = \underbrace{(R_L // R_E)(1+\beta)}_{\simeq R_L \beta \text{ under most conditions.}}$

Example: improved power supply ---



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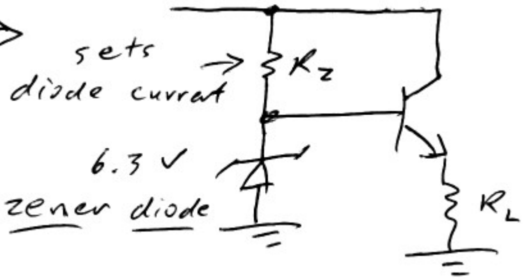
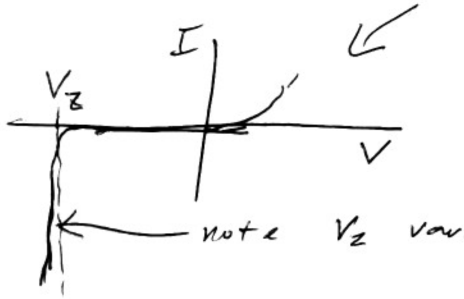
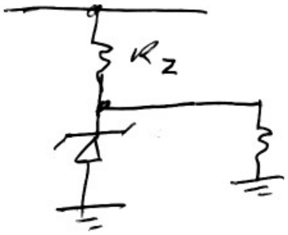




note V_z varies slightly with I_z

From the perspective of the zener diode,



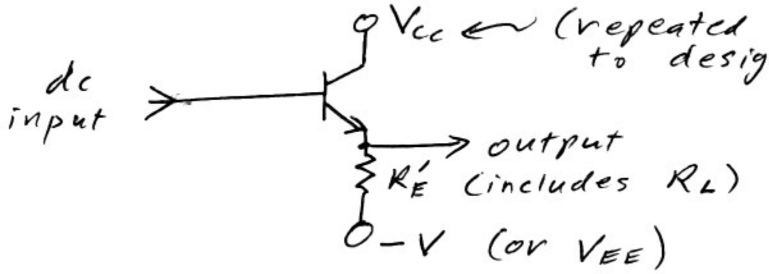
$Z_{in} = R_L(1+\beta)$ has reduced effect on I_Z as R_L varies.

⇒ Regulation improves by ~factor of β compared with circuit without follower.

⇒ Can use much lower current through zener diode: more thermally stable, does not need high-current rating.

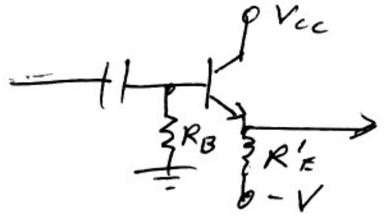
Biasing: What if input isn't always at ~ a few Volts?

Solution 1: Bipolar power supplies



(repeated subscripts often used to designate power supplies)

or



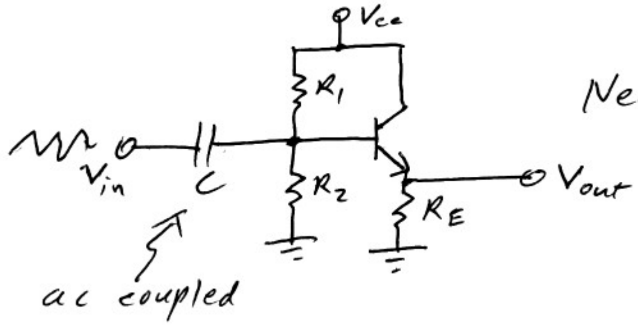
Here R_B supplies base current. To maintain a stable operating point, but avoid loading the input, make $R_B \lesssim \frac{1}{10} Z_{follower}$

$\simeq \frac{1}{10}(\beta+1) R_E'$

For $\beta \sim 100$, this yields $Z_{in} \sim 10 R_E'$.

Solution 2: Biasing circuit, most often a "stiff voltage divider"

Make $Z_{divider} << Z_{driven} \simeq \beta R_E'$



Need $R_1 // R_2 << \beta R_E'$

Design process:

① Make $R_{in} C >>$ slowest τ_{signal},

where $R_{in} \equiv R_1 // R_2$, since $R_{follower}$ is large by comparison.

② Make $\frac{R_2}{R_1 + R_2} \sim \frac{1}{2}$ to center range at $V_{cc}/2$.

i.e., pick $R_1 \simeq R_2$.

③ choose R_E for a small quiescent current,
but not so large that you can't
drive the load (can omit R_E if load fixed, always connected.)

If we have $V_c = 10V$, $R_E = 10k$,
then $I = 1mA$ max (ignoring Z_{load}) or 0.5 mA quiescent,
Pick $R_1 = R_2 = 10K$; note Z_{in} of transistor
is ~ 100 x 10K

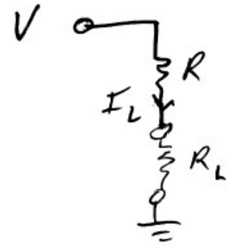
If 200 Hz is lowest input freq., need

$$RC >> \frac{1}{300(2\pi)} \quad \text{or} \quad \frac{10^4}{2} C >> \frac{1}{2000}$$

$$C >> 0.1 \mu F \quad (\text{pick } .47\mu F)$$

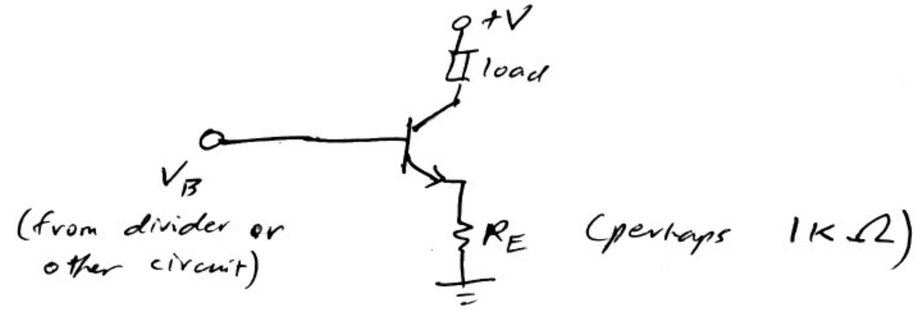
(we could make R_1, R_2 larger, but not *too* large.)

Current source:



I_L is very dependent on load
unless $R >> R_L$.
⇒ poor except at *very* low current.

Instead use



We have ~~∀~~ $I_c \cong I_E$ if β is large,
since $I_B = I_c / \beta$.

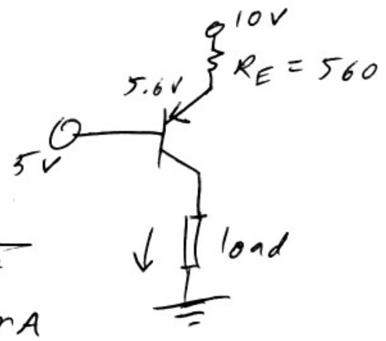
Also $V_E \cong V_B - 0.6V$, so

$$I_c \simeq \frac{V_E}{R_E} \cong \boxed{\frac{V_B - 0.6V}{R_E}} \quad (\simeq I_E).$$

sat. volt. drop.

Compliance range set by $V_c > V_E + 0.2V$; $V_E = I_E R_E$.

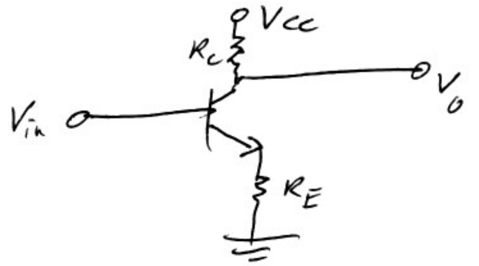
While this version <u>sinks</u> current, a pnp variation can <u>source</u> it—



$$I = \frac{4.4\,V}{560\,\Omega} = 7.9\ mA$$

Common emitter amplifier:

How to build a <u>voltage</u> amplifier? One way is to use a current source with a resistor as its load:



Here $V_o = V_c = V_{cc} - I_c R_c$ ① and $I_c \simeq I_e \overset{②}{=} \frac{V_B - 0.6V}{R_E}$

For a small change v_B in the input

$$V_B = V_B^{DC} + v_B),$$

I_c changes by $i_c = \frac{v_B}{R_E}$ per ② and

V_c changes by $v_c = -i_c R_c$ per ①.

Thus $v_c = -v_B \frac{R_c}{R_E} \Rightarrow$ gain $\boxed{G = \frac{v_o}{V_{in}} \simeq -\frac{R_c}{R_E}}$

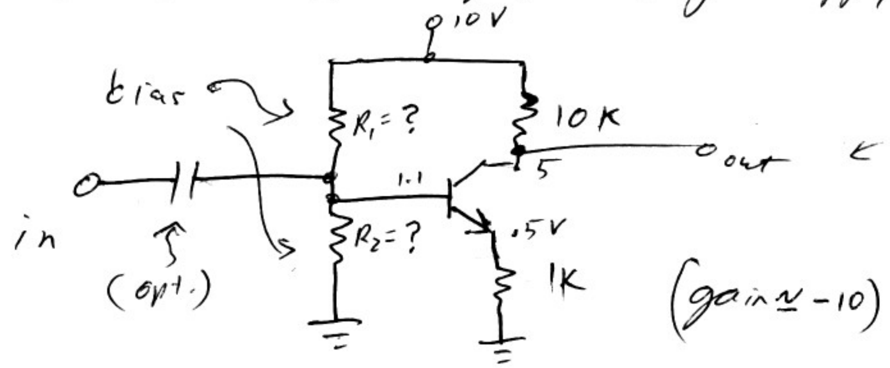
(called 'a' by Eggleston!)

Should be accurate if β is large and

1) Input is a small change in the quiescent operating point, and

2) R_E is not so small that something else limits G (more on this shortly).

Practical common emitter amplifier (single-supply):



← set $V_{quiescent} \cong 5V$

(gain ≅ −10)

Can this be done using 1K and 10K, as shown?
We need output ~5V when $V_{in} = 0$, to get max. swing.

To do this, we want 5V drop across 10K, or $I_c^{DC} = 0.5mA$

This requires $V_E^{DC} \sim 0.5V$ (gives 0.5 mA into 1K),
or $V_B^{DC} \sim 1.1V$. So $\frac{R_2}{R_1+R_2} = \frac{1.1}{10} = \frac{11}{100}$.

What are good values? We need
1) High input Z, but
2) operating point reasonably independent of input, and especially of base current.

Impedence looking into base is $\sim R_E \beta$, or around 100K. So we need $R_1 // R_2 << 100K$.
A reasonable choice is $R_2 = 10K$, so $R_1 = 80.9K$ would be ideal.

Quiescent current is just 0.5mA, so nothing heats up much.
But ~~max~~ current is only ~1mA, so not exactly a huge drive here.

Input impedence: $Z_{in} = R_1 // R_2 // (R_E(1+\beta))$, just under 10K

Output impedance: looking into collector it's $M\Omega$, so result is $Z_{out} = 10K$. (High Z is OK only if load is also high-Z.)

Looking into emitter: ~~hi~~ low Z
collector: high Z (current source)

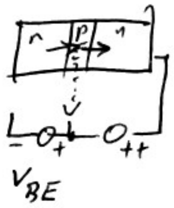
Why can't we make gain arbitrarily high by varying resistor ratios R_C / R_E? Of cource, it's mostly because our transistor model is so bad:

Ebers-Moll eq'n ; Improved model (residual problems summarized later.)

Recall for a diode, $I = I_s \left(e^{\frac{+V}{C}} - 1 \right)$.
Zero-order model: take $\Delta V \sim 0.6\,V$ always.
Not surprisingly, a qualitatively similar situation is a better view for a transistor, too.

Think of circuit in new terms -- as a transconductance amplifier --

$$g_m = \frac{I_{out}}{V_{in}}$$



The improved description is:

$$I_E \cong I_s \left(e^{\frac{V_{BE}}{V_T}} - 1 \right)$$

where $V_T = \frac{k_B T}{e} = 25.3\,mV$ at 20°C, just like diode but with $n = 1$.

and I_s = saturation current, varies with transistor <u>and</u> $\underline{\underline{T}}$.

So $$I_E = I_s \left(\frac{V_{BE}}{V_T} + \frac{1}{2}\left(\frac{V_{BE}}{V_T}\right)^2 + \cdots \right)$$

If we also assume $I_B \propto V_{BE}$,
I_E prop. to I_B ~~for~~ to lowest order.

So we see now that β depends on I_C, T, and also, it turns out, V_{CE}!

Main effect --- There is an effective resistance r_e looking into the emitter, intrinsic to the device.

Intrinsic emitter resistance — for small signals, use E-M model to find $\frac{i_E}{v_{BE}} = \frac{d I_E}{d V_{BE}} \cong I_s \frac{1}{V_T} e^{V_{BE}/V_T}$

Since $e^{V_{BE}/V_T} >> 1$ in normal use,

$$\frac{i_E}{v_{BE}} \simeq \frac{I_E}{V_T} \sim \left(\simeq I_s e^{V_{BE}/V_T}\right)$$

Thus the effective resistance is the reciprocal of this,

$$r_e = \frac{v_{BE}}{i_E} \simeq \frac{V_T}{I_E}$$, or, since $I_C \sim I_E$ in most cases,

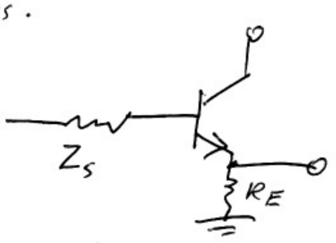
or $$\boxed{r_e \simeq \frac{25\,\Omega}{I_C \text{ (in mA)}} = \frac{0.025\,\Omega}{I_C \text{ (in A)}}}$$ (using $V_T \sim 25$ mV)

r_e appears in series with any external impedance.

Secondary consequences: Because I_S varies with T as well as with V_{BE}/V_T, it turns out that for constant I_C, $V_{BE} \propto 1/T$, and it <u>decreases</u> by $\simeq 2.1$ mV/°C for Si transistors. (Thus thermal runaway becomes a concern!)

Effects on our circuits:

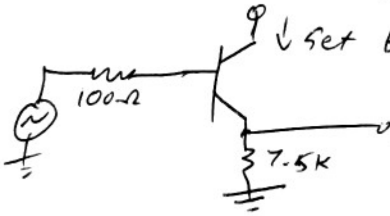
i) Emitter follower:



$$Z_{out} = \left(\frac{Z_s}{\beta+1} + r_e\right) // R_E$$

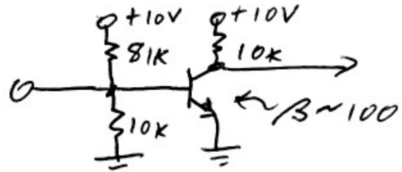
almost never important

For example,



$Z_{out} \simeq 25\,\Omega$, not $\frac{100\,\Omega}{100} \sim 1\,\Omega$.

2) Common-emitter amplifier – consider grounded emitter:



Gain ≠ ∞ !
At quiescent point, 5 V out corresponds to 0.5 mA, so $r_e = 50\Omega$,

$$G = \frac{10k}{50} = 200, \text{ not } \infty.$$

Also, because I_c varies widely with signal, gain varies from 0 to 400 over the full output range!

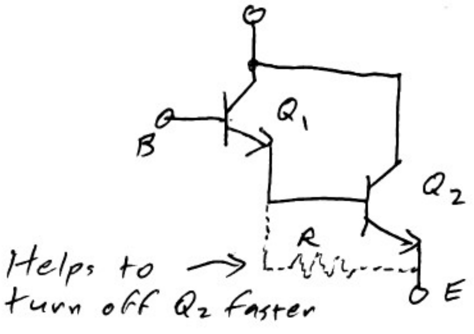
Further, input impedance βr_e is unstable, and base is very hard to bias – for fixed V_{BE}, I_C will rise tenfold for 30 C of heating!

Solutions:

1) Add emitter resistors as before with $R_E >> r_e$, defining quiescent current and limiting gain.
2) Bypass emitter resistor, decoupling gain from dc bias. ⇒
3) Feedback – see Horowitz and Hill.

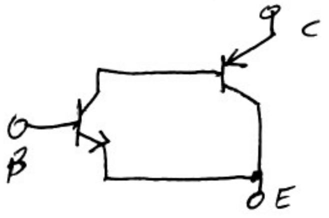
Other bipolar transistor circuits

Darlington:



Readily available pre-packaged.
$\beta \simeq \beta_1 \beta_2$
(but drops at least 1.2 V on base)

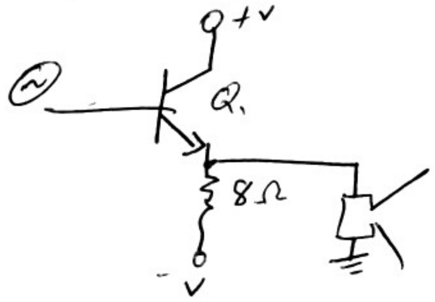
Another variation is



(Behaves as an npn transistor)

Push-pull power amps:

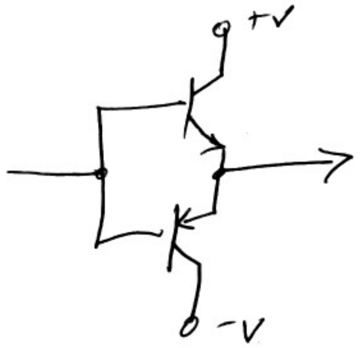
It's very inefficient to use a single transistor for high-power applications:



need quiescent current ≥ max output current.

resistor, Q₁ both dissipate more power than load.

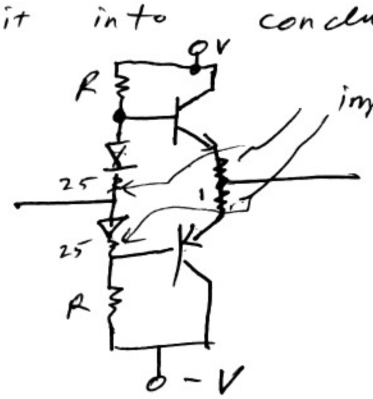
Instead, try



"Class B"

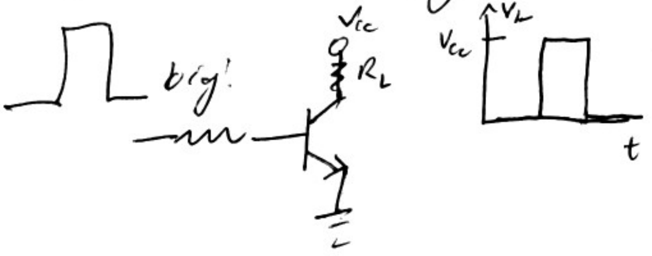
Works OK, but clips near 0V:

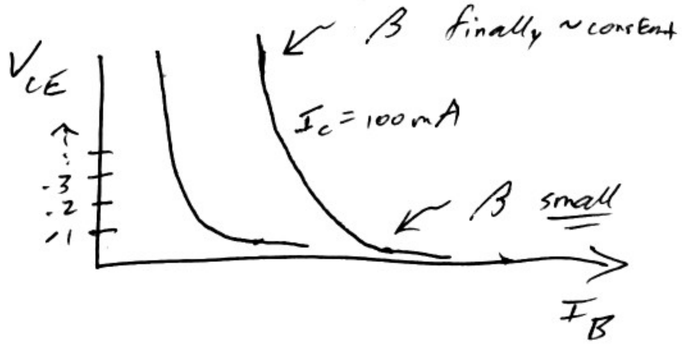
So bias it into conduction --



improves thermal stability -- Change in V_{BE} doesn't change quiescent current as quickly

Saturated switching:





- For high power, use lots of base current (C-B junction is starting to conduct, reducing β.)
- For fast switching, don't turn on or off quite so hard.