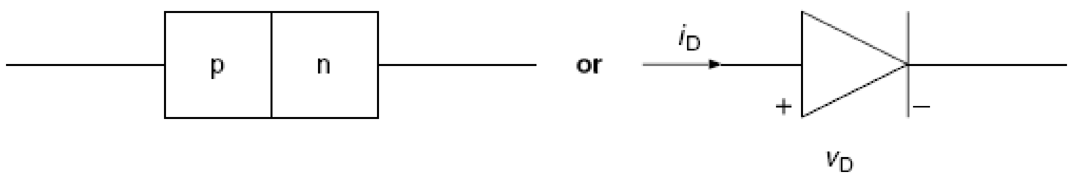
Active devices and amplifiers

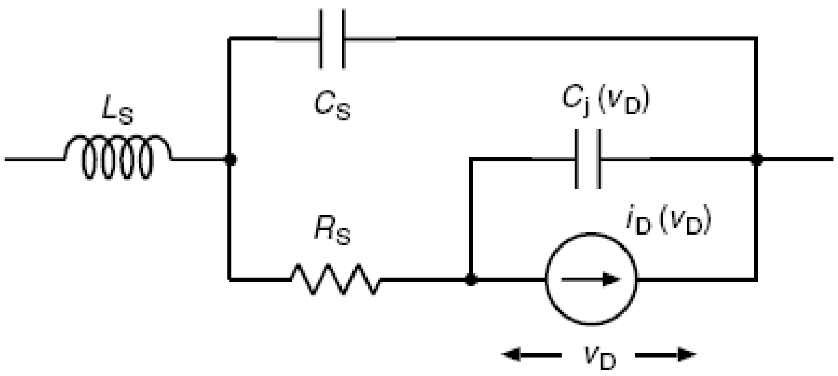
- Active devices are important elements in RF systems where they perform functions such as amplification, mixing and rectification.
- For mixing and rectification, the non-linear properties of the device are of paramount importance.
- In the case of amplification, however, the non-linear properties can have a damaging effect upon performance.
- The high frequency performance of transistor amplifiers is limited by what is known as the Miller effect and a large part of the chapter is devoted to techniques for overcoming this phenomenon.

The semiconductor diode

- The current i_D in a diode is related to the potential difference v_D across the diode through

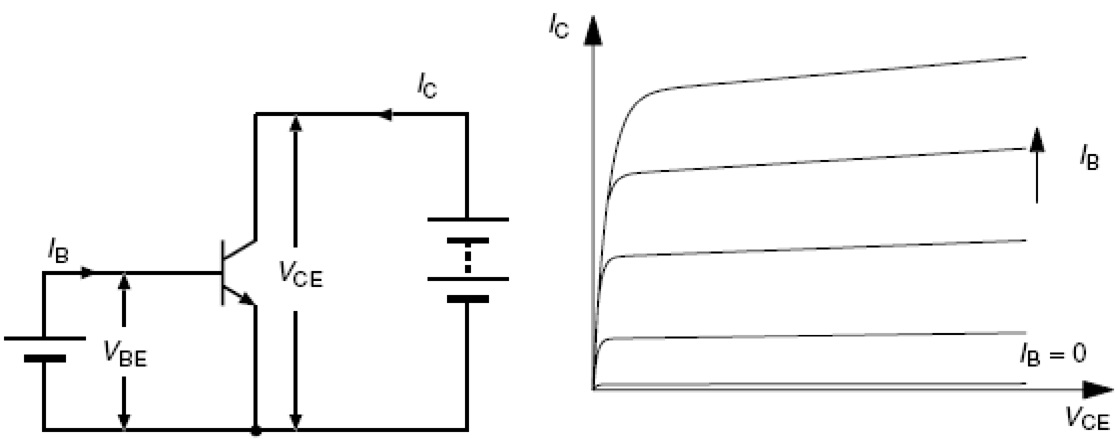
$$i_D = I_s \left[\exp\left(\frac{v_D}{n V_T} \right) - 1 \right]$$





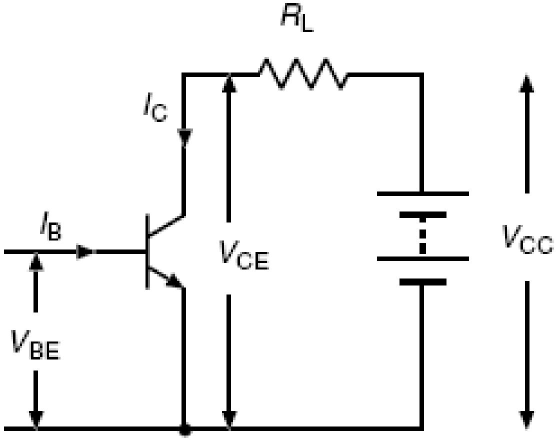
Bipolar junction transistors

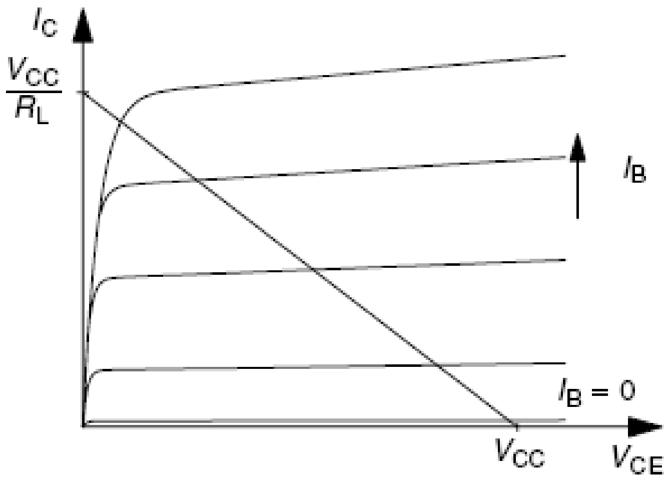
- A bipolar junction transistor (BJT) is a device consisting of a sandwich of n- and p-type semiconductors (npn or pnp) that acts as a current amplifier.
- If the base current I_B is fixed, there is a level of emitter–collector voltage (V_{CE}) above which the relationship between collector current I_C and V_{CE} is fairly linear.
- Furthermore, in the active region, there is a fairly linear relationship between the collector and base currents for a fixed collector voltage.
- As a consequence, the collector and base currents are related through $I_C = \beta I_B$ where β is known as the *current gain*.
- The emitter–base junction will act as a diode and so the base current, and hence the collector current, will be exponentially dependent upon the emitter–base voltage V_{BE}.



Bipolar junction transistors

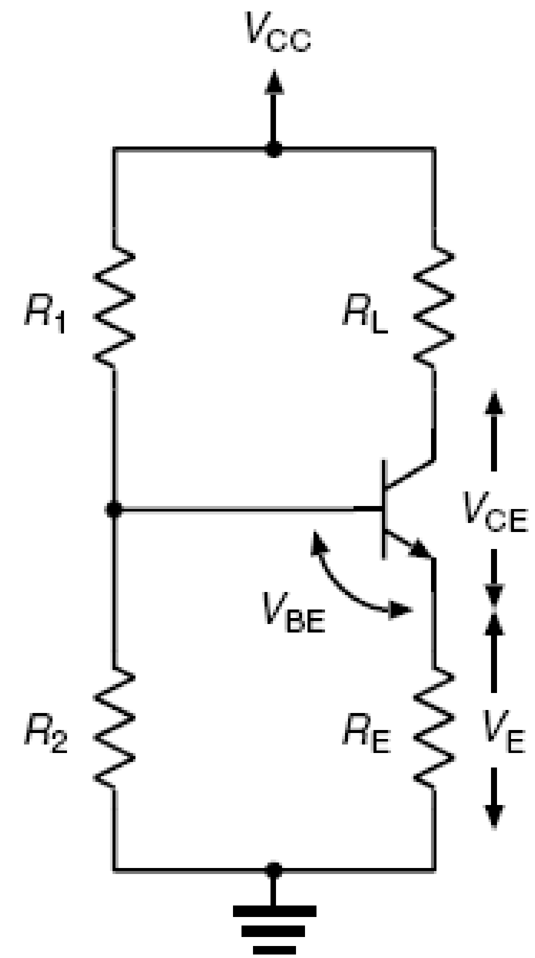
- For a given base current (I_B), the collector current will be given by the intersection of the *load line* $I_C = (V_{CC} - V_{CE})/R_L$ and the relevant transistor characteristic curve.
- If we suitably choose the load R_L and intersection, there will be a large variation in collector voltage for only a small variation in base–emitter voltage (voltage amplification).
- For amplification, we will normally select a quiescent collector current I_C such that voltage fluctuations at the base are linearly transformed to voltage fluctuations at the load.





Bipolar junction transistors

- The simplest bias configuration for small signal applications is the 4 resistors net.
- RE provides the negative feedback that stabilises the amplifier bias against variations in transistor properties.
- β can vary considerably between device samples and the VBE can vary significantly with temperature.



- The voltage drop across resistor $R2$:

  $VBE + (IC + IB)RE$

  $I2 = [VBE + (IC + IB)RE]/R2$

- The current through resistor $R1$:

  $I1 = I2 + IB$

  $VCC = I1R1 + I2R2$, we obtain:

  $VCC = R1\ IB + I2(R1 + R2)$

  $= R1\ IB + (R1 + R2)[VBE + (IC + IB)RE]/R2$.

Bipolar junction transistors

- As a consequence

$$R_B V_{CC} - R_1 V_{BE} = R_1 R_B I_B + (I_C + I_B) R_1 R_E$$

$$I_C = \frac{V_{CC}\frac{R_B}{R_1} - V_{BE}}{R_E + \frac{R_E + R_B}{\beta}}$$

- We need $R_E >> (R_E + R_B)/\beta$ for I_C to be insensitive to variations in the current gain β, that is $R_B << R_E(\beta - 1)$, thus.

$$\Delta I_C \approx -\frac{\Delta V_{BE}}{R_E}$$

- I_C will be related to the base–emitter voltage V_{BE} through $I_C \approx \beta I_s \exp(V_{BE}/V_T)$.
- Since V_T is temperature dependent, I_C is temperature dependent.

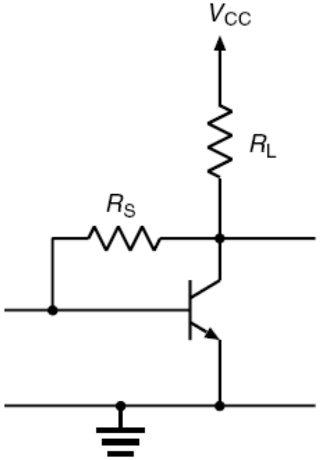
Bipolar junction transistors

Biasing rule:

- The stabilizing is achieved with V_E ($\approx I_C R_E$) of 2V or more.
- $R_2/(R_1 + R_2)$ is thus fixed by $(V_{BE} + V_E)/V_{CC}$ ($V_{BE} \approx 0.7$V for a silicon transistor).
- Although we need to keep R_B small in order to reduce sensitivity to variations in β, too small a value for this base resistance will end up reducing the gain of the amplifier. The usual compromise is to choose resistors R_1 and R_2 such that they carry about one tenth of the current that R_E carries.
- Consequently, once the quiescent collector current I_C is known, it is possible to complete the specification of R_1 and R_2.
- The value of the load resistance R_L needs to be chosen so that it causes a voltage drop of less than $V_{CC} - V_E$. The one-third rule, whereby V_{CE} and V_E are both approximately $V_{CC}/3$, is usually a good compromise.

Bipolar junction transistors

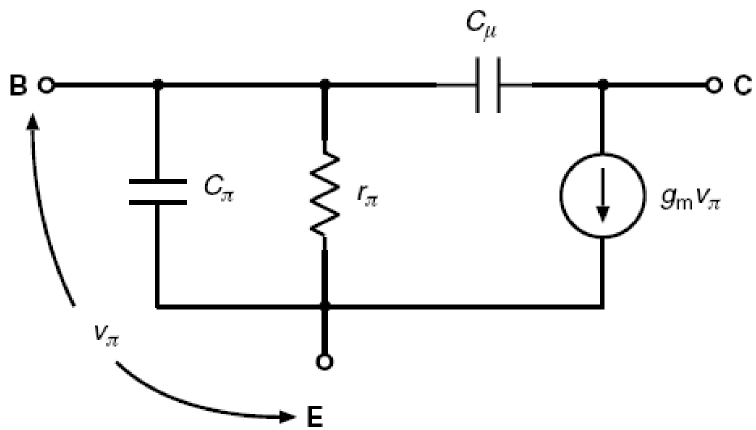
- This alternative topology does not provide stabilization.



- $I_B = I_C/\beta$; the total current through R_L is $I_C + I_B = I_C(1 + 1/\beta)$ and hence: $R_L = (V_{CC} - V_{CE})\beta / I_C(\beta + 1)$.
- In addition, we note that $R_S = (V_{CE} - V_{BE})/I_B = (V_{CE} - V_{BE})\beta/I_C$.
- In order to allow the maximum voltage fluctuation, we normally set $V_{CE} \approx V_{CC}/2$.
- Consequently, since $V_{BE} \approx 0.7$V for silicon devices, R_L and R_S can be determined once the supply voltage V_{CC} and collector current are known.

BJT model

- In order to analyze the behavior of the amplifier, we need a transistor model.



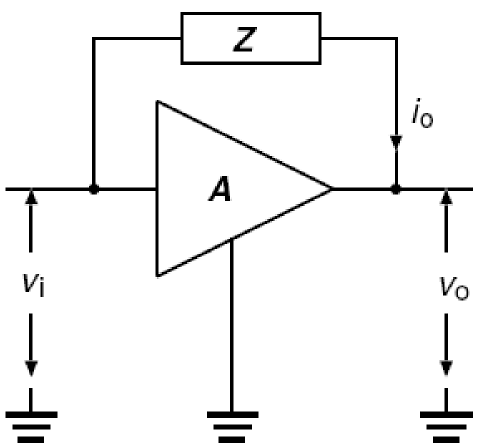
- *transconductance g*m can be derived from the expression $g_m r_\pi = \beta$ (i.e., $gm \approx IC/VT$).
- *Cπ* and *Cμ* are usually of the order of a few picofarads.
- The transistor capacitances cause the current gain to fall as frequency rises and the cut-off frequency *f*T (the frequency at which the short circuit common-emitter current gain of the transistor is unity)
- Parameter *f*T is related to *Cπ* and *Cμ* through

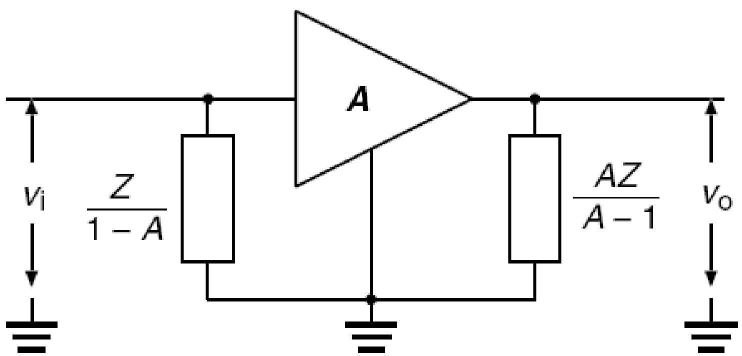
$$f_T = g_m / 2\pi (C_\mu + C_\pi)$$

The Miller effect and BJT amplifiers

- The *Miller result* is extremely useful tool for understanding the constraints on transistor performance at higher frequencies.
- For an amplifying device, it relates a feedback impedance to equivalent input and output impedances (infinite input impedance and zero output impedance).

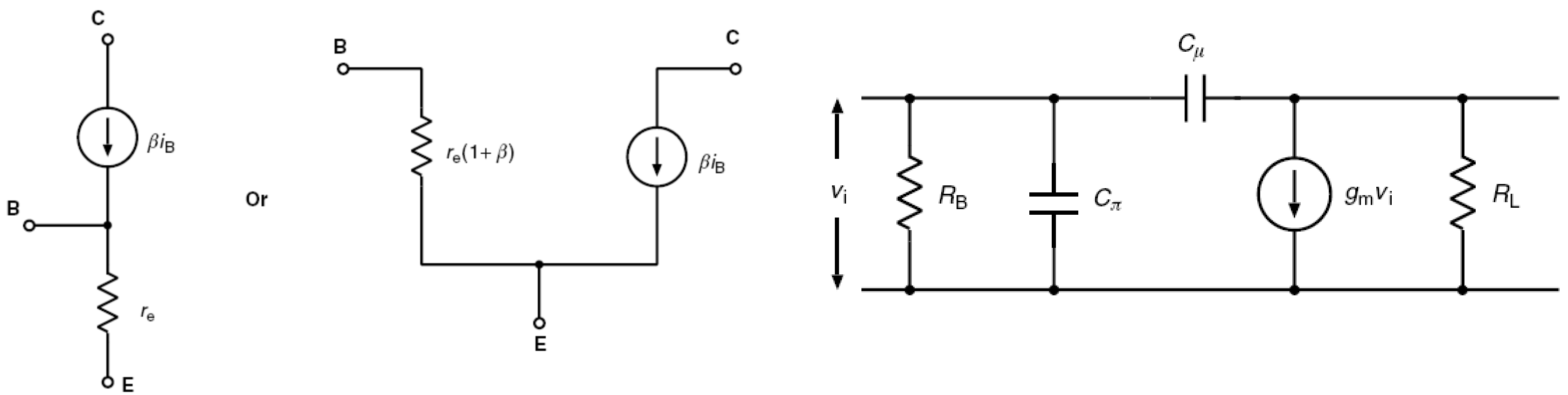
$$v_i = v_o + Z i_o \qquad v_o = A v_i$$



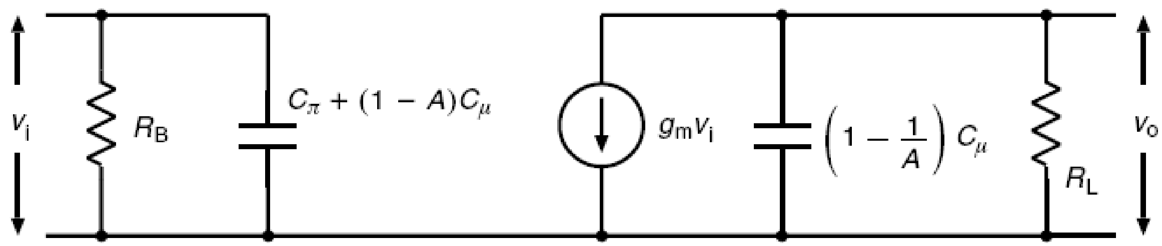


The Miller effect and BJT amplifiers

- The RF behaviour of the common-emitter BJT amplifier can be analysed by the small-signal model and Miller theorem



- The low freqeuncy gain is $A \approx -R_L g_m$



$$A_v = \frac{v_o}{v_i} = -\frac{g_m R_L}{1 + j\omega C_\mu R_L}$$

The Miller effect and BJT amplifiers

- The input and output impedance are frequency and gain dependent

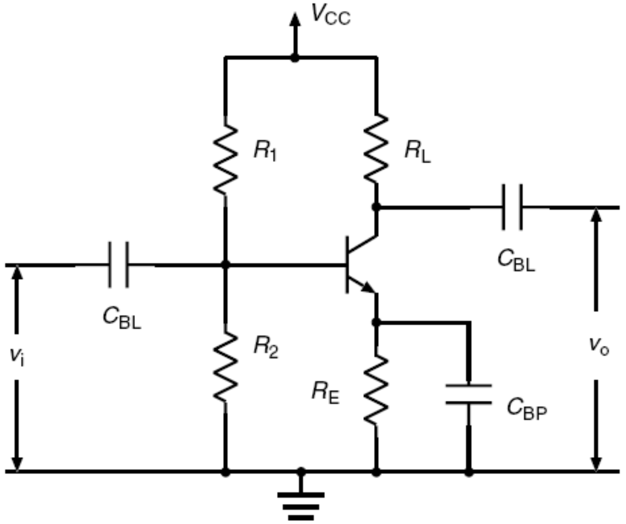
$$Z_{\text{in}} \approx R_{\text{B}} \parallel \frac{1}{\text{j}\omega[C_\pi + (1 - A)C_\mu]} = \frac{R_{\text{B}}}{1 + \text{j}\omega R_{\text{B}}[C_\pi + (1 - A)C_\mu]}$$

$$Z_{\text{out}} \approx \frac{1}{\text{j}\omega C_\mu} \parallel R_{\text{L}}$$

- *Z*in is the strong frequency dependent by the *Miller capacitance* (1 − *A*)*Cμ*.
- *Miller effect: f*or a signal source with impedance *ZS*, the open circuit source voltage v_S is reduced to v_S*Z*in/(*Z*in + *Z*S).
- In narrow band amplifier an inductance cancels out the Miller capacitance at the required operating frequency.
- There are, however, amplifier configurations that can avoid the Miller effect without such a device and hence provide broadband amplification.

Example

- Design an npn BJT common-emitter amplifier that has a peak voltage gain of −100 for a 1 kohm collector resistor (assume β = 200, $C\mu$ = 5 pF, $C\pi$ = 50 pF and a 9V supply).
- Calculate the bandwidth of the amplifier for a 2 kohm signal source.



- The voltage gain of a BJT is given by $Av = -gmRL/(1 + j\omega C\mu RL)$ which has a peak value of $A = -gmRL$. Since $gm \approx IC/0.026$ we need a quiescent current of $IC = 0.026 \times 100/RL = 2.6$mA for a peak gain of −100.
- Noting that the voltage drop across the load resistance will be 2.6V, the one-third rule will be approximately satisfied if we choose $RL = RE = 1$ kohm ($IC \approx IE$).

Example

- The base resistors (*R*1 and *R*2) will need to be chosen such that (voltage divider)

$$\frac{R_1}{R_1 + R_2} = \frac{V_{BE} + V_E}{V_{CC}} = \frac{0.7 + 2.6}{9.0} = 0.367$$

- from which *R*2 = 0*.*58*R*1.
- Introducing the static rπ=re(β + 1) with re ≈ VT/IC, the d.c. BJT input resistance of is (*β* + 1)(*R*E + *r*e) which is approximately 200 kohm and hence negligible.
- R1,R2 carry one-tenth of the current that *R*E carries, thus we need *V*CC*/*(*R*1 + *R*2) = *V*E*/*10*R*E and from which *R*1 + *R*2 = 34*.*6 kohm. Consequently, *R*1 = 21*.*9 kohm and *R*2 = 12*.*7 kohm.
- The frequency ω_B = 1*/C*μ*R*L = 2×10^8 rad/s (about 30 MHz) defines the limit of operation with a perfect voltage source (zero internal impedance).
- The source will have a non-zero impedance *R*S and this will reduce the voltage at the amplifier input: *v*i = *Z*in*v*S*/*(*Z*in + *R*S), we obtain (using for Zin the eq. found above, and *R*B = *R*1||*R*2||*r*π)

$$v_i = \frac{R_B}{R_S + R_B} \frac{v_S}{1 + j\omega(R_S \parallel R_B)[C_\pi + (1 - A)C_\mu]}$$

Example

- Since $C\pi + (1 - A)C\mu = 555$ pF, the Miller capacitance will dominate unless the source impedance is very small.
- Consequently, we can ignore the frequency dependence of Av and approximate the output voltage by

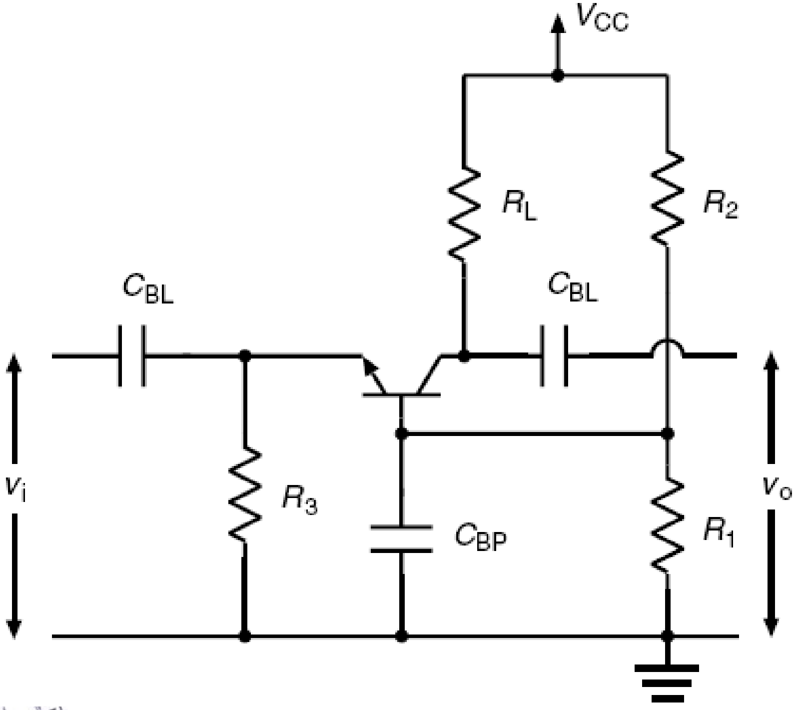
$$v_o \approx \frac{R_B}{R_S + R_B} \frac{-g_m R_L v_S}{1 + j\omega(R_S \parallel R_B)[C_\pi + (1 - A)C_\mu]}$$

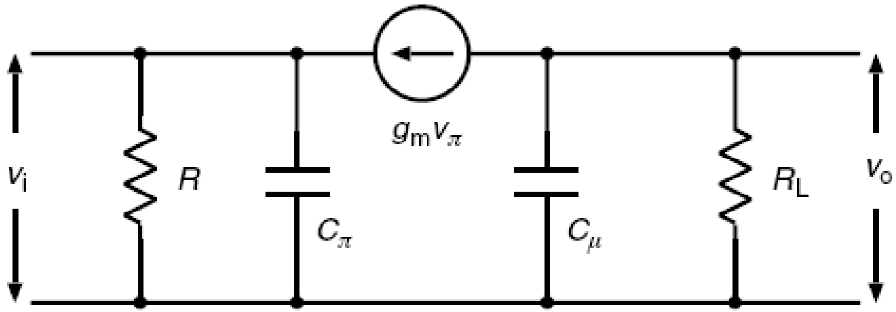
- For a transfer function of the form $T(s) = \dfrac{T_0}{1 + s/\omega_B}$
- $\omega_B = (RS\ \ RB)^{-1}\ [C\pi + (1 - A)C\mu]^{-1}$

A common-base amplifier

- An alternative amplifier configuration is the common-base amplifier
- the common-base amplifier: with $R = R3||r\pi$, the voltage gain is:

$$A_o = \frac{v_o}{v_i} = g_m \left(R_L \parallel \frac{1}{j\omega C_\mu} \right)$$





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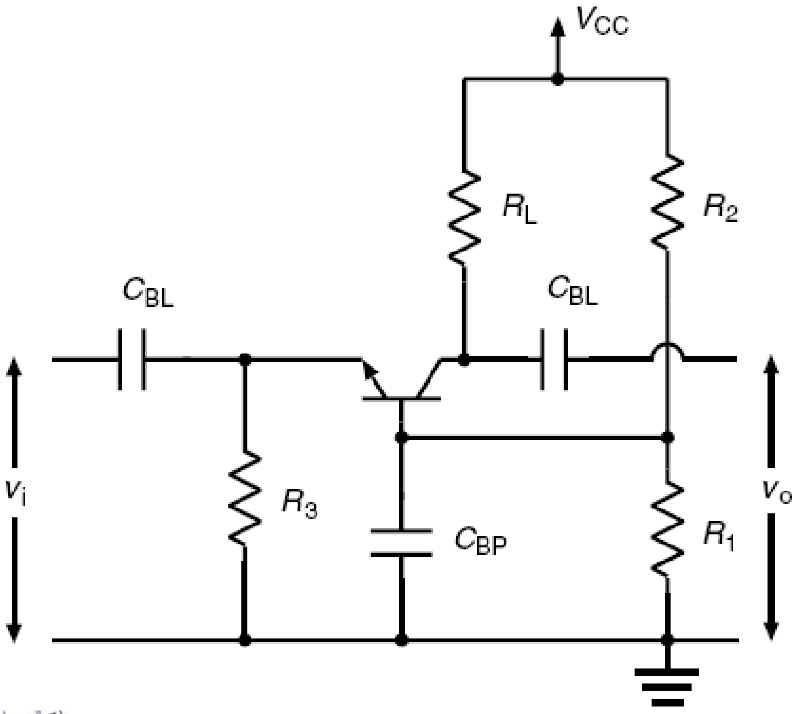
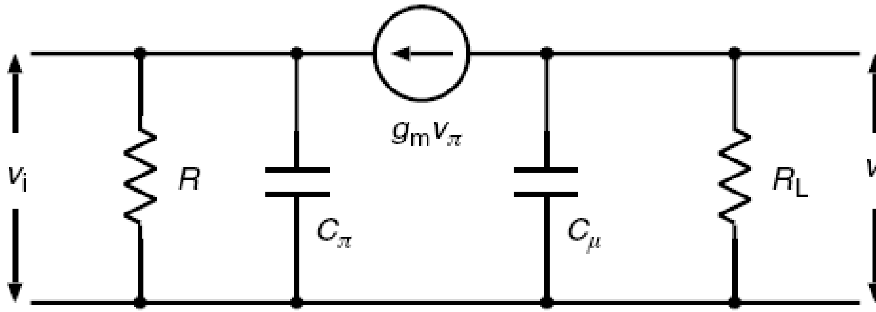




A common-base amplifier

- the output impedance by:

$$Z_{\text{out}} = R_{\text{L}} \parallel \frac{1}{j\omega C_{\mu}}$$

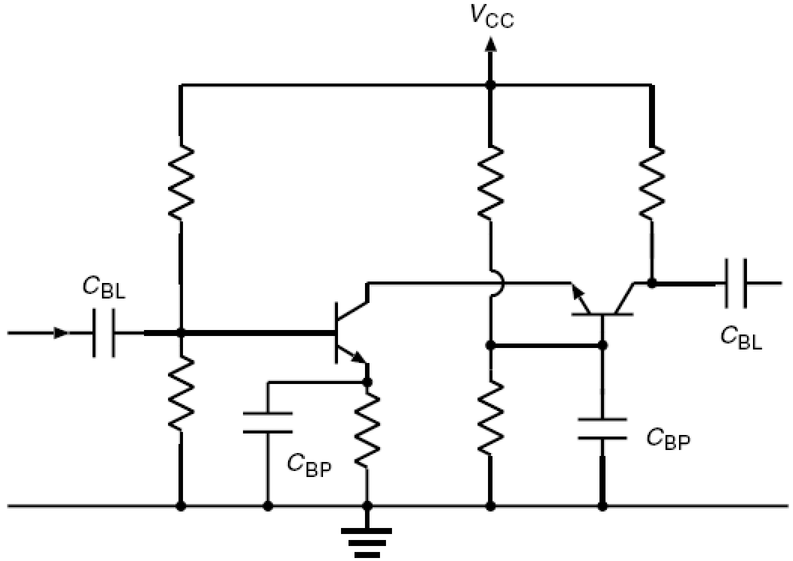
- noting that *vi* = −*vπ* , the input impedance will be given by:

$$Z_{\text{in}} = \frac{\text{input voltage}}{\text{input current}} = \frac{v_{\text{i}}}{v_{\text{i}} Z^{-1} - g_{\text{m}} v_{\pi}} = \frac{Z}{1 + g_{\text{m}} Z} \qquad Z = R \parallel \frac{1}{j\omega C_{\pi}} = \frac{R}{1 + j\omega R C_{\mu}}$$

- because there is no Miller effect, the CB exhibits a much improved frequency response compared to the common-emitter amplifier.
- However, there will be a much lower input impedance.
- In general, a common-base amplifier
- will have high voltage gain, low input impedance, high output impedance and good frequency response.

cascode configuration

- Common-emitter and common-base amplifiers can be combined to produce the cascode configuration.
- At its input this circuit has a common-emitter amplifier and hence a relatively high input impedance. This input stage, however, has a low voltage gain (the low input impedance of the common-base amplifier reduces the gain) to counteract the Miller effect.
- The common-base amplifier at the output of the circuit provides considerable voltage gain.
- As a consequence, the cascode amplifier has relatively high input and output impedances, high voltage gain and good frequency response.

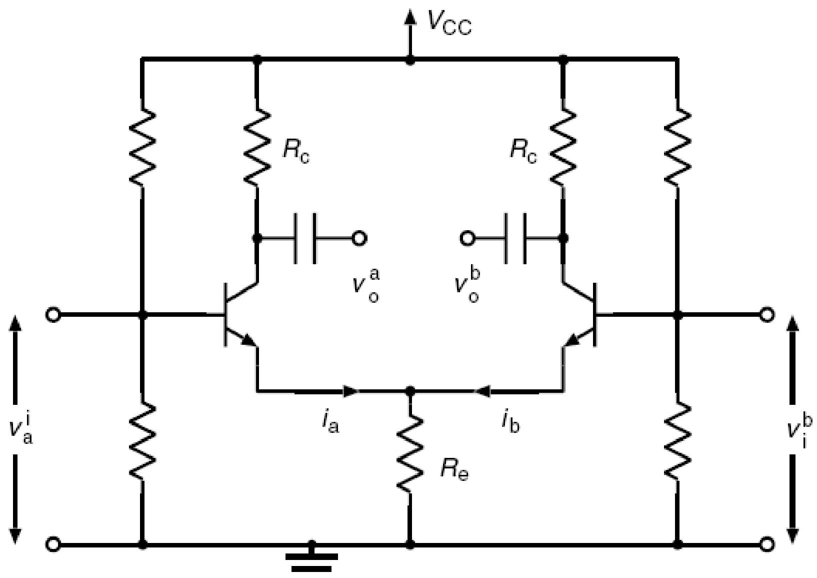


Differential amplifiers

- Differential amplifiers have the ability to amplify voltage differences and are the basis of many integrated circuit designs. The gain of these amplifiers can be electronically controlled and, because of this, they can also be configured as mixers.
- If we assume the simple small signal BJT model of Figure 3.9, we obtain the following expressions for the emitter currents

$$i_{\mathrm{a}} = \left[v_{\mathrm{i}}^{a} - (i_{\mathrm{a}} + i_{\mathrm{b}})R_{\mathrm{e}}\right]g_{\mathrm{m}}$$

$$i_{\mathrm{b}} = \left[v_{\mathrm{i}}^{b} - (i_{\mathrm{a}} + i_{\mathrm{b}})R_{\mathrm{e}}\right]g_{\mathrm{m}}$$



Differential amplifiers

- which imply that

$$i_a - i_b = g_m\left(v_i^a - v_i^b\right) \qquad i_a + i_b = g_m \frac{v_i^a + v_i^b}{1 + 2g_m R_e}$$

- assuming a high current gain, we obtain the differential output voltage

$$v_d = v_o^a - v_o^b \approx -g_m R_c\left(v_i^a - v_i^b\right)$$

- and the common output voltage

$$v_c = \frac{v_o^a + v_o^b}{2} \approx -\frac{g_m R_c}{1 + 2g_m R_e}\left(\frac{v_i^a + v_i^b}{2}\right)$$

$$\approx -\frac{R_c}{2R_e}\left(\frac{v_i^a + v_i^b}{2}\right).$$

Differential amplifiers

- which imply that

$$i_a - i_b = g_m\left(v_i^a - v_i^b\right) \qquad i_a + i_b = g_m \frac{v_i^a + v_i^b}{1 + 2g_m R_e}$$

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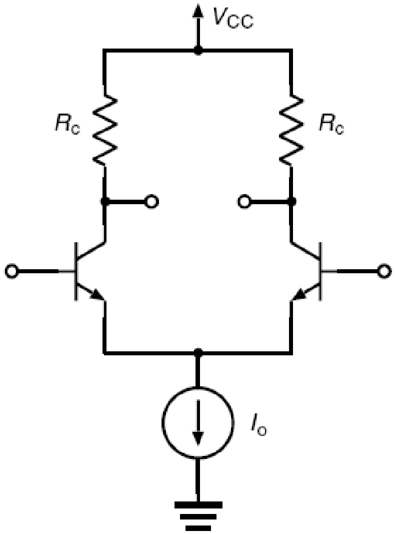
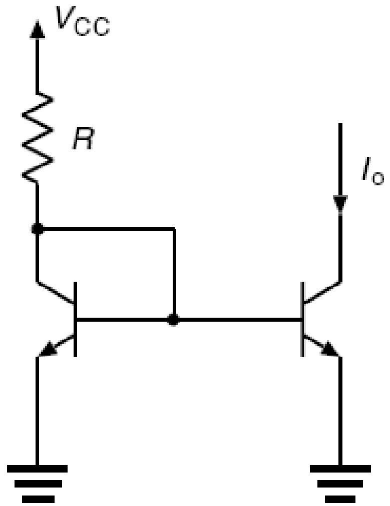
$$\approx -\frac{R_c}{2R_e}\left(\frac{v_i^a + v_i^b}{2}\right).$$

- For *Rc/Re* 1, we have a differential output alone

Differential amplifiers

- A practical current source, the current in the left transistor is set by its base–emitter voltage ($V_{BE} = V_{CC} - I_o R$).
- Given the sensitivity of I_o to changes in V_{BE}, the current will be very stable when the collector resistor R is large. Since both transistors have identical base–emitter voltages, the current in the right–hand transistor will be the same as that in left-hand transistor. As a consequence, the configuration is known as a *current mirror*.





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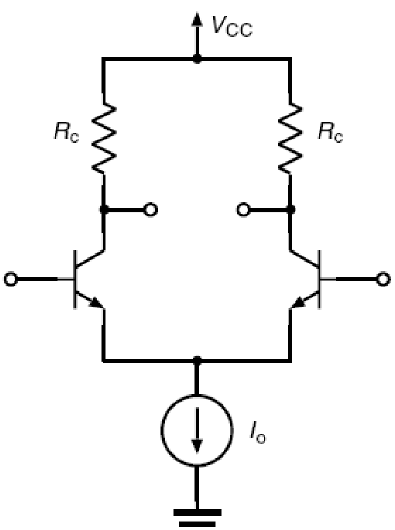
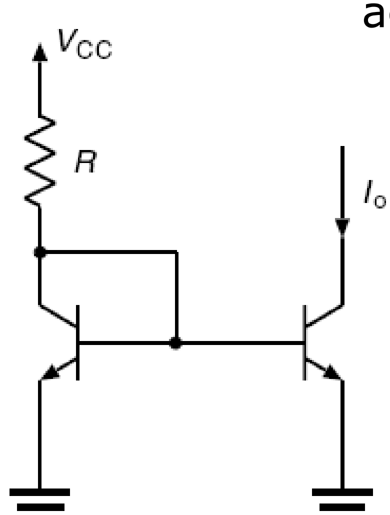
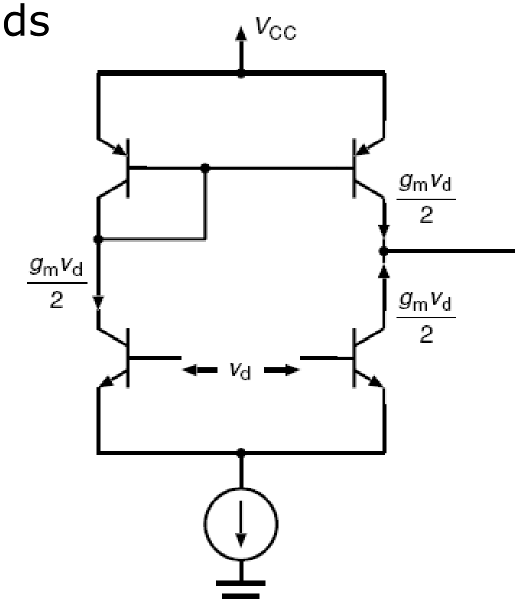


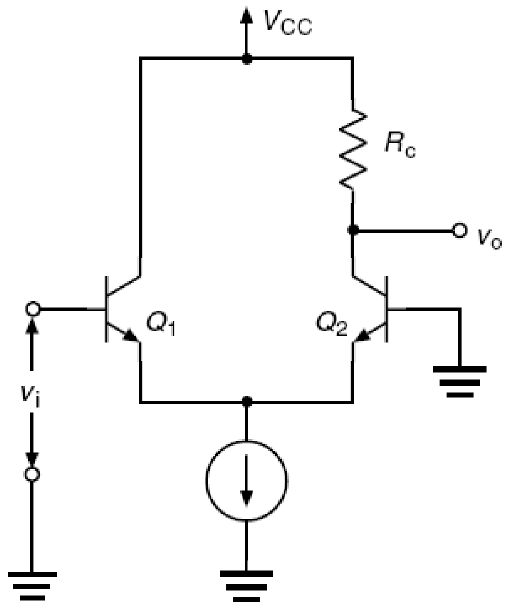


Differential amplifier with active loads



Differential amplifiers

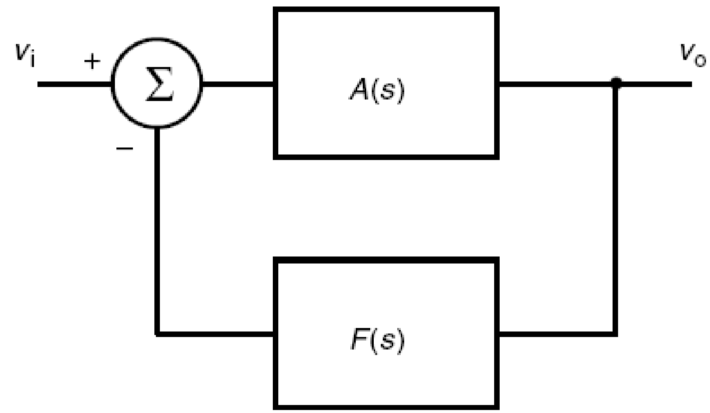
- a differential amplifier can be regarded as a pair of back-to-back common-emitter amplifiers with inputs $vd/2$ and $-vd/2$, respectively. As a consequence, it will suffer from the Miller effect.
- It can, however, be used to produce a single-ended amplifier that is immune from the Miller effect.
- The RF grounding of the $Q1$ collector eliminates the Miller effect on transistor $Q1$ and the RF grounding of the base of $Q2$ eliminates the Miller effect on $Q2$.



Feedback

- Feedback is very useful for RF circuits; negative feedback loop is shown. The amplifier has a transfer function *A*(*s*) and the feedback loop has a transfer function *F*(*s*) (note that the feedback is subtracted from the input).

$$\frac{v_o}{v_i} = A_F(s) = \frac{A(s)}{1 + A(s)F(s)}$$

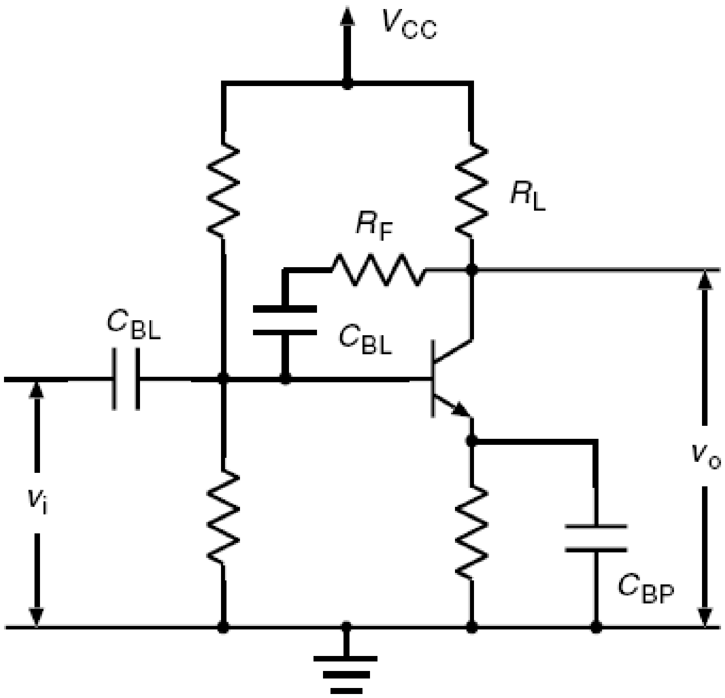


- feedback can be used to manipulate the transfer characteristics of an amplifier.
- In particular, feedback can be used to reduce the effect of the distortion caused by amplifier non-linearity. In the limit |*AF*|>>1,
- that is, the behaviour of the system is primarily dictated by its feedback characteristics.

$$\frac{v_o}{v_i} \approx \frac{1}{F(s)}$$

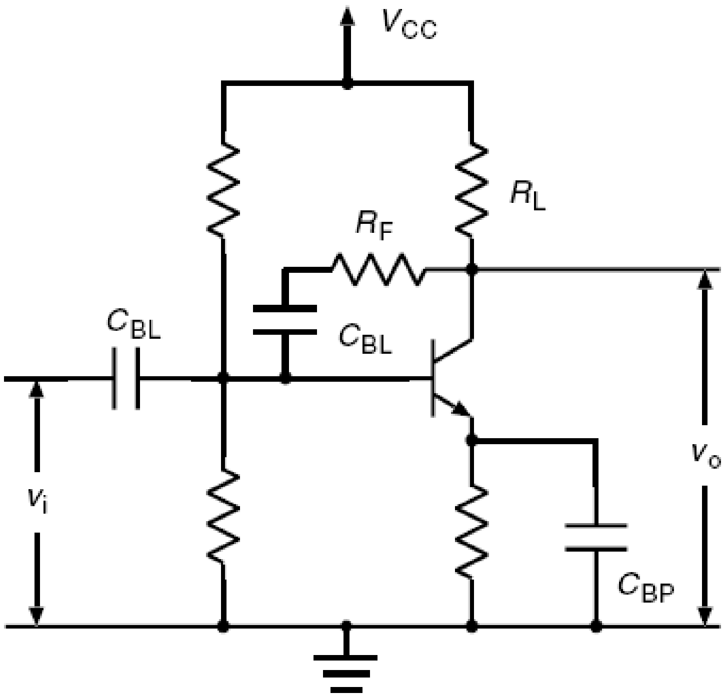
Feedback

- assume a high gain BJT and *R*F>>*R*S where *R*S is the combined impedance of the signal source, the bias network and the transistor.
- The feedback loop will have a gain of ~−*R*S/*R*F and so the amplifier will have a voltage gain of approximately −*R*F/*R*S.



Feedback

- assume a high gain BJT and *RF*>>*RS* where *RS* is the combined impedance of the signal source, the bias network and the transistor.
- The feedback loop will have a gain of ~−*RS*/*RF* and so the amplifier will have a voltage gain of approximately −*RF*/*RS*.



Feedback

- Another use of feedback is to increase the bandwidth of an amplifier. Let the dominant frequency dependence of an amplifier be described by

$$A(s) = \frac{A_0}{1 + s/\omega_B}$$

- where ω_B sets the bandwidth of the amplifier. Consider a feedback loop with constant gain $F(s) = A$f, then

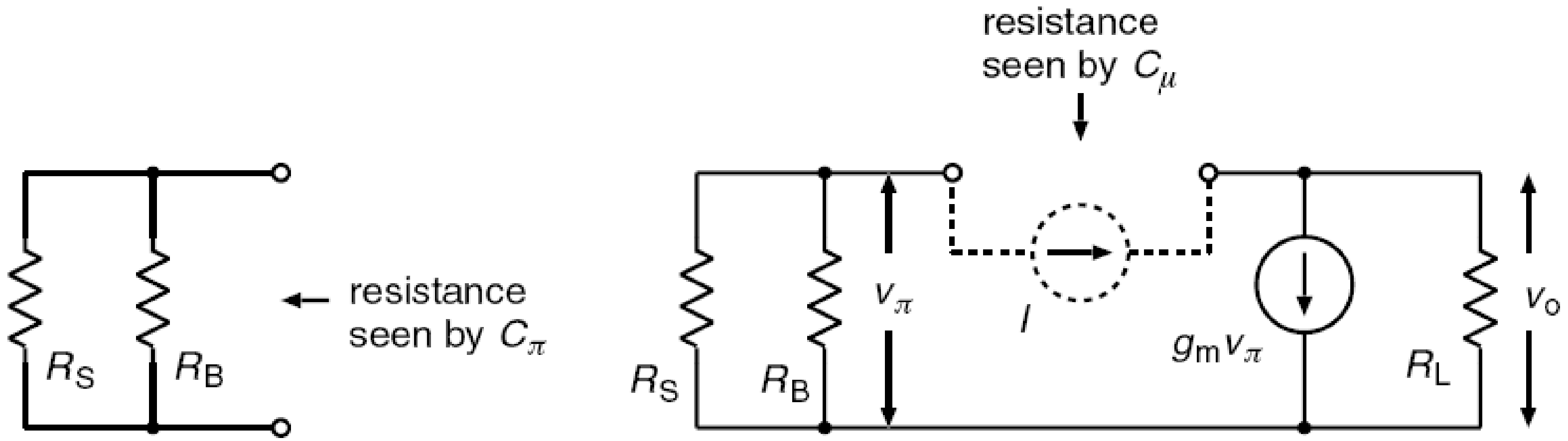
$$A_F(s) = \frac{A(s)}{1 + A(s)A_f} = \left(\frac{A_0}{1 + A_0 A_f}\right)\left[\frac{1}{1 + \frac{s}{\omega_B(1+A_0 A_f)}}\right]$$

- Note that the bandwidth has increased from ω_B to $\omega_F = \omega_B(1 + A_0 A_f)$, but that the d.c. gain has decreased from $A0$ to AF(0) = $A0$/(1 + $A0A$f).
- This is because, the gain–bandwidth product remains constant (i.e., $A0\omega$B = AF(0)ωF).

Feedback

- Since bandwidth is such an important consideration in RF amplifiers, it is useful to have a means of estimating its value.
- For a simple RC circuit, the frequency response will be dictated by the time constant, this leads to an estimate of $\omega B = 1/RC$ for the bandwidth. For a more complex system, a general estimate of the bandwidth is

$$\omega_{\mathrm{B}} = 1/(\sum_{i=1}^{N} R_{\mathrm{i}} C_{\mathrm{i}})$$



Example

- For the common-emitter amplifier, use the approximate technique to estimate the bandwidth when the source has a 2 k impedance.

- $C\pi$ will see the resistance of $R\pi = R\text{S}||R\text{B}$.
- The resistance $R\mu$ seen by $C\mu$, can be found by forcing a current I to flow through the circuit.
  - The voltage $v\pi$ across the input resistors will be $v\pi = -I\,(R\text{S}||R\text{B})$ and, since a current $I - gmv\pi$ will flow through $R\text{L}$, the voltage $v\text{o}$ across this resistor will be given by

    $v\text{o} = (I - g_\text{m}v_\pi)R\text{L} = I\,R\text{L}[1 + g_\text{m}(R\text{S}||R\text{B})]$.
- Consequently, since the voltage across the source of current I will be $v\text{o} - v\pi$, the resistance $R\mu$ will be given by $R\mu = (v\text{o} - v\pi)/I = R\text{L} + (1 + g_\text{m}R\text{L})(R\text{S}||R\text{B})$. Using values from the
- The bandwidth can be approximated by $\omega_\text{B} = 1/(R\mu C\mu + R\pi C\pi)$
- The exact solution is $\omega_\text{B} = (R\text{S}||R\text{B})^{-1}[C\pi + (1 - A)C\mu]^{-1}$

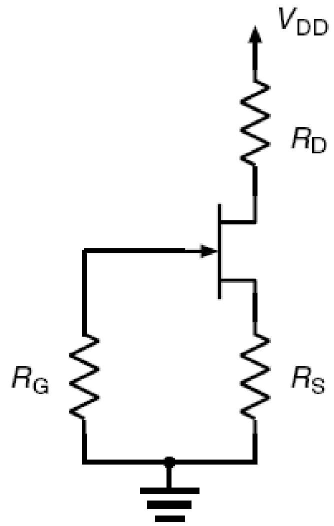
FET amplifiers

- A field-effect transistor (FET) is a semiconductor device in which the current flow (between source and drain terminals) is controlled by the voltage on a high input impedance terminal (the gate).
- Characteristic FET behaviour is shown in Figure 3.34. In the triode region, the current in the drain–source channel can be approximated by

$$I_D = K\left[2(V_{GS} - V_t)V_{DS} - V_{DS}^2\right]$$

- and in the saturation region by

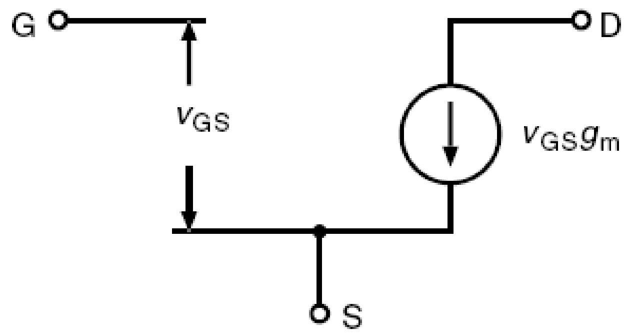
$$I_D = K(V_{GS} - V_t)^2$$



In the case of an n-channel JFET, there is the option of self-bias. The gate is set at ground voltage through a high value resistance *RG* (significantly less than the FET input resistance). *RS* is chosen to give a quiescent current around which there is sufficient linearity for the expected input swing.

FET amplifiers

- Low freq. model



- Once the quiescent conditions have been set, the transconductance *g*m is derived from:

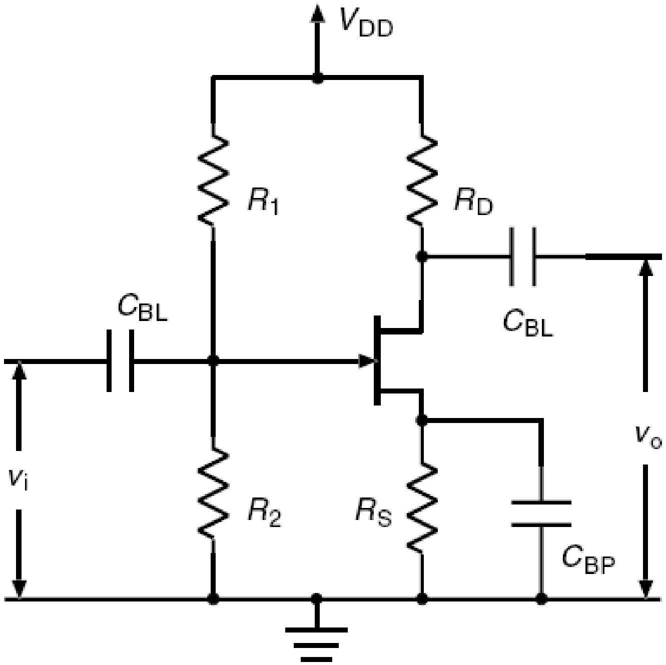
$$g_m = \frac{\partial I_D}{\partial V_{GS}}$$

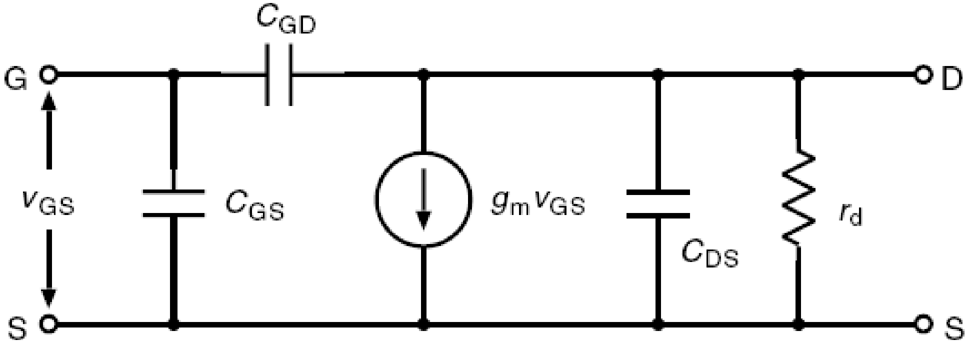
- with the derivative evaluated at the quiescent operating point. In terms of quiescent current

$$I_{DQ},\ g_m = 2\sqrt{K I_{DQ}}.$$

FET amplifiers

- The common-source amplifier is shown in Figure along with the small signal FET model, suitable for RF frequencies.
- Without the effect of capacitances, the gain will be $A0 = -gm(RD||rd)$. Consequently, assuming $|A0|>>1$, we can use the Miller theorem to transform CGD into equivalent input and output capacitances.





FET amplifiers

- The small signal gain is:

$$\frac{v_o}{v_i} = \frac{-g_m(R_D \parallel r_d)}{1 + j\omega(R_D \parallel r_d)(C_{GD} + C_{DS})}$$

- the input, output impedances:

$$Z_{in} = \frac{1}{j\omega[C_{GS} + (1 - A_0)C_{GD}]} \qquad Z_{out} = R_D \parallel r_d \parallel \frac{1}{j\omega(C_{GD} + C_{DS})}.$$

- the Miller effect is a significant factor in FET amplifiers at high frequencies.
- There are, however, additional effects are the gate-to-drain and gate-to-source capacitances that include a small amount of series resistance. This will result in an input resistance that falls in value as frequency rises (typically to a value of a few kilohms at 100 MHz), an effect that must be taken into account when matching FETs at very high frequencies.

FET amplifiers Example

- Design an JFET common-source amplifier that has a peak RF voltage gain of−10 and harmonic components 26 dB below the fundamental when there is a 100mV
- amplitude RF drive. For the chosen transistor, and a 1 k signal source, calculate the bandwidth of the amplifier.

- We will design the amplifier around the 2N3819 JFET for which typical parameters are $K = 1.3 \times 10^{-3}$ A/V^2, Vt = −3 V, CGS = 4 pF, CGD = 1.6 pF, CDS ≈ 0 and rd =30 k.
- The circuit is based on the self-bias topology, we can remove resistor R1, while for R2 100 kohm.
- The output voltage at the drain will have the form

$$v_o = V_{DD} - R_D K(V_{bias} + v_{RF} - V_t)$$

- where Vbias is the gate–source voltage under quiescent conditions and vRF is the RF input voltage.
- For vRF = VRF cos ωt,

$$v_o = V_{DD} - R_D K(V_{bias} - V_t)^2 - 2R_D K(V_{bias} - V_t)V_{RF}\cos(\omega t) - R_D K V^2\mathrm{RF}\cos^2(\omega t)$$

FET amplifiers Example

- and so the output voltage *vo* will have amplitude $2R_DK(V_{bias} - V_t)V_{RF}$ at the fundamental frequency ω and $1/2R_DKV_{RF}^2$ at the harmonic frequency 2ω.
- Consequently, we require

$$\frac{V_{RF}}{4(V_{bias} - V_t)} < \frac{1}{20}$$

- for the second harmonic to be at the requisite level (26 dB implies that the harmonic voltage be 1/20 of the fundamental voltage). This will imply that $V_{bias} > -2.5$V for input signal at the level $V_{RF} = 100$ mV. Assuming the transistor to be saturated, we obtain from the characteristic equation that:

$$-V_{bias} = R_S K (V_{bias} - V_t)^2$$

- We choose a bias voltage $V_{bias} = -1.5$V and from Equation 3.42 obtain that R_S =
- 513ohm.

FET amplifiers Example

- On noting that $gm = 2K(V\text{bias} - V\text{t})$, the amplifier will have a voltage gain given by $-2R_D K(V\text{bias} - V\text{t})$ and to attain a value of −10 will require RD to have a value of 2.6 kohm.
- At high freq, the dominant effect will come from the Miller capacitance at the input. This will combine with the finite impedance R of the signal source to
- reduce the amplifier input voltage.
- The voltage will be given by $vi = Zin\,vS/(Zin + R)$, where vS is the open circuit voltage of the source and Zin is the input impedance of the amplifier.

$$v_i = \frac{v_S}{1 + j\omega R[C_{GS} + (1 - A_0)C_{GD}]}, \qquad v_o = \frac{A v_S}{1 + j\omega R[C_{GS} + (1 - A_0)C_{GD}]}$$

- for a 1 kohm signal source impedance, this will imply a bandwidth of
- $\omega B = R^{-1}\,(CGS + (1 - A0)CGD)^{-1}$