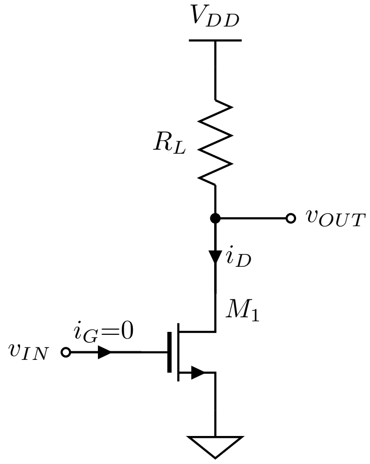


Figure 3.15: The common-source amplifier.

3.3 The Common-Source Amplifier

Similar to the common-emitter amplifier, the MOSFET common-source amplifier, shown in Fig. 3.15, utilizes the quadratic relationship between the drain current and gate-to-source voltage to provide voltage gain.

Given the input quiescent DC voltage, and recognizing that this is equal to the quiescent gate-to-source voltage, and assuming that the MOSFET is in the saturation region, we can determine the quiescent drain current, $I_{D,Q}$,

$$I_{D,Q} = k \cdot (V_{IN} - V_{TH})^2 \cdot (1 + \lambda V_{OUT}) \tag{3.57}$$

The channel length modulation parameter, λ, is usually in the order of $0.01\,\mathrm{V}^{-1}$, and for supply voltages of around $5\,\mathrm{V}$,

$$\lambda V_{OUT} \ll 1 \tag{3.58}$$

Thus Eq. 3.57 can be approximated as

$$I_{D,Q} = k \cdot (V_{IN} - V_{TH})^2 \tag{3.59}$$

The KVL equation for the output loop can be written as

$$V_{DD} - I_{D,Q} R_L - V_{OUT} = 0 \tag{3.60}$$

Plugging in Eq. 3.59 into Eq. 3.60 gives us the quiescent output DC voltage

$$V_{OUT} = V_{DD} - R_L \cdot k \cdot (V_{IN} - V_{TH})^2 \tag{3.61}$$

Recall that an important fundamental difference between BJTs and MOSFETs is the boundary between the desired operating region (forward-active for BJTs and saturation for MOSFETs) and the low V_{OUT} region (saturation for BJTs and linear or ohmic for MOSFETs). For a common-emitter amplifier, the BJT goes into saturation when $V_{CE} \leq V_{CE,sat}$. However, in a common-source amplifier, the MOSFET goes into the linear region when

$$V_{DS} \leq V_{GS} - V_{TH} \tag{3.62}$$

Again, note that this boundary between the MOSFET linear and saturation regions is dependent on the quiescent DC operating point, unlike $V_{CE,sat}$ which can be treated as a constant. In the linear or ohmic region, the MOSFET behaves like a nonlinear resistor, whose current is given by

$$I_{D,Q} = 2k \cdot \left((V_{IN} - V_{TH})\, V_{OUT} - \frac{V_{OUT}^2}{2} \right) \tag{3.63}$$

In the linear region, the relationship between $I_{D,Q}$ and V_{GS} is no longer quadratic, but is now linear, hence the name of the operating region. Thus, the transconductance, g_m, in the linear region is smaller, leading to lower voltage gain. Therefore, to prevent the MOSFET from going into the linear region, the minimum output voltage of the common-source amplifier is

$$V_{OUT,\min} = V_{IN} - V_{TH} \tag{3.64}$$

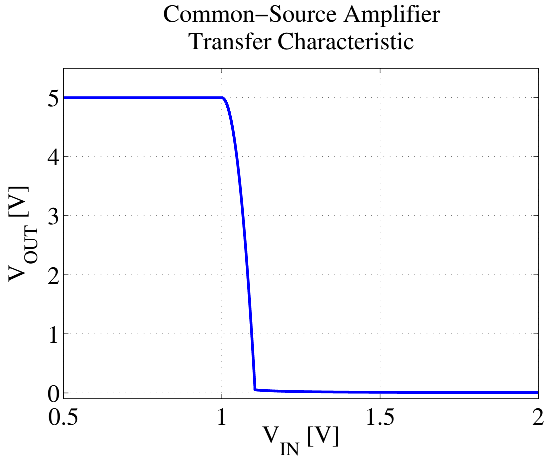


Figure 3.16: The common-source amplifier large signal transfer characteristic for $k = 90\,\frac{\text{mA}}{\text{V}^2}$, $V_{TH} = 1\,\text{V}$, $\lambda \to 0$, $V_{DD} = 5\,\text{V}$, and $R_L = 5\,\text{k}\Omega$.

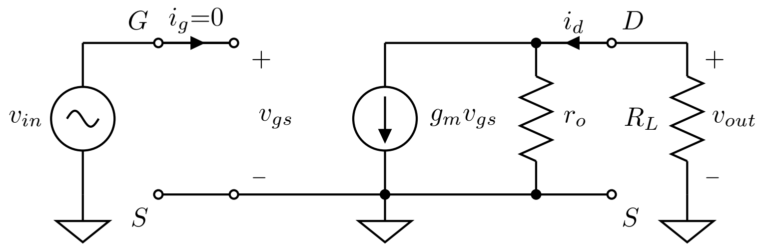


Figure 3.17: The small signal equivalent circuit of the common-source amplifier in Fig. 3.15.

Combining Eq. 3.63 into Eq. 3.60, and assuming that the output voltage is small such that the $\frac{V_{OUT}^2}{2}$ term in Eq. 3.63 is negligible, we can express the output quiescent DC voltage when the transistor is in the linear region as

$$V_{OUT} \approx \frac{V_{DD}}{1 + 2k \cdot (V_{IN} - V_{TH}) \cdot R_L} \tag{3.65}$$

Also, we can use the relationship $V_{OUT} \approx V_{DD}$ when $V_{IN} < V_{TH}$. This is due to our assumption that when the gate-to-source voltage is less than the threshold voltage, $I_{D,Q} \approx 0$. We can then use Eqs. 3.61 and 3.65 to plot the large signal common-source amplifier transfer characteristic, as shown in Fig. 3.16.

Once we have determined the quiescent DC drain current, we can now determine the small signal parameters of the MOSFET as

$$g_m = \left.\frac{\partial I_D}{\partial V_{GS}}\right|_{I_D = I_{D,Q}} = 2k \cdot (V_{IN} - V_{TH}) = \sqrt{4k \cdot I_{D,Q}} = \frac{2 \cdot I_{D,Q}}{V_{IN} - V_{TH}} \tag{3.66}$$

$$r_o = \left.\frac{\partial V_{DS}}{\partial I_D}\right|_{I_D = I_{D,Q}} = \frac{1}{\lambda I_{D,Q}} \tag{3.67}$$

$$r_\pi = \left.\frac{\partial V_{GS}}{\partial I_G}\right|_{I_D = I_{D,Q}} \to \infty \tag{3.68}$$

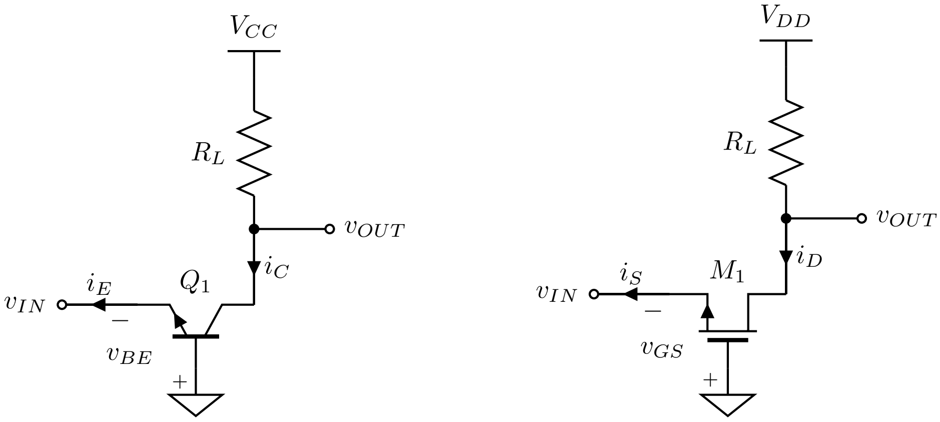
The small signal equivalent circuit of the common-source amplifier is shown in Fig. 3.17. It is interesting to note that the transconductance of a BJT is proportional to $I_{C,Q}$, while that of a MOSFET is proportional to $\sqrt{I_{D,Q}}$. Thus, for the same bias current, we can obtain a higher transconductances in BJTs.

After deriving the small signal transistor parameters, we can easily determine the hybrid-π two-port parameters. By inspection, we can see that

$$G_m = g_m = \sqrt{4k \cdot I_{D,Q}} \tag{3.69}$$

$$R_o = r_o \parallel R_L = \frac{1}{\lambda I_{D,Q}} \parallel R_L \tag{3.70}$$

$$R_i \to \infty \tag{3.71}$$



(a) The common-base amplifier (b) The common-gate amplifier

Figure 3.18

The voltage gain is then

$$A_v = -G_m R_o = -g_m (r_o \parallel R_L) \tag{3.72}$$

If $r_o \gg R_L$, the voltage gain can be approximated as

$$A_v \approx -g_m R_L = -2k \cdot R_L \cdot (V_{IN} - V_{TH}) = -R_L \sqrt{4k \cdot I_{D,Q}} \tag{3.73}$$

The intrinsic voltage gain, a_o, of a MOSFET is given as

$$a_o = -g_m r_o = -\frac{\sqrt{4k \cdot I_{D,Q}}}{\lambda I_{D,Q}} = -\frac{1}{\lambda}\sqrt{\frac{4k}{I_{D,Q}}} = -\frac{2}{\lambda (V_{IN} - V_{TH})} \tag{3.74}$$

Note that unlike the BJT, the MOSFET intrinsic gain is dependent on its quiescent DC operating point.

Though there are fundamental differences between BJTs and MOSFETs, as soon as we obtain (or select) the quiescent DC collector or drain currents, the small signal characteristics turn out to be the same. In the case of the common-emitter and common-source amplifiers, their small signal behavior is the described by the same equivalent circuit, and only the actual parameter values (g_m, r_o, and r_π) are dependent on the transistor used.

3.4 The Common-Base and Common-Gate Amplifiers

The second single-stage amplifier topology that we will look at is the BJT common-base (CB) amplifier in Fig. 3.18a, and its MOSFET equivalent, the common-gate (CG) amplifier in Fig. 3.18b.

DC Analysis: In order to get the quiescent DC operating point of the common-base amplifier, we write out the KVL equation at the input loop, and recognizing that $v_{BE} = -v_{IN}$ and $v_{CE} = v_{OUT} - v_{IN}$, we get

$$I_{C,Q} = I_S \cdot e^{\frac{-V_{IN}}{V_T}} \cdot \left(1 + \frac{V_{CE,Q}}{V_A}\right) = I_S \cdot e^{\frac{-V_{IN}}{V_T}} \cdot \left(1 + \frac{V_{OUT} - V_{IN}}{V_A}\right) \tag{3.75}$$

If we assume that we are given the input DC voltage, V_{IN}, and that $V_A \gg V_{OUT} - V_{IN}$, we can simplify Eq. 3.75 into

$$I_{C,Q} \approx I_S \cdot e^{\frac{-V_{IN}}{V_T}} \tag{3.76}$$

Using the KVL equation at the output loop

$$V_{CC} - I_{C,Q} R_L - V_{OUT} = 0 \tag{3.77}$$

We can get quiescent DC output voltage as

$$V_{OUT} = V_{CC} - R_L \cdot I_S \cdot e^{\frac{-V_{IN}}{V_T}} \tag{3.78}$$

Note that Eq. 3.78 describes the large signal relationship between the input voltage and the output voltage, and thus, is the transfer characteristic of the common-base amplifier. We can prevent the transistor from moving into its saturation region, by making sure that

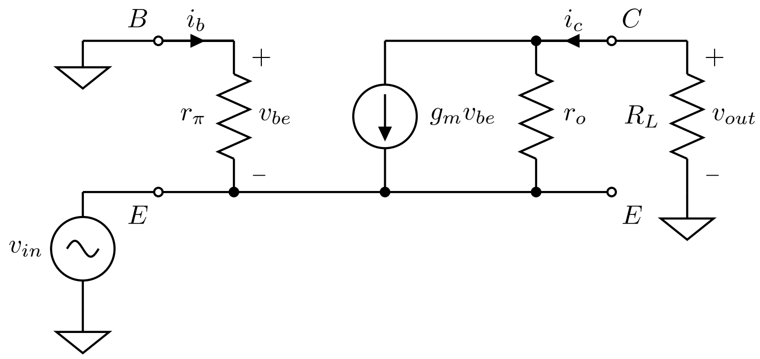


Figure 3.19: The small signal equivalent circuit of the common-base amplifier.

$$V_{CE} = V_{OUT} - V_{IN} > V_{CE,sat} \tag{3.79}$$

For the common-gate amplifier, with $v_{GS} = -v_{IN}$, and $V_{GS} = -V_{IN} > V_{TH}$, we can use the same procedure to get the quiescent DC collector current as

$$I_{D,Q} = k \cdot (-V_{IN} - V_{TH})^2 \cdot (1 + \lambda (V_{OUT} - V_{IN})) \tag{3.80}$$

and if we assume that $\lambda (V_{OUT} - V_{IN}) \ll 1$, we get

$$I_{D,Q} = k \cdot (-V_{IN} - V_{TH})^2 \tag{3.81}$$

Thus, the quiescent DC output voltage can be written as

$$V_{OUT} = V_{DD} - R_L \cdot k \cdot (-V_{IN} - V_{TH})^2 \tag{3.82}$$

Again, Eq. 3.82 is the transfer characteristic of the common-gate amplifier. To prevent the transistor from going into its linear region, we need to ensure that

$$V_{DS} = V_{OUT} - V_{IN} > -V_{IN} - V_{TH} \tag{3.83}$$

Simplifying, we get

$$V_{OUT} > -V_{TH} \tag{3.84}$$

Small Signal Behavior: As soon as we have the quiescent DC collector and drain currents, we can calculate the small signal parameters of the BJT and MOSFET. In general, Fig. 3.19 shows the small signal equivalent circuit of the common-base amplifier. Note that this is also the equivalent circuit of the common-gate amplifier. The only difference is that for the common-gate amplifier, $r_\pi \to \infty$.

To determine the effective transconductance, G_m, we short the output node of the circuit in Fig. 3.19 to ground, and calculate the output short-circuit current. Writing the KCL at the output node, and noting that $i_c = i_{out}$, $v_{be} = -v_{in}$, and $v_e = v_{in}$, we get

$$i_{out} = g_m v_{be} - \frac{v_e}{r_o} = -g_m v_{in} - \frac{v_{in}}{r_o} \tag{3.85}$$

Thus, if we assume that for the BJT,

$$|a_o| = |g_m r_o| \gg 1 \tag{3.86}$$

the effective transconductance becomes

$$G_m = \frac{i_{out}}{v_{in}} = -g_m - \frac{1}{r_o} \approx -g_m \tag{3.87}$$

This effective transconductance is slightly larger than the transconductance of the common-emitter amplifier because of the added feed-forward path from the input to the output, via r_o, leading to a larger output short-circuit current.

With the output shorted to small signal ground, we can also calculate the effective input resistance. The total current can be written as

$$i_{in} = \frac{v_{in}}{r_\pi} - i_{out} = \frac{v_{in}}{r_\pi} + g_m v_{in} + \frac{v_{in}}{r_o} \tag{3.88}$$

Table 3.2: Common-base and common-gate amplifier characteristics.

	Common-Base Amplifier	Common-Gate Amplifier
g_m	$\frac{I_{C,Q}}{V_T}$	$\sqrt{4k \cdot I_{D,Q}}$
r_o	$\frac{V_A}{I_{C,Q}}$	$\frac{1}{\lambda \cdot I_{D,Q}}$
r_π	$\frac{\beta}{g_m} = \frac{\beta \cdot V_T}{I_{C,Q}}$	∞
G_m	$-g_m = -\frac{I_{C,Q}}{V_T}$	$-g_m = -\sqrt{4k \cdot I_{D,Q}}$
R_i	$\frac{1}{g_m} = \frac{V_T}{I_{C,Q}}$	$\frac{1}{g_m} = \frac{1}{\sqrt{4k \cdot I_{D,Q}}}$
R_o	$r_o \parallel R_L = \frac{V_A}{I_{C,Q}} \parallel R_L$	$r_o \parallel R_L = \frac{1}{\lambda \cdot I_{D,Q}} \parallel R_L$
A_v	$-G_m R_o = g_m \left(r_o \parallel R_L\right) = \frac{I_{C,Q}}{V_T}\left(\frac{V_A}{I_{C,Q}} \parallel R_L\right)$	$-G_m R_o = g_m \left(r_o \parallel R_L\right) = \sqrt{4k \cdot I_{D,Q}}\left(\frac{1}{\lambda \cdot I_{D,Q}} \parallel R_L\right)$

Therefore, the small signal input resistance of the common-base amplifier is

$$R_i = \frac{v_{in}}{i_{in}} = \frac{1}{\frac{1}{r_\pi} + g_m + \frac{1}{r_o}} = r_\pi \parallel r_o \parallel \frac{1}{g_m} \tag{3.89}$$

Note that in Eq. 3.89, we treated the transistor transconductance as a resistor with value $\frac{1}{g_m}$, since the voltage that controls the dependent source is the same voltage across that dependent source. Again, using our assumption in given in Eq. 3.86, we can approximate the input resistance as

$$R_i \approx \frac{1}{g_m} \tag{3.90}$$

which is β times smaller than r_π.

The common-base output resistance can be easily found by inspection. Shorting out the input results in $v_{be} = 0$, thus zeroing out the current through the dependent source, leaving us with

$$R_o = r_o \parallel R_L \tag{3.91}$$

The common-base voltage gain is then obtained using Eqs. 3.87 and 3.91, as

$$A_v = -G_m R_o = g_m \left(r_o \parallel R_L\right) \tag{3.92}$$

It is important to note that even though the magnitude of the small signal voltage gain is the same as that of the common-emitter amplifier, the small signal voltage gain of the common-base amplifier is positive, and is thus a *non-inverting amplifier*.

The analysis is the same for a common-gate amplifier, as seen in Table 3.2. A few important things to note is that for the common-gate amplifier, even though the transistor $r_\pi \to \infty$, the input resistance of the whole amplifier is finite since the input is connected to the source and not to the gate of the MOSFET.

3.5 The Common-Collector and Common-Drain Amplifiers

The third single-stage amplifier topology is the BJT common-collector (CC) amplifier in Fig. 3.20a, and the MOSFET common-drain (CD) amplifier in Fig. 3.20b.

DC Analysis: To determine the quiescent DC operating point of the common-collector amplifier in Fig. 3.20a, we can write the KVL equation for the input loop, assuming that the transistor is in the forward-active region, as

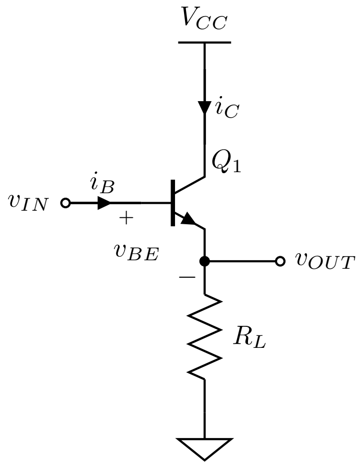
$$V_{IN} - V_{BE} - I_{E,Q} R_L = V_{IN} - V_T \ln\left(\frac{I_{C,Q}}{I_S}\right) - I_{C,Q}\left(1 + \frac{1}{\beta}\right) R_L = 0 \tag{3.93}$$

We can solve Eq. 3.93 iteratively, but for simplicity, and without losing generality, we can use the approximation that $V_{BE} = 0.7\,\mathrm{V}$. Using this approximation, and assuming $\beta \gg 1$, the quiescent DC collector current can then be expressed as

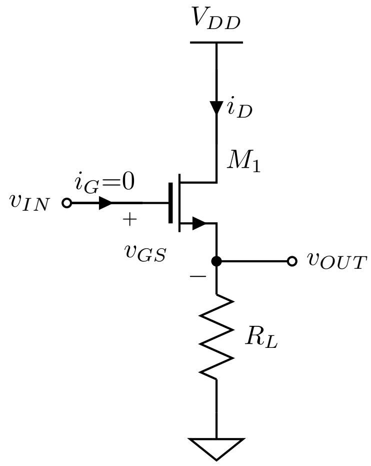
$$I_{C,Q} = \frac{V_{IN} - 0.7\,\mathrm{V}}{R_L\left(1 + \frac{1}{\beta}\right)} \approx \frac{V_{IN} - 0.7\,\mathrm{V}}{R_L} \tag{3.94}$$

Note that the quiescent output DC voltage is just

$$V_{OUT} = I_{E,Q} R_L = I_{C,Q}\left(1 + \frac{1}{\beta}\right) R_L \approx I_{C,Q} R_L \tag{3.95}$$



(a) The common-collector amplifier



(b) The common-drain amplifier

Figure 3.20

Combining Eqs. 3.93 and 3.95, we can express the transfer characteristic of the common-collector amplifier as

$$V_{OUT} = V_{IN} - V_{BE} = V_{IN} - 0.7\,\text{V} \tag{3.96}$$

From Eq. 3.96, we observe that the relationship between the input voltage and the output voltage is linear[a], or that the output voltage "follows" the input voltage. Thus, the common-collector amplifier is also known as an *emitter-follower* amplifier. Also, to keep the BJT is in its forward-active region, we need to ensure that

$$V_{CE} = V_{CC} - V_{OUT} > V_{CE,sat} \tag{3.97}$$

Repeating the same analysis for the common-drain amplifier in Fig. 3.20b, and with $I_{G,Q} = 0$, the KVL equation at the input loop, assuming that the MOSFET is in its saturation region and $\lambda \approx 0$, is

$$V_{IN} - V_{GS} - I_{S,Q}R_L = V_{IN} - \left(V_{TH} + \sqrt{\frac{I_{D,Q}}{k}}\right) - I_{D,Q}R_L = 0 \tag{3.98}$$

Since Eq. 3.98 is a quadratic equation, we can easily obtain the quiescent DC drain current. Also, recognizing that

$$V_{OUT} = I_{D,Q}R_L \tag{3.99}$$

the common-drain transfer characteristic can be written as

$$V_{OUT} = V_{IN} - V_{GS} = V_{IN} - \left(V_{TH} + \sqrt{\frac{I_{D,Q}}{k}}\right) \tag{3.100}$$

If there is little change in the quiescent drain current as we vary the DC input voltage, the relationship in Eq. 3.100 can be thought of as approximately linear, and that the output voltage again "follows" the input voltage. Thus, the common-drain amplifier is also known as a *source-follower* amplifier. In order to prevent the MOSFET from going into its linear region, we need to ensure that

$$V_{DS} = V_{DD} - V_{OUT} > V_{GS} - V_{TH} = V_{IN} - V_{OUT} - V_{TH} \tag{3.101}$$

or

$$V_{DD} > V_{IN} - V_{TH} \tag{3.102}$$

Small Signal Analysis: The small signal equivalent circuit of the common-collector amplifier is shown in Fig. 3.21.

To get the effective circuit transconductance, we connect the output to small signal ground, and calculate the short-circuit current. By inspection, if we connect the output to ground, then no current will flow through the resistors r_o and R_L, and recognizing that $v_{be} = v_{in}$ when $v_{out} = 0$, we get

[a]Remember that we have made a few approximations, so it is important to note that the relationship between V_{IN} and V_{OUT} is close to linear, but not exactly linear.

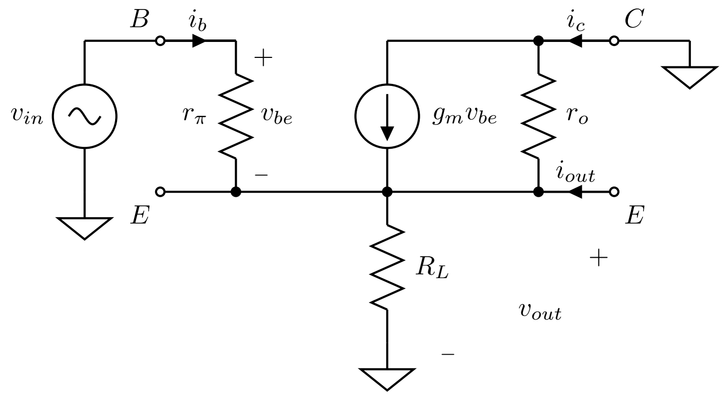


Figure 3.21: The common-collector small signal equivalent circuit.

$$i_{out} = -g_m v_{in} - \frac{v_{in}}{r_\pi} \tag{3.103}$$

The negative sign is due to the fact that the current is flowing out of the circuit, and into the short, opposite that of our convention for output current. Thus, from Eq. 3.103, and if we assume that

$$\beta = g_m r_\pi \gg 1 \tag{3.104}$$

the circuit transconductance is then

$$G_m = \frac{i_{out}}{v_{in}} = -g_m - \frac{1}{r_\pi} \approx -g_m \tag{3.105}$$

Again, by inspection, the input resistance when the output is shorted to ground, is equal to

$$R_i = r_\pi \tag{3.106}$$

To find the output resistance, the input is set to zero, and recognizing that $v_{be} = -v_{out}$, and that all the resistors are now in parallel, the output current when a test voltage source is placed at the output is

$$i_{out} = \frac{v_{out}}{R_L} + \frac{v_{out}}{r_o} + \frac{v_{out}}{r_\pi} + g_m v_{out} \tag{3.107}$$

Therefore, the output resistance is

$$R_o = \frac{v_{out}}{i_{out}} = \frac{1}{\frac{1}{R_L} + \frac{1}{r_o} + \frac{1}{r_\pi} + g_m} = R_L \parallel r_o \parallel r_\pi \parallel \frac{1}{g_m} \tag{3.108}$$

Using our assumptions in Eqs. 3.86 and 3.104, we can simplify Eq. 3.108 as

$$R_o \approx \frac{1}{g_m} \parallel R_L = \frac{R_L}{1 + g_m R_L} \tag{3.109}$$

Finally, using Eqs. 3.105 and 3.109, the small signal voltage gain of the common-collector amplifier is equal to

$$A_v = -G_m R_o = \frac{g_m R_L}{1 + g_m R_L} \tag{3.110}$$

Note that if $g_m R_L \gg 1$, then $A_v \approx 1$. This is consistent with our earlier observation that for the common-collector, the output "follows" the input.

The small signal analysis for the common-drain amplifier is exactly the same, except that for the MOSFET, $r_\pi \to \infty$.

3.6 Input Resistance Revisited

Except for the common-emitter and common-source amplifier, the input resistance is dependent on the state of the output port, and can change significantly.

Let us calculate the input resistance of the common-base and common-gate amplifiers, when the output is open-circuited. Writing the KCL equation at the output node of the circuit in Fig. 3.19, we get

$$\frac{v_{out}}{R_L} + \frac{v_{out} - v_{in}}{r_o} - g_m v_{in} = 0 \tag{3.111}$$

Table 3.3: Input resistance vs. output condition.

	$v_{out} = 0$	$i_{out} = 0$
CB/CG Amplifier	$\frac{1}{g_m}$	$\frac{1}{g_m} + \frac{R_L}{g_m r_o}$
CC/CD Amplifier	r_π	$r_\pi \left(1 + g_m R_L\right)$

Table 3.4: Summary of single-stage amplifier parameters.

	CE/CS	CB/CG	CC/CD
G_m	g_m	$-g_m$	$-g_m$
R_o	$r_o \parallel R_L$	$r_o \parallel R_L$	$\frac{1}{g_m} \parallel R_L = \frac{R_L}{1+g_m R_L}$
R_i	r_π	$\frac{1}{g_m}$	r_π
A_v	$-g_m \left(r_o \parallel R_L\right)$	$g_m \left(r_o \parallel R_L\right)$	$\frac{g_m R_L}{1+g_m R_L}$

Thus, we have

$$v_{out} = v_{in} \frac{g_m + \frac{1}{r_o}}{\frac{1}{R_L} + \frac{1}{r_o}} \approx g_m \left(R_L \parallel r_o\right) \cdot v_{in} \tag{3.112}$$

Calculating the input current,

$$i_{in} = \frac{v_{in}}{r_\pi} + \frac{v_{out}}{R_L} = \frac{v_{in}}{r_\pi} + \frac{1}{R_L} \cdot g_m \cdot \frac{R_L r_o}{R_L + r_o} v_{in} \tag{3.113}$$

Therefore, the input resistance is

$$R_i = \frac{v_{in}}{i_{in}} = \frac{1}{\frac{1}{r_\pi} + \frac{g_m r_o}{R_L + r_o}} = r_\pi \parallel \left(\frac{1}{g_m} + \frac{R_L}{g_m r_o}\right) \approx \frac{1}{g_m} + \frac{R_L}{g_m r_o} \tag{3.114}$$

As expected, Eq. 3.114 will be equal to Eq. 3.90 when $R_L = 0$. Note that the load resistance affects the input resistance of the common-base and common-gate amplifier.

This output dependent input resistance is also observed in the common-collector and common-drain amplifiers. It turns out that we have already calculated the input resistance of the circuit in Fig. 3.21 when the emitter node is open-circuited, when we calculated the input resistance of the emitter-degenerated transistor. Thus, the input resistance is

$$R_i \approx r_\pi + R_L + g_m r_\pi R_L \approx r_\pi + g_m r_\pi R_L = r_\pi \left(1 + g_m R_L\right) \tag{3.115}$$

The differences in input resistances are summarized in Table 3.3. Since the CB/CG and CC/CD amplifiers are inherently bilateral, as seen from the small signal equivalent circuits in Figs. 3.19 and 3.21, our unilateral two-port model using just three parameters (A_v, R_i and R_o) will not be enough to describe the behavior of these amplifiers for all possible output conditions.

However, for the CB/CG case, unless R_L is very large, the approximation $R_i \approx \frac{1}{g_m}$ is a good one. Note that even if $R_L = r_o$, the input resistance increases by just a factor of 2. On the other hand the input resistance of the CC/CD amplifier can change drastically, so it is important that we understand the implications of our unilateral amplifier models.

3.7 Single-Stage Amplifier Comparisons

We have looked at the small signal characteristics of the three single-stage transistor amplifier topologies, and the effective two-port parameters are summarized in Table 3.4.

The CE/CS amplifier and the CB/CG amplifier both have relatively large voltage gains, but one is inverting, the other is non-inverting. In larger, more complex amplifiers and applications, a designer might prefer one over the other, as we will see in the succeeding topics. It is interesting to note that the output impedances of the CE/CS and CB/CG amplifiers are rather large, and can only drive resistive loads larger than their output resistance to amplify voltage signals without much loading effects.

The CC/CD amplifier, on the other hand, has a voltage gain very close to unity. However, the output resistance of these amplifiers can be made quite low, depending on the value of g_m. This is very useful in driving small resistive loads, such as audio speakers, whose impedances are typically around $8\,\Omega$.

The input resistance of the CE/CS and CC/CD amplifiers, are relatively large, and in the case of MOSFET amplifiers, $R_i \to \infty$. This is desirable if the driving circuit is a voltage source with a small source resistance, since the gain degradation due to loading effects are reduced. However, if the input is a current source with a high output impedance, it is desirable to use an amplifier with a small input resistance such as the CB/CG amplifier, again reducing the effects of loading.

As we have seen, the different single-stage amplifier topologies can provide us with a varied set of effective two-port small signal parameters. In the remainder of EEE 51, we will use these amplifiers as building blocks towards creating more useful and practical, but more complex amplifiers.