BJT Amplifier Circuits

As we have developed different models for DC signals (simple large-signal model) and AC signals (small-signal model), analysis of BJT circuits follows these steps:

**DC biasing analysis:** Assume all capacitors are open circuit. Analyze the transistor circuit using the simple large signal mode as described in pages 77-78.

**AC analysis:**
1) Kill all DC sources
2) Assume coupling capacitors are short circuit. The effect of these capacitors is to set a lower cut-off frequency for the circuit. This is analyzed in the last step.
3) Inspect the circuit. If you identify the circuit as a prototype circuit, you can directly use the formulas for that circuit. Otherwise go to step 4.
4) Replace the BJT with its small signal model.
5) Solve for voltage and current transfer functions and input and output impedances (node-voltage method is the best).
6) Compute the cut-off frequency of the amplifier circuit.

Several standard BJT amplifier configurations are discussed below and are analyzed. For completeness, circuits include standard bias resistors $R_1$ and $R_2$. For bias configurations that do not utilize these resistors (e.g., current mirror), simply set $R_B = R_1 \parallel R_2 \rightarrow \infty$.

**Common Collector Amplifier (Emitter Follower)**

**DC analysis:** With the capacitors open circuit, this circuit is the same as our good biasing circuit of page 110 with $R_C = 0$. The bias point currents and voltages can be found using procedure of pages 110-112.

**AC analysis:** To start the analysis, we kill all DC sources:
We can combine $R_1$ and $R_2$ into $R_B$ (same resistance that we encountered in the biasing analysis) and replace the BJT with its small signal model:

The figure above shows why this is a common collector configuration: collector is shared between input and output AC signals. We can now proceed with the analysis. Node voltage method is usually the best approach to solve these circuits. For example, the above circuit will have only one node equation for node at point E with a voltage $v_o$:

$$\frac{v_o - v_i}{r_\pi} + \frac{v_o - 0}{r_o} - \beta \Delta i_B + \frac{v_o - 0}{R_E} = 0$$

Because of the controlled source, we need to write an “auxiliary” equation relating the control current ($\Delta i_B$) to node voltages:

$$\Delta i_B = \frac{v_i - v_o}{r_\pi}$$

Substituting the expression for $\Delta i_B$ in our node equation, multiplying both sides by $r_\pi$, and collecting terms, we get:

$$v_i (1 + \beta) = v_o \left[ 1 + \beta + r_\pi \left( \frac{1}{r_o} + \frac{1}{R_E} \right) \right] = v_o \left[ 1 + \beta + \frac{r_\pi}{r_o \parallel R_E} \right]$$

Amplifier Gain can now be directly calculated:

$$A_v \equiv \frac{v_o}{v_i} = \frac{1}{1 + \frac{1}{(1 + \beta)(r_o \parallel R_E)}}$$

Unless $R_E$ is very small (tens of $\Omega$), the fraction in the denominator is quite small compared to 1 and $A_v \approx 1$.

To find the input impedance, we calculate $i_1$ by KCL:

$$i_1 = i_1 + \Delta i_B = \frac{v_i}{R_B} + \frac{v_i - v_o}{r_\pi}$$
Since \( v_o \approx v_i \), we have \( i_i = v_i / R_B \) or
\[
R_i \equiv \frac{v_i}{i_i} = R_B
\]

Note that \( R_B \) is the combination of our biasing resistors \( R_1 \) and \( R_2 \). With alternative biasing schemes which do not require \( R_1 \) and \( R_2 \), (and, therefore \( R_B \to \infty \)), the input resistance of the emitter follower circuit will become large. In this case, we cannot use \( v_o \approx v_i \). Using the full expression for \( v_o \) from above, the input resistance of the emitter follower circuit becomes:
\[
R_i \equiv \frac{v_i}{i_i} = R_B \parallel \left[ r_\pi + \left( R_E \parallel r_o \right) (1 + \beta) \right]
\]

and it is quite large (hundreds of k\(\Omega\) to several M\(\Omega\)) for \( R_B \to \infty \). Such a circuit is in fact the first stage of the 741 OpAmp.

The output resistance of the common collector amplifier (in fact for all transistor amplifiers) is somewhat complicated because the load can be configured in two ways (see figure): First, \( R_E \), itself, is the load. This is the case when the common collector is used as a “current amplifier” to raise the power level and to drive the load. The output resistance of the circuit is \( R_o \) as is shown in the circuit model. This is usually the case when values of \( R_o \) and \( A_i \) (current gain) is quoted in electronic text books.

Alternatively, the load can be placed in parallel to \( R_E \). This is done when the common collector amplifier is used as a buffer (\( A_v \approx 1, R_i \) large). In this case, the output resistance is denoted by \( R'_o \) (see figure). For this circuit, BJT sees a resistance of \( R_E \parallel R_L \). Obviously, if we want the load not to affect the emitter follower circuit, we should use \( R_L \) to be much larger than \( R_E \).
larger than $R_E$. In this case, little current flows in $R_L$ which is fine because we are using this configuration as a buffer and not to amplify the current and power. As such, value of $R'_o$ or $A_i$ does not have much use.

When $R_E$ is the load, the output resistance can be found by killing the source (short $v_i$) and finding the Thevenin resistance of the two-terminal network (using a test voltage source).

KCL: \[ i_T = -\Delta i_B + \frac{v_T}{r_o} - \beta \Delta i_B \]

KVL (outside loop): \[ -r_\pi \Delta i_B = v_T \]

Substituting for $\Delta i_B$ from the 2nd equation in the first and rearranging terms we get:

\[
R_o \equiv \frac{v_T}{i_T} = \frac{(r_o) r_\pi}{(1 + \beta)(r_o) + r_\pi}
\]

Since, $(1 + \beta)(r_o) \gg r_\pi$, the expression for $R_o$ simplifies to

\[
R_o \approx \frac{(r_o) r_\pi}{(1 + \beta)(r_o) + r_\pi} = \frac{r_\pi}{(1 + \beta)} \approx \frac{r_\pi}{\beta} = R_e
\]

As mentioned above, when $R_E$ is the load the common collector is used as a “current amplifier” to raise the current and power levels. This can be seen by checking the current gain in this amplifier: $i_o = v_o/R_E$, $i_i \approx v_i/R_B$ and

\[
A_i \equiv \frac{i_o}{i_i} = \frac{R_B}{R_E}
\]

We can calculate $R'_o$, the output resistance when an additional load is attached to the circuit (i.e., $R_E$ is not the load) with a similar procedure: we need to find the Thevenin resistance of the two-terminal network (using a test voltage source).

We can use our previous results by noting that we can replace $r_o$ and $R_E$ with $r'_o = r_o \parallel R_E$ which results in a circuit similar to the case with no $R_L$. Therefore, $R'_o$ has a similar expression as $R_O$ if we replace $r_o$ with $r'_o$.
\[ R'_o \equiv \frac{v_T}{i_T} = \frac{(r'_o) r_\pi}{(1 + \beta)(r'_o) + r_\pi} \]

In most circuits, \((1 + \beta)(r'_o) \gg r_\pi\) (unless we choose a small value for \(R_E\)) and \(R'_o \approx r_e\).

In summary, the general properties of the common collector amplifier (emitter follower) include a voltage gain of unity \((A_v \approx 1)\), a very large input resistance \(R_i \approx R_B\) (and can be made much larger with alternate biasing schemes). This circuit can be used as buffer for matching impedance, at the first stage of an amplifier to provide very large input resistance (such in 741 OpAmp). The common collector amplifier can be also used as the last stage of some amplifier system to amplify the current (and thus, power) and drive a load. In this case, \(R_E\) is the load, \(R_o\) is small: \(R_o = r_e\) and current gain can be substantial: \(A_i = R_B/R_E\).

**Impact of Coupling Capacitor:**

Up to now, we have neglected the impact of the coupling capacitor in the circuit (assumed it was a short circuit). This is not a correct assumption at low frequencies. The coupling capacitor results in a lower cut-off frequency for the transistor amplifiers. In order to find the cut-off frequency, we need to repeat the above analysis and include the coupling capacitor impedance in the calculation. In most cases, however, the impact of the coupling capacitor and the lower cut-off frequency can be deduced be examining the amplifier circuit model.

Consider our general model for any amplifier circuit. If we assume that coupling capacitor is short circuit (similar to our AC analysis of BJT amplifier), \(v'_i = v_i\).

When we account for impedance of the capacitor, we have set up a high pass filter in the input part of the circuit (combination of the coupling capacitor and the input resistance of the amplifier). This combination introduces a lower cut-off frequency for our amplifier which is the same as the cut-off frequency of the high-pass filter:

\[ \omega_l = 2\pi f_l = \frac{1}{R_i C_c} \]

Lastly, our small signal model is a low-frequency model. As such, our analysis indicates that the amplifier has no upper cut-off frequency (which is not true). At high frequencies, the capacitance between BE, BC, CE layers become important and a high-frequency small-signal model for BJT should be used for analysis. You will see these models in upper division.
courses. Basically, these capacitances results in amplifier gain to drop at high frequencies. PSpice includes a high-frequency model for BJT, so your simulation should show the upper cut-off frequency for BJT amplifiers.

**Common Emitter Amplifier**

DC analysis: Recall that an emitter resistor is necessary to provide stability of the bias point. As such, the circuit configuration as is shown has as a poor bias. We need to include $R_E$ for good biasing (DC signals) and eliminate it for AC signals. The solution to include an emitter resistance and use a “bypass” capacitor to short it out for AC signals as is shown.

For this new circuit and with the capacitors open circuit, this circuit is the same as our good biasing circuit of page 110. The bias point currents and voltages can be found using procedure of pages 110-112.

AC analysis: To start the analysis, we kill all DC sources, combine $R_1$ and $R_2$ into $R_B$ and replace the BJT with its small signal model. We see that emitter is now common between input and output AC signals (thus, common emitter amplifier. Analysis of this circuit is straightforward. Examination of the circuit shows that:

$$v_i = r_\pi \Delta i_B \quad v_o = -(R_C \parallel r_o) \beta \Delta i_B$$

$$A_v \equiv \frac{v_o}{v_i} = -\frac{\beta}{r_\pi} (R_C \parallel r_o) \approx -\frac{\beta}{r_\pi} R_C = -\frac{R_C}{r_e}$$

$$R_i = R_B \parallel r_\pi$$

The negative sign in $A_v$ indicates $180^\circ$ phase shift between input and output. The circuit has a large voltage gain but has a medium value for input resistance.

As with the emitter follower circuit, the load can be configured in two ways: 1) $R_C$ is the load; or 2) load is placed in parallel to $R_C$. The output resistance can be found by killing the source (short $v_i$) and finding the Thevenin resistance of the two-terminal network. For this circuit, we see that if $v_i = 0$ (killing the source), $\Delta i_B = 0$. In this case, the strength of
the dependent current source would be zero and this element would become an open circuit. Therefore,

\[ R_o = r_o \quad R'_o = R_C \parallel r_o \]

**Lower cut-off frequency:** Both the coupling and bypass capacitors contribute to setting the lower cut-off frequency for this amplifier, both act as a high-pass filter with:

\[
\omega_l(\text{coupling}) = 2\pi f_l = \frac{1}{R_i C_c} \\
\omega_l(\text{bypass}) = 2\pi f_l = \frac{1}{R'_E C_b}
\]

where \( R'_E \equiv R_E \parallel r_e \)

Note that usually \( R_E \gg r_e \) and, therefore, \( R'_E \approx r_e \).

In the case when these two frequencies are far apart, the cut-off frequency of the amplifier is set by the “larger” cut-off frequency. *i.e.,*

\[
\omega_l(\text{bypass}) \ll \omega_l(\text{coupling}) \quad \rightarrow \quad \omega_l = 2\pi f_l = \frac{1}{R_i C_c} \\
\omega_l(\text{coupling}) \ll \omega_l(\text{bypass}) \quad \rightarrow \quad \omega_l = 2\pi f_l = \frac{1}{R'_E C_b}
\]

When the two frequencies are close to each other, there is no exact analytical formulas, the cut-off frequency should be found from simulations. An approximate formula for the cut-off frequency (accurate within a factor of two and exact at the limits) is:

\[
\omega_l = 2\pi f_l = \frac{1}{R_i C_c} + \frac{1}{R'_E C_b}
\]
Common Emitter Amplifier with Emitter resistance

A problem with the common emitter amplifier is that its gain depend on BJT parameters $A_v \approx (\beta/r_\pi)R_C$. Some form of feedback is necessary to ensure stable gain for this amplifier. One way to achieve this is to add an emitter resistance. Recall impact of negative feedback on OpAmp circuits: we traded gain for stability of the output. Same principles apply here.

DC analysis: With the capacitors open circuit, this circuit is the same as our good biasing circuit of page 110. The bias point currents and voltages can be found using procedure of pages 110-112.

AC analysis: To start the analysis, we kill all DC sources, combine $R_1$ and $R_2$ into $R_B$ and replace the BJT with its small signal model. Analysis is straight forward using node-voltage method.

$$\frac{v_E - v_i}{r_\pi} + \frac{v_E}{R_E} - \beta \Delta i_B + \frac{v_E - v_o}{r_o} = 0$$

$$\frac{v_o}{R_C} + \frac{v_o - v_E}{r_o} + \beta \Delta i_B = 0$$

$$\Delta i_B = \frac{v_i - v_E}{r_\pi} \quad \text{(Controlled source aux. Eq.)}$$

Substituting for $\Delta i_B$ in the node equations and noting $1 + \beta \approx \beta$, we get:

$$\frac{v_E}{R_E} + \frac{\beta v_E - v_i}{r_\pi} + \frac{v_E - v_o}{r_o} = 0$$

$$\frac{v_o}{R_C} + \frac{v_o - v_E}{r_o} - \beta \frac{v_E - v_i}{r_\pi} = 0$$

Above are two equations in two unknowns ($v_E$ and $v_o$). Adding the two equation together we get $v_E = -(R_E/R_C)v_o$ and substituting that in either equations we can find $v_o$. Using $r_\pi/\beta = r_e$, we get:

$$A_v = \frac{v_o}{v_i} = \frac{R_C}{r_e(1 + R_C/r_o) + R_E(1 + r_e/r_o)} \approx \frac{R_C}{r_e(1 + R_C/r_o) + R_E}$$

where we have simplified the equation noting $r_e \ll r_o$. For most circuits, $R_C \ll r_o$ and $r_e \ll R_E$. In this case, the voltage gain is simply $A_v = -R_C/R_E$.  

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The input resistance of the circuit can be found from (prove it!)

\[ R_i = R_B \parallel \frac{v_i}{\Delta i_B} \]

Noting that \( \Delta i_B = (v_i - v_E)/r_\pi \) and \( v_E = -(R_E/R_C)v_o = -(R_E/R_C)A_v v_i \), we get:

\[ R_i = R_B \parallel \frac{r_\pi}{1 + A_v R_C/R_E} \]

Substituting for \( A_v \) from above (complete expression for \( A_v \) with \( r_e/r_o \ll 1 \)), we get:

\[ R_i = R_B \parallel \left[ \beta \left( \frac{R_E}{1 + R_C/r_o} + r_e \right) \right] \]

For most circuits, \( R_C \ll r_o \) and \( r_e \ll R_E \). In this case, the input resistance is simply \( R_i = R_B \parallel (\beta R_E) \).

As before the minus sign in \( A_v \) indicates a 180° phase shift between input and output signals. Note the impact of negative feedback introduced by the emitter resistance: The voltage gain is independent of BJT parameters and is set by \( R_C \) and \( R_E \) (recall OpAmp inverting amplifier!). The input resistance is also increased dramatically.

As with the emitter follower circuit, the load can be configured in two ways: 1) \( R_C \) is the load. 2) Load is placed in parallel to \( R_C \). The output resistance can be found by killing the source (short \( v_i \)) and finding the Thevenin resistance of the two-terminal network (by attaching a test voltage source to the circuit).

Resistor \( R_B \) drops out of the circuit because it is shorted out. Resistors \( r_\pi \) and \( R_E \) are in parallel. Therefore, \( i_1 = (r_\pi/R_E)\Delta i_B \) and by KCL, \( i_2 = (\beta + 1 + r_\pi/R_E)\Delta i_B \). Then:

\[ i_T = -\Delta i_B - i_1 = -\Delta i_B \left( 1 + \frac{r_\pi}{R_E} \right) \]

\[ v_T = -\Delta i_B r_\pi - i_2 r_o = -\Delta i_B \left[ r_o \left( \beta + 1 + \frac{r_\pi}{R_E} \right) + r_\pi \right] \]

Then:

\[ R_o = \frac{v_T}{i_T} = r_o + R_E \times \frac{1 + r_o/r_\pi}{1 + R_E/r_\pi} \]
where we have used \( r_\pi / \beta = r_e \). Generally \( r_o \gg r_e \) (first approximation below) and for most circuit, \( R_E \ll r_\pi \) (second approximation) leading to

\[
R_o \approx r_o + r_o \times \frac{R_E/r_e}{1 + R_E/r_\pi} \approx r_o + \frac{R_E r_o}{r_e} = r_o \left( \frac{R_E}{r_e} + 1 \right)
\]

Value of \( R'_o \) can be found by a similar procedure. Alternatively, examination of the circuit shows that

\[
R'_o = R_C \parallel R_o \approx R_C
\]

Lower cut-off frequency: The coupling capacitor together with the input resistance of the amplifier lead to a lower cut-off frequency for this amplifier (similar to emitter follower). The lower cut-off frequency is given by:

\[
\omega_l = 2\pi f_l = \frac{1}{R_i C_C}
\]
A Possible Biasing Problem: The gain of the common emitter amplifier with the emitter resistance is approximately $R_C/R_E$. For cases when a high gain (gains larger than 5-10) is needed, $R_E$ may be become so small that the necessary good biasing condition, $V_E = R_E I_E > 1$ V cannot be fulfilled. The solution is to use a by-pass capacitor as is shown. The AC signal sees an emitter resistance of $R_{E1}$ while for DC signal the emitter resistance is the larger value of $R_E = R_{E1} + R_{E2}$. Obviously formulas for common emitter amplifier with emitter resistance can be applied here by replacing $R_E$ with $R_{E1}$ as in deriving the amplifier gain, and input and output impedances, we “short” the bypass capacitor so $R_{E2}$ is effectively removed from the circuit.

The addition of by-pass capacitor, however, modifies the lower cut-off frequency of the circuit. Similar to a regular common emitter amplifier with no emitter resistance, both the coupling and bypass capacitors contribute to setting the lower cut-off frequency for this amplifier. Similarly we find that an approximate formula for the cut-off frequency (accurate within a factor of two and exact at the limits) is:

$$\omega_l = 2\pi f_l = \frac{1}{R_i C_c} + \frac{1}{R'_E C_b}$$

where $R'_E \equiv R_{E2} \parallel (R_{E1} + r_e)$
Summary of BJT Amplifiers*

Common Collector (Emitter Follower):

\[
A_v = \frac{(R_E \| r_o)(1 + \beta)}{r_\pi + (R_E \| r_o)(1 + \beta)} \approx 1 \\
R_i = R_B \| [r_\pi + (R_E \| r_o)(1 + \beta)] \approx R_B \\
R_o = \frac{(r_o) r_\pi}{(1 + \beta)(r_o) + r_\pi} \approx \frac{r_\pi}{\beta} = r_e \\
R'_o = \frac{(r_o') r_\pi}{(1 + \beta)(r_o') + r_\pi} \approx \frac{r_\pi}{\beta} \text{ where } r_o = r_o \| R_C
\]

Common Emitter:

\[
A_v = -\frac{\beta}{r_\pi}(R_C \| r_o) \approx \frac{\beta}{r_\pi} R_C = -\frac{R_C}{r_e} \\
R_i = R_B \| r_\pi \\
R_o = r_o \\
R'_o = R_C \| r_o \approx R_C \\
2\pi f_i = \frac{1}{R_i C_c} + \frac{1}{R'_o C_b} \text{ where } R'_E \equiv R_E \| r_e
\]

Common Emitter with Emitter Resistance:

\[
A_v = -\frac{R_C}{r_e(1 + R_C/r_o) + R_E} \approx \frac{R_C}{r_e + R_E} \approx -\frac{R_C}{R_E} \\
R_i = R_B \| \left[ \beta \left( \frac{R_E}{1 + R_C/r_o} + r_e \right) \right] \approx R_B \| \beta R_E \approx R_B \\
R_o \approx r_o + r_o \times \frac{R_E/r_e}{1 + R_E/r_\pi} \approx r_o \left( \frac{R_E}{r_e} + 1 \right) \\
R'_o = R_C \| R_o \approx R_C \text{ and } 2\pi f_i = \frac{1}{R_i C_c}
\]

Replace \( R_E \) with \( R_{E1} \) in the above formulas except

\[
2\pi f_i = \frac{1}{R_i C_c} + \frac{1}{R'_E C_b} \\
\text{where } R'_E \equiv R_{E2} \| (R_{E1} + r_e)
\]

*If bias resistors are not present (e.g., bias with current mirror), let \( R_B \to \infty \) in the “full” expression for \( R_i \).
Examples of Analysis and Design of BJT Amplifiers

Example 1: Find the bias point and AC amplifier parameters of this circuit (Manufacturers’ spec sheets give: $h_{fe} = 200$, $h_{ie} = 5 \, k\Omega$, $h_{oe} = 10 \, \mu S$).

$$r_\pi = h_{ie} = 5 \, k\Omega \quad r_o = \frac{1}{h_{oe}} = 100 \, k\Omega \quad \beta = h_{fe} = 200 \quad r_e = \frac{r_\pi}{\beta} = 25 \, \Omega$$

DC analysis:

Replace $R_1$ and $R_2$ with their Thevenin equivalent and proceed with DC analysis (all DC current and voltages are denoted by capital letters):

$$R_B = 18 \, k \parallel 22 \, k = 9.9 \, k\Omega$$

$$V_{BB} = \frac{22}{18 + 22} \cdot 9 = 4.95 \, V$$

KVL: $V_{BB} = R_B I_B + V_{BE} + 10^3 I_E$ \quad $I_B = \frac{I_E}{1 + \beta} = \frac{I_E}{201}$

$$4.95 - 0.7 = I_E \left( \frac{9.9 \times 10^3}{201} + 10^3 \right)$$

$I_E = 4 \, mA \approx I_C$, \quad $I_B = \frac{I_C}{\beta} = 20 \, \mu A$

KVL: $V_{CC} = V_{CE} + 10^3 I_E$

$$V_{CE} = 9 - 10^3 \times 4 \times 10^{-3} = 5 \, V$$

DC Bias summary: $I_E \approx I_C = 4 \, mA$, \quad $I_B = 20 \, \mu A$, \quad $V_{CE} = 5 \, V$

AC analysis: The circuit is a common collector amplifier. Using the formulas in page 134,

$$A_v \approx 1$$

$$R_i \approx R_B = 9.9 \, k\Omega$$

$$R_o \approx r_e = 25 \, \Omega$$

$$f_l = \frac{\omega_l}{2\pi} = \frac{1}{2\pi R_B C_c} = \frac{1}{2\pi \times 9.9 \times 10^3 \times 0.47 \times 10^{-6}} = 36 \, Hz$$
Example 2: Find the bias point and AC amplifier parameters of this circuit (Manufacturers’ spec sheets give: $h_{fe} = 200$, $h_{ie} = 5$ kΩ, $h_{oe} = 10$ μS).

$$r_\pi = h_{ie} = 5 \text{ kΩ} \quad r_o = \frac{1}{h_{oe}} = 100 \text{ kΩ} \quad \beta = h_{fe} = 200 \quad r_e = \frac{r_\pi}{\beta} = 25 \text{ Ω}$$

DC analysis: Replace $R_1$ and $R_2$ with their Thevenin equivalent and proceed with DC analysis (all DC current and voltages are denoted by capital letters). Since all capacitors are replaced with open circuit, the emitter resistance for DC analysis is $270 + 240 = 510$ Ω.

$$R_B = 5.9 \text{ k} \parallel 34 \text{ k} = 5.0 \text{ kΩ}$$

$$V_{BB} = \frac{5.9}{5.9 + 34} \times 15 = 2.22 \text{ V}$$

KVL: $V_{BB} = R_B I_B + V_{BE} + 510 I_E \quad I_B = \frac{I_E}{1 + \beta} = \frac{I_E}{201}$

$$2.22 - 0.7 = I_E \left( \frac{5.0 \times 10^3}{201} + 510 \right)$$

$I_E = 3 \text{ mA} \approx I_C, \quad I_B = \frac{I_C}{\beta} = 15 \mu\text{A}$

KVL: $V_{CC} = 1000 I_C + V_{CE} + 510 I_E$

$$V_{CE} = 15 - 1.510 \times 3 \times 10^{-3} = 10.5 \text{ V}$$

DC Bias: $I_E \approx I_C = 3 \text{ mA}, \quad I_B = 15 \mu\text{A}, \quad V_{CE} = 10.5 \text{ V}$

AC analysis: The circuit is a common collector amplifier with an emitter resistance. Note that the 240 Ω resistor is shorted out with the by-pass capacitor. It only enters the formula for the lower cut-off frequency. Using the formulas in page 134 (with $R_{E1} = 270$ Ω):

$$A_v = \frac{R_C}{R_{E1}} = \frac{1.000}{270} = 3.70$$

$$R_i \approx R_B \parallel \beta R_{E1} \approx R_B = 5.0 \text{ kΩ} \quad R_o \approx r_e \left( \frac{R_{E1}}{r_e} + 1 \right) = 1.2 \text{ MΩ}$$

$$R'_E = R_{E2} \parallel (R_{E1} + r_e) = 240 \parallel (270 + 25) = 132 \Omega$$

$$f_l = \frac{\omega_l}{2\pi} = \frac{1}{2\pi R_i C_c} + \frac{1}{2\pi R'_E C_b} =$$

$$\frac{1}{2\pi \times 5,000 \times 4.7 \times 10^{-6}} + \frac{1}{2\pi \times 132 \times 47 \times 10^{-6}} = 31.5 \text{ Hz}$$
Example 3: Design a BJT amplifier with a gain of 4 and a lower cut-off frequency of 100 Hz. The Q point parameters should be $I_C = 3\, mA$ and $V_{CE} = 7.5\, V$. (Manufacturers’ spec sheets give: $\beta_{\text{min}} = 100$, $\beta = 200$, $h_{ie} = 5\, k\Omega$, $h_{oe} = 10\, \mu S$).

\[ r_\pi = h_{ie} = 5\, k\Omega \quad r_o = \frac{1}{h_{oe}} = 100\, k\Omega \quad r_e = \frac{r_\pi}{\beta} = 25\, \Omega \]

The prototype of this circuit is a common emitter amplifier with an emitter resistance. Using formulas of page 134

\[ |A_v| \approx \frac{R_C}{R_E} = 4 \]

The lower cut-off frequency will set the value of $C_c$.

We start with the DC bias: As $V_{CC}$ is not given, we need to choose it. To set the Q-point in the middle of load line, set $V_{CC} = 2V_{CE} = 15\, V$. Then, noting $I_C \approx I_E$:

\[ V_{CC} = R_CI_C + V_{CE} + R_EI_E \quad 15 - 7.5 = 3 \times 10^{-3}(R_C + R_E) \quad \rightarrow \quad R_C + R_E = 2.5\, k\Omega \]

Values of $R_C$ and $R_E$ can be found from the above equation together with the AC gain of the amplifier, $A_V = 4$. Ignoring $r_e$ compared to $R_E$ (usually a good approximation), we get:

\[ \frac{R_C}{R_E} = 4 \quad \rightarrow \quad 4R_E + R_E = 2.5\, k\Omega \quad \rightarrow \quad R_E = 500\, \Omega, R_C = 2. \, k\Omega \]

Commercial values are $R_E = 510\, \Omega$ and $R_C = 2\, k\Omega$. Use these commercial values for the rest of analysis.

We need to check if $V_E > 1\, V$, the condition for good biasing. $V_E = R_EI_E = 510 \times 3 \times 10^{-3} = 1.5 > 1$, it is OK (See next example for the case when $V_E$ is smaller than 1 V).

We now proceed to find $R_B$ and $V_{BB}$. $R_B$ is found from good bias condition and $V_{BB}$ from a KVL in BE loop:

\[ R_B \ll (\beta + 1)R_E \quad \rightarrow \quad R_B = 0.1(\beta_{\text{min}} + 1)R_E = 0.1 \times 101 \times 510 = 5.1\, k\Omega \]

KVL: $V_{BB} = R_BI_B + V_{BE} + R_EI_E$

\[ V_{BB} = 5.1 \times 10^{-3} \times 10^{-3} + 0.7 + 510 \times 3 \times 10^{-3} = 2.28\, V \]
Bias resistors $R_1$ and $R_2$ are now found from $R_B$ and $V_{BB}$:

\[
R_B = R_1 \parallel R_2 = \frac{R_1 R_2}{R_1 + R_2} = 5 \, \text{k}\Omega
\]

\[
\frac{V_{BB}}{V_{CC}} = \frac{R_2}{R_1 + R_2} = \frac{2.28}{15} = 0.152
\]

$R_1$ can be found by dividing the two equations: $R_1 = 33 \, \text{k}\Omega$. $R_2$ is found from the equation for $V_{BB}$ to be $R_2 = 5.9 \, \text{k}\Omega$. Commercial values are $R_1 = 33 \, \text{k}\Omega$ and $R_2 = 6.2 \, \text{k}\Omega$.

Lastly, we have to find the value of the coupling capacitor:

\[
\omega_l = \frac{1}{R_i C_c} = 2\pi \times 100
\]

Using $R_i \approx R_B = 5.1 \, \text{k}\Omega$, we find $C_c = 3 \times 10^{-7} \, \text{F}$ or a commercial values of $C_c = 300 \, \text{nF}$.

So, are design values are: $R_1 = 33 \, \text{k}\Omega$, $R_2 = 6.2 \, \text{k}\Omega$, $R_E = 510 \, \Omega$, $R_C = 2 \, \text{k}\Omega$. and $C_c = 300 \, \text{nF}$.

**Example 4**: Design a BJT amplifier with a gain of 10 and a lower cut-off frequency of 100 Hz. The Q point parameters should be $I_C = 3 \, \text{mA}$ and $V_{CE} = 7.5 \, \text{V}$. A power supply of 15 V is available. Manufacturers’ spec sheets give: $\beta_{min} = 100$, $h_{fe} = 200$, $r_\pi = 5 \, \text{k}\Omega$, $h_{oe} = 10 \, \mu\text{S}$.

\[
\begin{align*}
    r_\pi &= h_{ie} = 5 \, \text{k}\Omega \\
    r_o &= \frac{1}{h_{oe}} = 100 \, \text{k}\Omega \\
    r_e &= \frac{r_\pi}{\beta} = 25 \, \Omega
\end{align*}
\]

The prototype of this circuit is a common emitter amplifier with an emitter resistance. Using formulas of page 134:

\[
|A_v| \approx \frac{R_C}{R_E} = 10
\]

The lower cut-off frequency will set the value of $C_c$.

We start with the DC bias: As the power supply voltage is given, we set $V_{CC} = 15 \, \text{V}$. Then, noting $I_C \approx I_E$:

\[
V_{CC} = R_C I_C + V_{CE} + R_E I_E
\]

\[
15 - 7.5 = 3 \times 10^{-3}(R_C + R_E) \quad \rightarrow \quad R_C + R_E = 2.5 \, \text{k}\Omega
\]
Values of $R_C$ and $R_E$ can be found from the above equation together with the AC gain of the amplifier $A_V = 10$. Ignoring $r_e$ compared to $R_E$ (usually a good approximation), we get:

$$\frac{R_C}{R_E} = 10 \rightarrow 10R_E + R_E = 2.5 \text{ k}\Omega \rightarrow R_E = 227 \text{ } \Omega, R_C = 2.27 \text{ } \text{k}\Omega$$

We need to check if $V_E > 1 \text{ V}$ which is the condition for good biasing: $V_E = R_E I_E = 227 \times 3 \times 10^{-3} = 0.69 < 1$. Therefore, we need to use a bypass capacitor and modify our circuits as is shown.

For DC analysis, the emitter resistance is $R_{E1} + R_{E2}$ while for AC analysis, the emitter resistance will be $R_{E1}$. Therefore:

DC Bias: $R_C + R_{E1} + R_{E2} = 2.5 \text{ k}\Omega$

AC gain: $A_v = \frac{R_C}{R_{E1}} = 10$

Above are two equations in three unknowns. A third equation is derived by setting $V_E = 1 \text{ V}$ to minimize the value of $R_{E1} + R_{E2}$.

$$V_E = (R_{E1} + R_{E2})I_E$$

$$R_{E1} + R_{E2} = \frac{1}{3 \times 10^{-3}} = 333$$

Now, solving for $R_C$, $R_{E1}$, and $R_{E2}$, we find $R_C = 2.2 \text{ k}\Omega$, $R_{E1} = 220 \text{ } \Omega$, and $R_{E2} = 110 \text{ } \Omega$ (All commercial values).

We can now proceed to find $R_B$ and $V_{BB}$:

$$R_B \ll (\beta + 1)(R_{E1} + R_{E2})$$

$$R_B = 0.1(\beta_{\text{min}} + 1)(R_{E1} + R_{E2}) = 0.1 \times 101 \times 330 = 3.3 \text{ k}\Omega$$

KVL: $V_{BB} = R_B I_B + V_{BE} + R_E I_E$

$$V_{BB} = 3.3 \times 10^3 \frac{3 \times 10^{-3}}{201} + 0.7 + 330 \times 3 \times 10^{-3} = 1.7 \text{ V}$$

Bias resistors $R_1$ and $R_2$ are now found from $R_B$ and $V_{BB}$:

$$R_B = R_1 \parallel R_2 = \frac{R_1 R_2}{R_1 + R_2} = 3.3 \text{ k}\Omega$$

$$\frac{V_{BB}}{V_{CC}} = \frac{R_2}{R_1 + R_2} = \frac{1}{15} = 0.066$$
$R_1$ can be found by dividing the two equations: $R_1 = 50$ kΩ and $R_2$ is found from the equation for $V_{BB}$ to be $R_2 = 3.6$ kΩ. Commercial values are $R_1 = 51$ kΩ and $R_2 = 3.6$ kΩ.

Lastly, we have to find the value of the coupling and bypass capacitors:

\[
R'_E = R_{E2} \parallel (R_{E1} + r_e) = 110 \parallel (220 + 25) = 76 \, \Omega
\]

\[
R_i \approx R_B = 3.3 \, k\Omega
\]

\[
\omega_l = \frac{1}{R_i C_c} + \frac{1}{R'_E C_b} = 2\pi \times 100
\]

This is one equation in two unknowns ($C_c$ and $C_B$) so one can be chosen freely. Typically $C_b \gg C_c$ as $R_i \approx R_B \gg R_E \gg R'_E$. This means that unless we choose $C_c$ to be very small, the cut-off frequency is set by the bypass capacitor. The usual approach is the choose $C_b$ based on the cut-off frequency of the amplifier and choose $C_c$ such that cut-off frequency of the $R_i C_c$ filter is at least a factor of ten lower than that of the bypass capacitor. Note that in this case, our formula for the cut-off frequency is quite accurate (see discussion in page 129) and is

\[
\omega_l \approx \frac{1}{R'_E C_b} = 2\pi \times 100
\]

This gives $C_b = 20 \, \mu F$. Then, setting

\[
\frac{1}{R_i C_c} \ll \frac{1}{R'_E C_b}
\]

\[
\frac{1}{R_i C_c} = 0.1 \frac{1}{R'_E C_b}
\]

\[
R_i C_c = 10R'_E C_b \quad \rightarrow \quad C_c = 4.7 \times 10^{-6} = 4.7 \, \mu F
\]

So, are design values are: $R_1 = 50$ kΩ, $R_2 = 3.6$ kΩ, $R_{E1} = 220$ Ω, $R_{E2} = 110$ Ω, $R_C = 2.2$ kΩ, $C_b = 20 \, \mu F$, and $C_c = 4.7 \, \mu F$.

An alternative approach is to choose $C_b$ (or $C_c$) and compute the value of the other from the formula for the cut-off frequency. For example, if we choose $C_b = 47 \, \mu F$, we find $C_c = 0.86 \, \mu F$. 
Example 5: Find the bias point and AC amplifier parameters of this circuit (Manufacturers’ spec sheets give: $\beta = 200$, $r_\pi = 5 \, \text{k}\Omega$, $r_o = 100 \, \text{k}\Omega$).

This is a two-stage amplifier. The first stage (Tr1) is a common emitter amplifier and the second stage (Tr2) is an emitter follower. The two stages are coupled by a coupling capacitor (0.47 $\mu\text{F}$).

DC analysis:
When we replace the coupling capacitors with open circuits, we see the that bias circuits for the two transistors are independent of each other. Each bias circuit can be solved independently.

For Tr1, we replace the bias resistors (6.2k and 33k) with their Thevenin equivalent and proceed with DC analysis:

\[ R_{B1} = 6.2 \, \text{k} \parallel 33 \, \text{k} = 5.22 \, \text{k}\Omega \quad \text{and} \quad V_{BB1} = \frac{6.2}{6.2 + 33} \times 15 = 2.37 \, \text{V} \]

BE-KVL: \[ V_{BB1} = R_{B1}I_{B1} + V_{BE1} + 10^3I_{E1} \quad I_{B1} = \frac{I_{E1}}{1 + \beta} = \frac{I_{E1}}{201} \]
\[ 2.37 - 0.7 = I_{E1}\left(\frac{5.22 \times 10^3}{201} + 500\right) \]
\[ I_{E1} = 3.17 \, \text{mA} \approx I_{C1}, \quad I_{B1} = \frac{I_{C1}}{\beta} = 16 \, \mu\text{A} \]

CE-KVL: \[ V_{CC} = 2 \times 10^3I_{C1} + V_{CE1} + 500I_{E1} \]
\[ V_{CE1} = 15 - 2.5 \times 10^3 \times 3.17 \times 10^{-3} = 7.1 \, \text{V} \]

DC Bias summary for Tr1: \[ I_{E1} \approx I_{C1} = 3.17 \, \text{mA}, \quad I_{B1} = 16 \, \mu\text{A}, \quad V_{CE1} = 7.1 \, \text{V} \]

Following similar procedure for Tr2, we get:

\[ R_{B2} = 18 \, \text{k} \parallel 22 \, \text{k} = 9.9 \, \text{k}\Omega \quad \text{and} \quad V_{BB2} = \frac{22}{18 + 22} \times 15 = 8.25 \, \text{V} \]

BE-KVL: \[ V_{BB2} = R_{B2}I_{B2} + V_{BE2} + 10^3I_{E2} \quad I_{B2} = \frac{I_{E2}}{1 + \beta} = \frac{I_{E2}}{201} \]
\[ 8.25 - 0.7 = I_{E2}\left(\frac{9.9 \times 10^3}{201} + 10^3\right) \]
\[ I_{E2} = 7.2 \text{ mA} \approx I_{C2}, \quad I_{B2} = \frac{I_{C2}}{\beta} = 36 \mu\text{A} \]

CE-KVL: \[ V_{CC} = V_{CE2} + 10^3 I_{E2} \]
\[ V_{CE2} = 15 - 10^3 \times 7.2 \times 10^{-3} = 7.8 \text{ V} \]

DC Bias summary for TR2: \( I_{E2} \approx I_{C2} = 7.2 \text{ mA}, \quad I_{B2} = 36 \mu\text{A}, \quad V_{CE2} = 7.8 \text{ V} \)

**AC analysis:**

We start with the emitter follower circuit (Tr2) as the input resistance of this circuit will appear as the load for the common emitter amplifier (Tr1). Using the formulas in page 134:

\[ A_{v2} \approx 1 \]
\[ R_{i2} \approx R_{B2} = 9.9 \text{ kΩ} \]
\[ f_{l2} = \frac{\omega_{l2}}{2\pi} = \frac{1}{2\pi R_{B2} C_{c2}} = \frac{1}{2\pi \times 9.9 \times 10^3 \times 0.47 \times 10^{-6}} = 34 \text{ Hz} \]

Since \( R_{i2} = 9.9 \text{ kΩ} \) is NOT much larger than the collector resistor of common emitter amplifier (Tr1), it will affect the first circuit. Following discussion in pages 125 and 128, the effect of this load can be taken into by replacing \( R_C \) in common emitter amplifiers formulas with \( R_C' = R_C \parallel R_L = R_{C1} \parallel R_{i2} = 2 \text{ k} \parallel 9.9 \text{ kΩ} = 1.66 \text{ kΩ} \).

\[ |A_{v1}| \approx \frac{R_C'}{R_E} = \frac{1.66k}{500} = 3.3 \]
\[ R_{i1} \approx R_{B1} = 5.22 \text{ kΩ} \]
\[ f_{l1} = \frac{\omega_{l1}}{2\pi} = \frac{1}{2\pi R_{B1} C_{c1}} = \frac{1}{2\pi \times 5.22 \times 10^3 \times 4.7 \times 10^{-6}} = 6.5 \text{ Hz} \]

The overall gain of the two-stage amplifier is then \( A_v = A_{v1} \times A_{v2} = 3.3 \). The input resistance of the two-stage amplifier is the input resistance of the first-stage (Tr1), \( R_i = 9.9 \text{ kΩ} \). To find the lower cut-off frequency of the two-stage amplifier, we note that:

\[ A_{v1}(j\omega) = \frac{A_{v1}}{1 - j\omega_{l1}/\omega} \quad \text{and} \quad A_{v2}(j\omega) = \frac{A_{v2}}{1 - j\omega_{l2}/\omega} \]
\[ A_v(j\omega) = A_{v1}(j\omega) \times A_{v2}(j\omega) = \frac{A_{v1} A_{v2}}{(1 - j\omega_{l1}/\omega)(1 - j\omega_{l2}/\omega)} \]

From above, it is clear that the maximum value of \( A_v(j\omega) \) is \( A_{v1} A_{v2} \) and the cut-off frequency, \( \omega_l \) can be found from \(|A_v(j\omega = \omega_l)| = A_{v1} A_{v2} / \sqrt{2} \) (similar to procedure we used for filters). For the circuit above, since \( \omega_{l2} \gg \omega_{l1} \) the lower cut-off frequency would be very close to \( \omega_{l2} \). So, the lower-cut-off frequency of this amplifier is 34 Hz.
Example 6: Find the bias point and AC amplifier parameters of this circuit (Manufacturers’ spec sheets give: $\beta = 200$, $r_e = 5 \, \text{k}\Omega$, $r_o = 100 \, \text{k}\Omega$).

This is a two-stage amplifier. The first stage (Tr1) is a common emitter amplifier and the second stage (Tr2) is an emitter follower. The circuit is similar to the two-stage amplifier of Example 5. The only difference is that Tr2 is directly biased from Tr1 and there is no coupling capacitor between the two stages. This approach has its own advantages and disadvantages that are discussed at the end of this example.

DC analysis:

Since the base current in BJTs are typically much smaller than the collector current, we start by assuming $I_{C1} \gg I_{B2}$. In this case, $I_1 = I_{C1} + I_{B2} \approx I_{C1} \approx I_{E1}$ (the bias current $I_{B2}$ has no effect on bias parameters of Tr1). This assumption simplifies the analysis considerably and we will check the validity of this assumption later.

For Tr1, we replace the bias resistors (6.2k and 33k) with their Thevenin equivalent and proceed with DC analysis:

$$R_{B1} = 6.2 \, \text{k} \parallel 33 \, \text{k} = 5.22 \, \text{k}\Omega \quad \text{and} \quad V_{BB1} = \frac{6.2}{6.2 + 33} \times 15 = 2.37 \, \text{V}$$

BE-KVL: $V_{BB1} = R_{B1}I_{B1} + V_{BE1} + 10^3I_{E1}$

$$I_{B1} = \frac{I_{E1}}{1 + \beta} = \frac{I_{E1}}{201}$$

$$2.37 - 0.7 = I_{E1} \left( \frac{5.22 \times 10^3}{201} + 500 \right)$$

$$I_{E1} = 3.17 \, \text{mA} \approx I_{C1}, \quad I_{B1} = \frac{I_{C1}}{\beta} = 16 \, \mu\text{A}$$

CE-KVL: $V_{CC} = 2 \times 10^3I_{C1} + V_{CE1} + 500I_{E1}$

$$V_{CE1} = 15 - 2.5 \times 10^3 \times 3.17 \times 10^{-3} = 7.1 \, \text{V}$$

DC Bias summary for Tr1: $I_{E1} \approx I_{C1} = 3.17 \, \text{mA}$, $I_{B1} = 16 \, \mu\text{A}$, $V_{CE1} = 7.1 \, \text{V}$

To find the bias point of TR2, we note:

$$V_{B2} = V_{CE1} + 500 \times I_{E1} = 7.1 + 500 \times 3.17 \times 10^{-3} = 8.68 \, \text{V}$$
BE-KVL: \[ V_{B2} = V_{BE2} + 10^3 I_{E2} \]
\[ 8.68 - 0.7 = 10^3 I_{E2} \]
\[ I_{E2} = 8.0 \text{ mA} \approx I_{C2}, \quad I_{B2} = \frac{I_{C2}}{\beta} = 40 \mu\text{A} \]

KVL: \[ V_{CC} = V_{CE2} + 10^3 I_{E2} \]
\[ V_{CE2} = 15 - 10^3 \times 8.0 \times 10^{-3} = 7.0 \text{ V} \]

DC Bias summary for TR2: \[ I_{E2} \approx I_{C2} = 8.0 \text{ mA}, \quad I_{B2} = 40 \mu\text{A}, \quad V_{CE2} = 7.0 \text{ V} \]

We now check our assumption of \( I_{C1} \gg I_{B2} \). We find \( I_{C1} = 3.17 \text{ mA} \gg I_{B2} = 41 \mu\text{A} \). So, our assumption was justified.

It should be noted that this bias arrangement is also stable to variation in transistor \( \beta \). The bias resistors in the first stage will ensure that \( I_{C1} (\approx I_{E1}) \) and \( V_{CE1} \) is stable to variation of TR1 \( \beta \). Since \( V_{B2} = V_{CE1} + R_{E1} \times I_{E1} \), \( V_{B2} \) will also be stable to variation in transistor \( \beta \). Finally, \( V_{B2} = V_{BE2} + R_{E2} I_{E2} \). Thus, \( I_{C2} (\approx I_{E2}) \) will also be stable (and \( V_{CE2} \) because of CE-KVL).

AC analysis:

As in Example 5, we start with the emitter follower circuit (Tr2) as the input resistance of this circuit will appear as the load for the common emitter amplifier (Tr1). Using the formulas in page 134 and noting that this amplifier does not have bias resistors \( (R_{B1} \rightarrow \infty) \):

\[ A_{v2} \approx 1 \]
\[ R_{i2} = r_\pi + (R_E \parallel r_o)(1 + \beta) = 5 \times 10^3 + 201 \times 10^3 = 201 \text{ k}\Omega \]

Note that because of the absence of the bias resistors, the input resistance of the circuit is very large, and because of the absence of the coupling capacitors, there is no lower cut-off frequency for this stage.

Since \( R_{i2} = 201 \text{ k}\Omega \) is much larger than the collector resistor of common emitter amplifier (Tr1), it will NOT affect the first circuit. The parameters of the first-stage common emitter amplifier can be found using formulas of page 134.

\[ |A_{v1}| \approx \frac{R_C}{R_E} = \frac{2,000}{500} = 4 \]
\[ R_{i1} \approx R_{B1} = 5.22 \text{ k}\Omega \]
\[ f_{11} = \frac{\omega_{11}}{2\pi} = \frac{1}{2\pi R_{B1}C_{c1}} = \frac{1}{2\pi \times 5.22 \times 10^3 \times 4.7 \times 10^{-6}} = 6.5 \text{ Hz} \]
The overall gain of the two-stage amplifier is then \( A_v = A_{v1} \times A_{v2} = 4 \). The input resistance of the two-stage amplifier is the input resistance of the first-stage (Tr1), \( R_i = 9.9 \, \text{k}\Omega \). The find the lower cut-off frequency of the two-stage amplifier is 6.5 Hz.

The two-stage amplifier of Example 6 has many advantages over that of Example 5. It has three less elements. Because of the absence of bias resistors, the second-stage does not load the first stage and the overall gain is higher. Also because of the absence of a coupling capacitor between the two-stages, the overall cut-off frequency of the circuit is lower. Some of these issues can be resolved by design, e.g., use a large capacitor for coupling the two stages, use a large \( R_{E2} \), etc.. The drawback of the Example 6 circuit is that the bias circuit is more complicated and harder to design.