1. For Darlington emitter follower circuit, obtain expression for overall current gain.

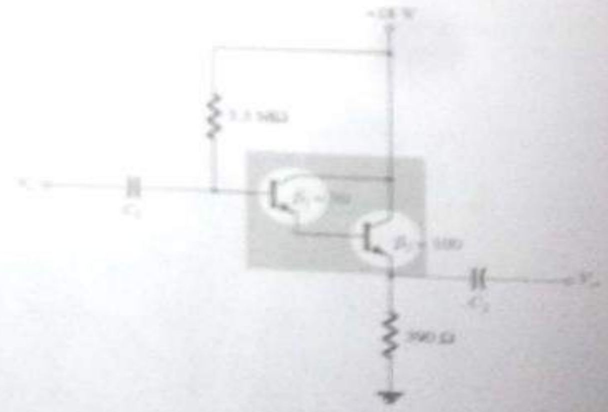


FIG. 5.76
Circuit for Example 5.17

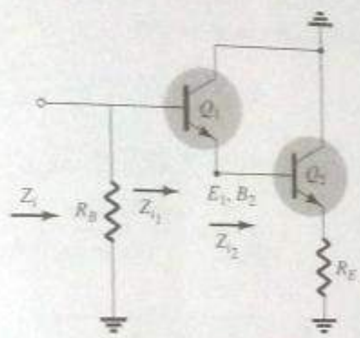


FIG. 5.77
Finding Z_i

As defined in Fig. 5.77:

$$Z_{i_2} = \beta_2(r_{e_2} + R_E)$$
$$Z_{i_1} = \beta_1(r_{e_1} + Z_{i_2})$$

so that

$$Z_{i_1} = \beta_1(r_{e_1} + \beta_2(r_{e_2} + R_E))$$

Assuming

$$R_E \gg r_{e_2}$$

and

$$Z_{i_1} = \beta_1(r_{e_1} + \beta_2 R_E)$$

Since

$$\beta_2 R_E \gg r_{e_1}$$
$$Z_{i_1} \cong \beta_1\beta_2 R_E$$

and since

$$Z_i = R_B \| Z_{i_1}$$

$$\boxed{Z_i = R_B \| \beta_1\beta_2 R_E = R_B \| \beta_D R_E} \qquad (5.108)$$

For the network of Fig. 5.76

$$\begin{aligned} Z_i &= R_B \| \beta_D R_E \\ &= 3.3\ \text{M}\Omega \| (5000)(390\ \Omega) = 3.3\ \text{M}\Omega \| 1.95\ \text{M}\Omega \\ &= \mathbf{1.38\ M\Omega} \end{aligned}$$

not compared but dropped com-

AC Current Gain The current gain can be determined from the equivalent network of Fig. 5.78. The output impedance of each transistor is ignored and the parameters for each transistor are employed.

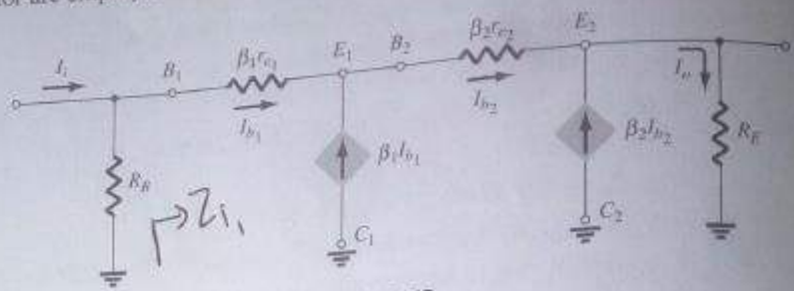


FIG. 5.78
Determining A_i for the network of Fig. 5.75.

Solving for the output current: $I_o = I_{b_2} + \beta_2 I_{b_2} = (\beta_2 + 1)I_{b_2}$

with $I_{b_2} = \beta_1 I_{b_1} + I_{b_1} = (\beta_1 + 1)I_{b_1}$

Then $I_o = (\beta_2 + 1)(\beta_1 + 1)I_{b_1}$

Using the current-divider rule on the input circuit:

$$I_{b_1} = \frac{R_B}{R_B + Z_i}I_i = \frac{R_B}{R_B + \beta_1\beta_2 R_E}I_i$$

and

$$I_o = (\beta_2 + 1)(\beta_1 + 1)\left(\frac{R_B}{R_B + \beta_1\beta_2 R_E}\right)I_i$$

so that

$$A_i = \frac{I_o}{I_i} = \frac{(\beta_1 + 1)(\beta_2 + 1)R_B}{R_B + \beta_1\beta_2 R_E}$$

Using $\beta_1, \beta_2 \gg 1$

$$\boxed{A_i = \frac{I_o}{I_i} \cong \frac{\beta_1\beta_2 R_B}{R_B + \beta_1\beta_2 R_E}} \qquad (5.109)$$

or

$$\boxed{A_i = \frac{I_o}{I_i} \cong \frac{\beta_D R_B}{R_B + \beta_D R_E}} \qquad (5.110)$$

For Fig. 5.76:

$$A_i = \frac{I_o}{I_i} = \frac{\beta_D R_B}{R_B + \beta_D R_E} = \frac{(5000)(3.3\ \text{M}\Omega)}{3.3\ \text{M}\Omega + 1.95\ \text{M}\Omega}$$
$$= \mathbf{3.14 \times 10^3}$$

AC Voltage Gain The voltage gain can be determined using Fig. 5.77 and the following derivation:

$$V_o = I_o R_E$$
$$V_i = I_i(R_B \| Z_i)$$
$$R_B \| Z_i = R_B \| \beta_D R_E = \frac{\beta_D R_B R_E}{R_B + \beta_D R_E}$$

and

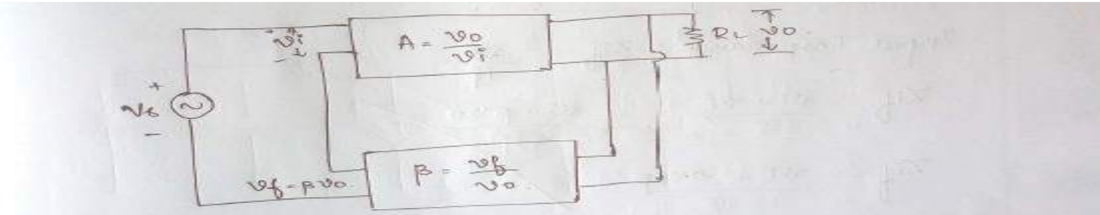
$$A_v = \frac{V_o}{V_i} = \frac{I_o R_E}{I_i(R_B \| Z_i)} = (A_i)\left(\frac{R_E}{R_B \| Z_i}\right)$$
$$= \left[\frac{\beta_D R_B}{R_B + \beta_D R_E}\right]\left[\frac{R_E}{\dfrac{\beta_D R_B R_E}{R_B + \beta_D R_E}}\right]$$

and

$$\boxed{A_v \cong 1 \text{ (in reality less than one)}} \qquad (5.111)$$

an expected result for the emitter-follower configuration.

2. For voltage series feedback topology obtain expressions for gain, input impedance and output impedance.

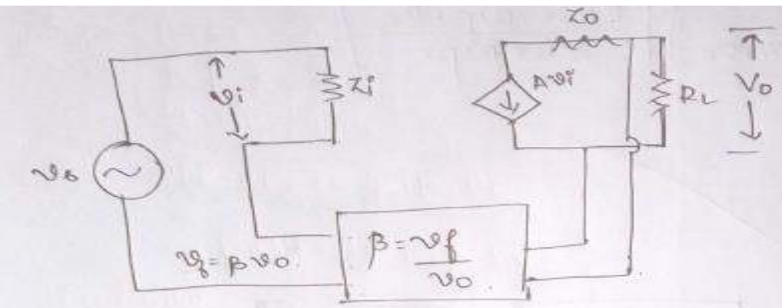


With feedback

$$\text{Gain} \quad A = \frac{v_o}{v_i} = \frac{v_o}{v_s} = \frac{v_o}{v_i + v_f} = \frac{v_o}{v_i + \beta v_o}$$

$$\text{Gain } A_v = \frac{v_o}{\frac{v_o}{A} + \beta v_o} = \frac{v_o}{v_o\left[\frac{1 + A\beta}{A}\right]}$$

$$\text{Gain} = \frac{A}{1 + A\beta}$$



Input impedance : $Z_{if} = \frac{v_s}{i_s}$

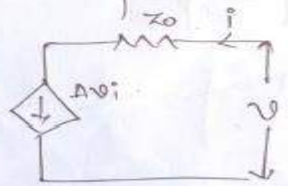
$$Z_{if} = \frac{v_i + v_f}{i_i} = \frac{v_i + \beta v_o}{i_i}$$

$$Z_{if} = \frac{v_i + A v_i \beta}{i_i}$$

$$Z_{if} = \frac{v_i}{i_i}\left[1 + A\beta\right]$$

$$Z_{if} = Z_i\left[1 + A\beta\right]$$

Output impedance : Z_{of}



Apply KVL to the loop

$$v - iZ_o - Av_i = 0$$

$$v = iZ_o + Av_i \qquad [v_i = -v_f]$$

$$v = iZ_o - Av_f$$

$$\cancel{v + Av_f}$$

$$v = iz_o - \beta v$$

$$v + \beta v = iz_o$$

$$v(1+\beta) = iz_o$$

$$\frac{v}{i}(1+\beta) = z_o$$

$$z_{of} = z_o(1+\beta).$$

3a. Explain the block diagram of feedback amplifier

Feedback has been mentioned previously, in particular, in op-amp circuits as described in Chapters 10 and 11. Depending on the relative polarity of the signal being fed back into a circuit, one may have negative or positive feedback. Negative feedback results in decreased voltage gain, for which a number of circuit features are improved, as summarized below. Positive feedback drives a circuit into oscillation as in various types of oscillator circuits.

A typical feedback connection is shown in Fig. 14.1. The input signal V_s is applied to a mixer network, where it is combined with a feedback signal V_f. The difference of these signals V_i is then the input voltage to the amplifier. A portion of the amplifier output V_o is connected to the feedback network (β), which provides a reduced portion of the output as feedback signal to the input mixer network.

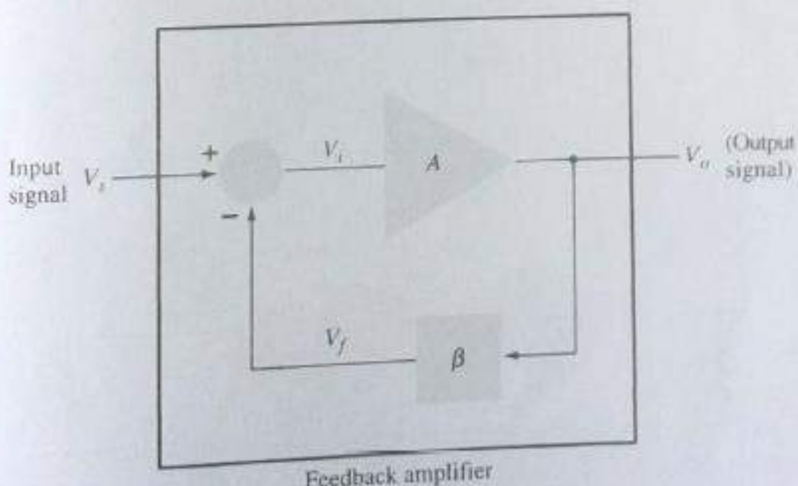


FIG. 14.1
Simple block diagram of feedback amplifier.

If the feedback signal is of opposite polarity to the input signal, as shown in Fig. 14.1, negative feedback results. Although negative feedback results in reduced overall voltage gain, a number of improvements are obtained, among them being:

1. Higher input impedance.
2. Better stabilized voltage gain.
3. Improved frequency response.
4. Lower output impedance.
5. Reduced noise.
6. More linear operation.

3b. With neat diagram explain cascade amplifier.

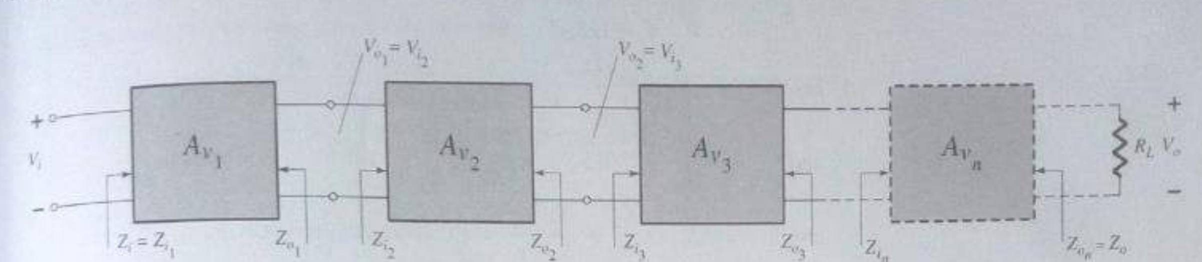
The two-port systems approach is particularly useful for cascaded systems such as that appearing in Fig. 5.67, where A_{v_1}, A_{v_2}, A_{v_3}, and so on, are the voltage gains of each stage *under loaded conditions*. That is, A_{v_1} is determined with the *input impedance to* A_{v_2} *acting as the load on* A_{v_1}. For A_{v_2}, A_{v_1} will determine the signal strength and source impedance at the input to A_{v_2}. The total gain of the system is then determined by the product of the individual gains as follows:

$$\boxed{A_{v_T} = A_{v_1} \cdot A_{v_2} \cdot A_{v_3} \cdot \cdots} \tag{5.99}$$

and the total current gain is given by

$$\boxed{A_{i_T} = -A_{v_T} \frac{Z_{i_1}}{R_L}} \tag{5.100}$$

No matter how perfect the system design, the application of a succeeding stage or load to a two-port system will affect the voltage gain. Therefore, there is no possibility of a situation where A_{v_1}, A_{v_2}, and so on, of Fig. 5.67 are simply the no-load values. The no-load parameters can be used to determine the loaded gains of each stage, but Eq. (5.99) requires the loaded values. The load on stage 1 is Z_{i_2}, on stage 2 Z_{i_3}, on stage 3 Z_{i_4}, and so on.



4. Explain the effect of negative feedback on (i) Gain Stability (ii) Distortion

amplifier circuit.

Reduction in Noise and Nonlinear Distortion

Signal feedback tends to hold down the amount of noise signal (such as power-supply hum) and nonlinear distortion. The factor $(1 + \beta A)$ reduces both input noise and resulting nonlinear distortion for considerable improvement. However, there is a reduction in overall gain (the price required for the improvement in circuit performance). If additional stages are used to bring the overall gain up to the level without feedback, the extra stage(s) might introduce as much noise back into the system as that reduced by the feedback amplifier. This problem can be somewhat alleviated by readjusting the gain of the feedback-amplifier circuit to obtain higher gain while also providing reduced noise signal.

Effect of Negative Feedback on Gain and Bandwidth

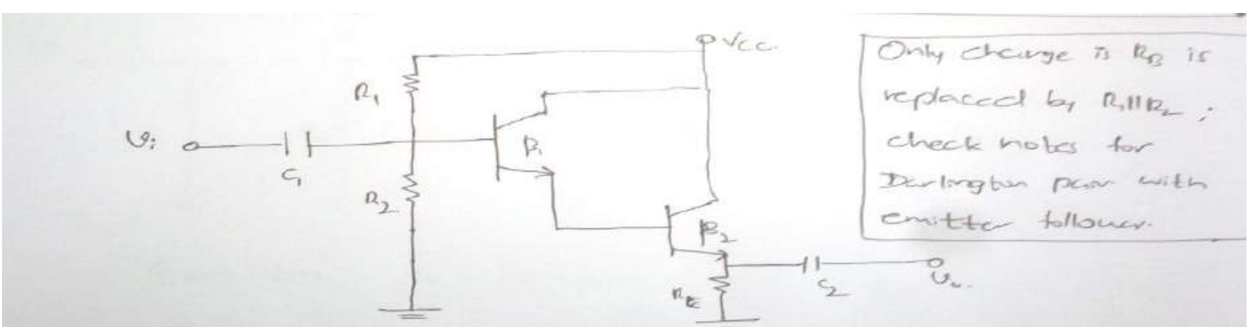
In Eq. (14.2), the overall gain with negative feedback is shown to be

$$A_f = \frac{A}{1 + \beta A} \cong \frac{A}{\beta A} = \frac{1}{\beta} \qquad \text{for } \beta A \gg 1$$

As long as $\beta A \gg 1$, the overall gain is approximately $1/\beta$. For a practical amplifier (for single low- and high-frequency breakpoints) the open-loop gain drops off at high frequencies due to the active device and circuit capacitances. Gain may also drop off at low frequencies for capacitively coupled amplifier stages. Once the open-loop gain A drops low enough and the factor βA is no longer much larger than 1, the conclusion of Eq. (14.2) that $A_f \cong 1/\beta$ no longer holds true.

Figure 14.6 shows that the amplifier with negative feedback has more bandwidth (B_f) than the amplifier without feedback (B). The feedback amplifier has a higher upper 3-dB frequency and smaller lower 3-dB frequency.

5. Draw the circuit of Darlington emitter follower with voltage divider bias. Calculate input impedance, voltage gain and output impedance. Take $\beta_1 = \beta_2 = 100$. $R_1 = R_2 = 100k\Omega$; $R_E = 5k\Omega$, Take $r_e = 100\Omega$.



No need to do DC analysis; as 're' value is given in the problem.

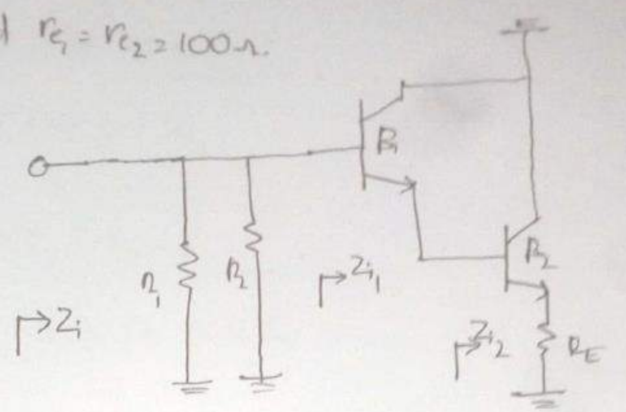
AC ANALYSIS TO FIND Z_i, A_v and Z_o :-

$\beta_1 = \beta_2 = 100$ and $r_{e_1} = r_{e_2} = 100\,\Omega$.

$$Z_{i_2} = \beta_2 (r_{e_2} + R_E)$$

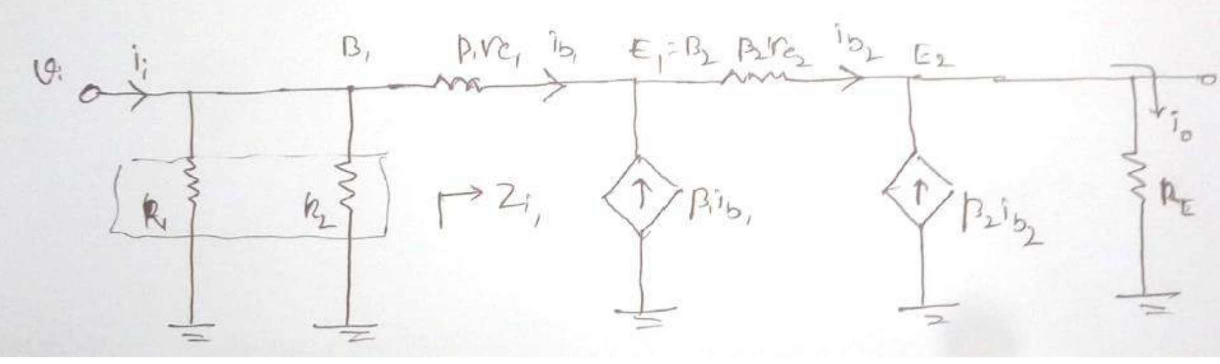
$$Z_{i_1} = \beta_1 (r_{e_1} + Z_{i_2}).$$

$$Z_{i_1} \simeq \beta_1 \beta_2 R_E$$



$$\boxed{Z_i = R_1 \| R_2 \| \beta_1 \beta_2 R_E = R_1 \| R_2 \| \beta^2 R_E}$$

CURRENT GAIN (A_i):-



from KCL; $i_o = i_{b_2} + \beta_2 i_{b_2}$

$$= i_{b_2}(1+\beta_2) = (i_{b_1} + \beta_1 i_{b_1})(1+\beta_2)$$

$$= i_{b_1}(1+\beta_1)(1+\beta_2) = \beta_1 \beta_2 i_{b_1}$$

from CDR @ i/p; $i_{b_1} = \dfrac{(R_1 \| R_2)}{(R_1 \| R_2) + Z_{i_1}} \, i_i$

$$i_i = i_{b_1}\left[\frac{(R_1 \| R_2) + Z_{i_1}}{(R_1 \| R_2)}\right]$$

$$\therefore A_{i_1} = \frac{i_o}{i_i} = \frac{\beta_1 \beta_2 \cancel{R_E} (R_1 \| R_2)}{(R_1 \| R_2) + Z_{i_1}}$$

$$A_i = \frac{\beta_1 \beta_2 (R_1 \| R_2)}{(R_1 \| R_2) + \beta_1 \beta_2 R_E}$$

VOLTAGE GAIN (A_V):-

$$A_V = \frac{v_o}{v_i} = \frac{i_o R_E}{Z_i \cdot i_i}$$

$$= \frac{i_o}{i_i} \cdot \frac{R_E}{Z_i}$$

$$A_V = \left[\frac{\beta_1 \beta_2 (R_1 \| R_2)}{(R_1 \| R_2) + \beta_1 \beta_2 R_E}\right]\left[\frac{R_E}{R_1 \| R_2 \| \beta^2 R_E}\right]$$

OUTPUT IMPEDANCE (Z_o):—

$$Z_o = \frac{r_{e_1}}{\beta_2} + r_{e_2}$$

$$\boxed{Z_o = \frac{100}{100} + 100 = 101\Omega}$$

Verify Your Answers; Z_i should be very high

$A_v \simeq 1$

A_i should be very high.

6. Two amplifiers are cascaded. The load resistance R_L = 20 kΩ, and internal resistance of the source is 2kΩ. Find the

(i)Loaded voltage gain of each stage

(ii)Total voltage gain of cascaded amplifier with R_s

(iii)Current gain of cascaded amplifier

(iv)Output Impedance

The first stage has No load voltage gain of 1, Input impedance = 500kΩ,Output Impedance = 1kΩ. The second stage has a No load gain of 300, input impedance of 1kΩ and output impedance of 50kΩ.

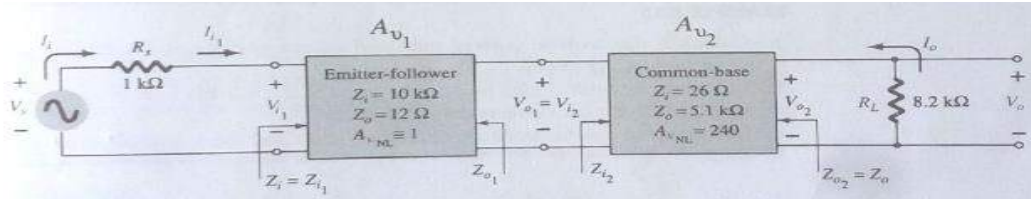


FIG. 5.68
Example 5.14.

Solution:

a. For the emitter-follower configuration, the loaded gain is (by Eq. (5.94))

$$V_{o_1} = \frac{Z_{i_2}}{Z_{i_2} + Z_{o_1}} A_{v_{NL}} V_{i_1} = \frac{26\ \Omega}{26\ \Omega + 12\ \Omega}(1)\, V_{i_1} = 0.684\, V_{i_1}$$

and $$A_{v_1} = \frac{V_{o_1}}{V_{i_1}} = \mathbf{0.684}$$

For the common-base configuration,

$$V_{o_2} = \frac{R_L}{R_L + R_{o_2}} A_{v_{NL}} V_{i_2} = \frac{8.2\ \text{k}\Omega}{8.2\ \text{k}\Omega + 5.1\ \text{k}\Omega}(240)\, V_{i_2} = 147.97\, V_{i_2}$$

and $$A_{v_2} = \frac{V_{o_2}}{V_{i_2}} = \mathbf{147.97}$$

b. Eq. (5.99): $A_{v_T} = A_{v_1} A_{v_2}$
$= (0.684)(147.97)$
$= \mathbf{101.20}$

$$\text{Eq. (5.91):}\quad A_{v_s} = \frac{Z_{i_1}}{Z_{i_1} + R_s} A_{v_T} = \frac{(10\text{ k}\Omega)(101.20)}{10\text{ k}\Omega + 1\text{ k}\Omega}$$

$$= \mathbf{92}$$

c. $$\text{Eq. (5.100):}\quad A_{i_T} = -A_{v_T}\frac{Z_{i_1}}{R_L} = -(101.20)\left(\frac{10\text{ k}\Omega}{8.2\text{ k}\Omega}\right)$$

$$= \mathbf{-123.41}$$

d. $$\text{Eq. (5.91):}\quad V_i = \frac{Z_{i_{CB}}}{Z_{i_{CB}} + R_s} V_s = \frac{26\ \Omega}{26\ \Omega + 1\text{ k}\Omega} V_s = 0.025\ V_s$$

and $\dfrac{V_i}{V_s} = 0.025$ with $\dfrac{V_o}{V_i} = 147.97$ from above

and $$A_{v_s} = \frac{V_o}{V_s} = \frac{V_i}{V_s} \cdot \frac{V_o}{V_i} = (0.025)(147.97) = \mathbf{3.7}$$

In total, therefore, the gain is about 25 times greater with the emitter-follower configuration to draw the signal to the amplifier stages. Note, however, that it is also important that the output impedance of the first stage is relatively close to the input impedance of the second stage, otherwise the signal would have been "lost" again by the voltage-divider action.