The small-signal circuit of the feedback circuit is seen in Fig. 31.14. Note that the forward path consists of the following nodes: 1, 2, 3. The feedback path consists of nodes 3 and 4, with the feedback variable, v_f, appearing across R_1. In the previous discussion, it was assumed that the β network did not load the amplifier circuit. However, to accurately calculate the open-loop gain, A_{OL}, the loading of R_1 and R_2 on both the input and the output of the amplifier circuit needs to be considered. Note that the resistor R_S is initially ignored, since it is essentially outside the feedback amplifier.

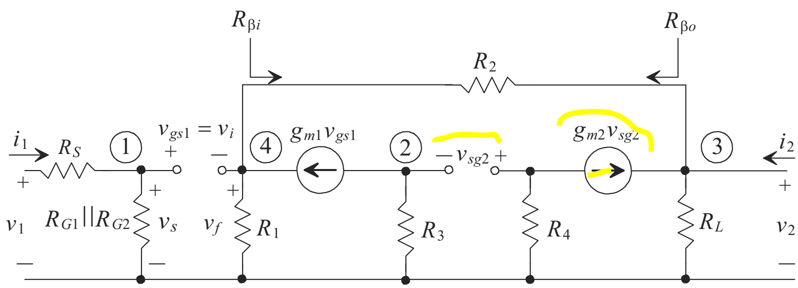

Figure 31.14 Closed-loop small-signal model of Fig. 31.13.

Since we are analyzing a series-shunt amplifier, we may determine the loading caused by the β network on the input, $R_{\beta i}$ and the output $R_{\beta o}$ in the following way (refer to Fig. 31.15). Looking into the β network from the input, we observe the resistance seen with the output terminal shorted to ground. The equivalent resistance to ground seen is the loading of the β network seen by the input of the amplifier. In this example, R_2 is seen. Therefore, in the open-loop model used to determine A_{OL}, we will include R_2 in parallel with R_1. The loading at the output is found similarly. Since the input mixing is series, we will remove M1 "out-of-socket" and look into the β network from the output. The equivalent resistance seen is then attached to the output of the open-loop model. In this example, the equivalent resistance is $R_2 + R_1$ and is attached to the output of the

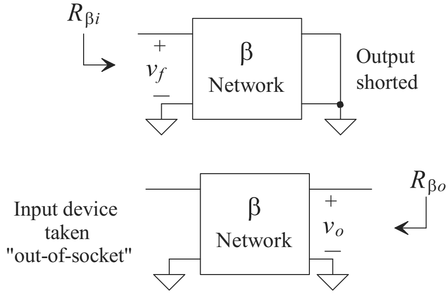

Figure 31.15 Determining the loading due to the feedback network for a series-shunt amplifier.

open-loop model. The resulting open-loop model is seen in Fig. 31.16. We will initially assume that r_o for the MOSFETs is much larger than the discrete resistors and that the bulk and source are tied together (v_{sb} = 0). As we progress through the chapter, more difficult circuits will include drain-to-source resistances in our small-signal analysis.

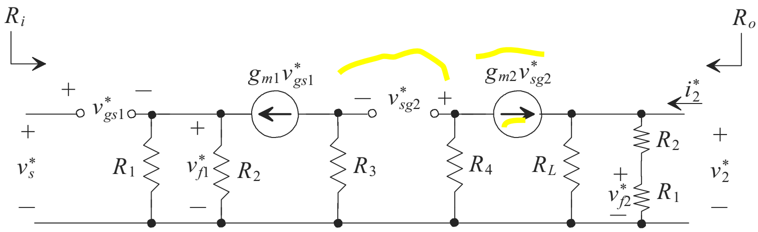

Figure 31.16 Open-loop small-signal model of Fig. 31.13.

The open-loop model is now ready to be analyzed in order to calculate A_{OL}. Since we are using a series (voltage)-shunt (voltage) feedback amplifier, the units of A_{OL} will be V/V and

$$A_{OL} = \frac{v_2^*}{v_s^*} \tag{31.34}$$

Solving for A_{OL} yields

$$A_{OL} = \frac{v_2^*}{v_s^*} = \left(\frac{v_2^*}{v_{sg2}^*}\right)\left(\frac{v_{sg2}^*}{v_{gs1}^*}\right)\left(\frac{v_{gs1}^*}{v_s^*}\right) = [-g_{m2}R_L||(R_2+R_1)]\left[-\frac{g_{m1}R_3}{1+g_{m2}R_4}\right]\left[\frac{1}{1+g_{m1}(R_1||R_2)}\right] \tag{31.35}$$

Next, the value of β can also be calculated from the open-loop model. Remembering that β is defined as the gain from the output back to the input mixing variable, v_f, we can write

$$\beta = \frac{v_f^*}{v_2^*} = \frac{R_1}{R_1+R_2} \tag{31.36}$$

since the β network is simply a voltage divider relationship. Notice that the open-loop circuit now contains two values of R_2 and v_f^*. In this example, since r_o was assumed to be infinite, the gain from v_2^* to v_{f1}^* will be zero. If r_o had not been neglected, the gain from v_2^* to v_{f1}^* would have been small but finite. Therefore, it can be said that a reverse path exists through the basic amplifier as well as through the feedback network. However, the gain from v_2^* to v_{f2}^*, though less than one, will be significantly larger than from v_2^* to v_{f1}^*. *Therefore, just as the forward path through the feedback network was neglected, the reverse path through the basic amplifier is assumed to be much smaller than the reverse path through the feedback path. Therefore, the value of* β *is calculated using the resistor, R_2, closest to the output.*

Next, the value for R_i and R_o will be calculated. These values are determined using the open-loop model generated in Fig. 31.16. Since we are using MOS devices, it should

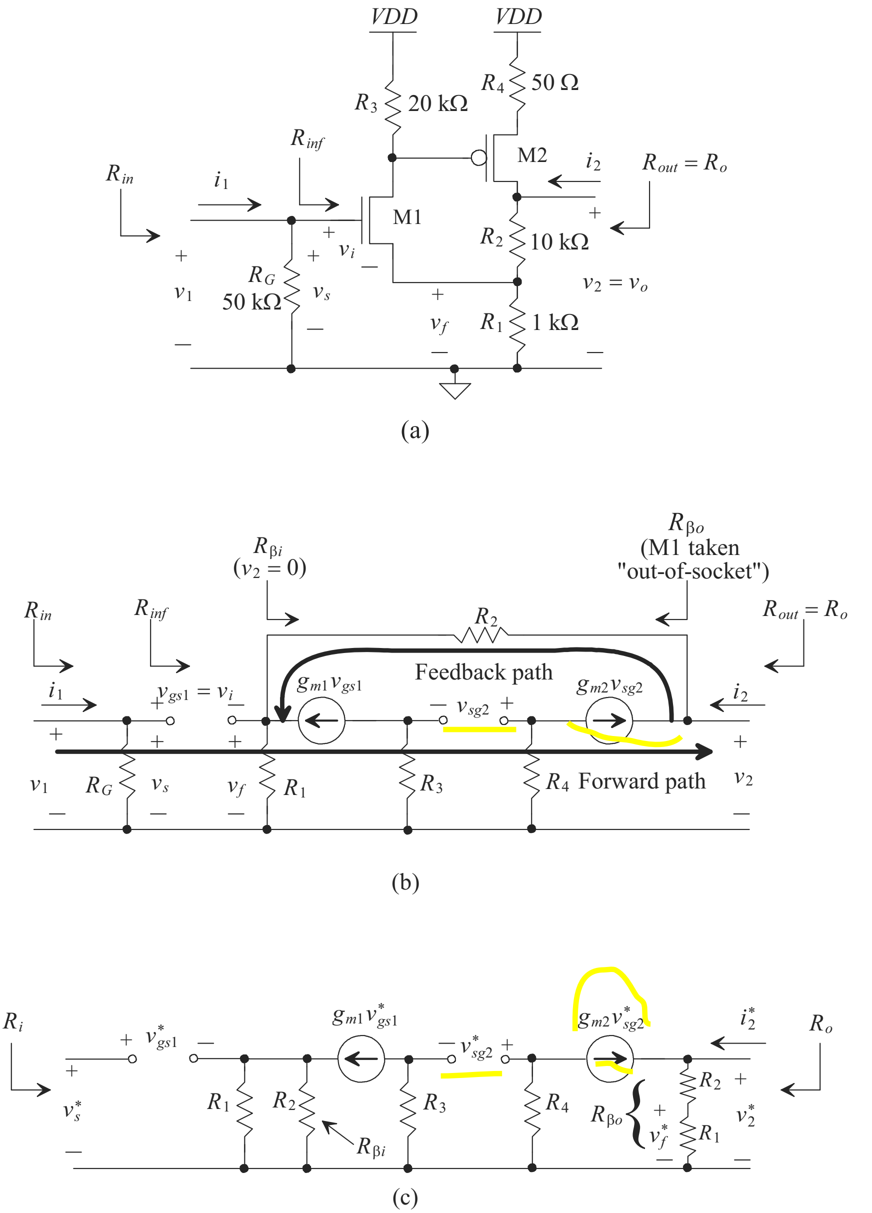

Figure 31.17 (a) Series-shunt circuit used in Ex. 31.1; (b) its closed-loop small-signal model; and (c) the resulting open-loop model.

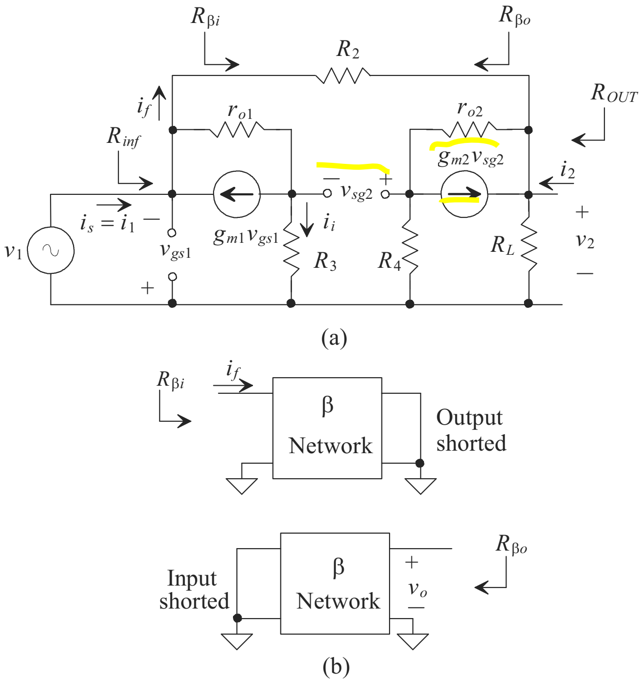

Figure 31.20 (a) Closed-loop small-signal model of Fig. 31.19 and (b) method for determining the feedback network loading.

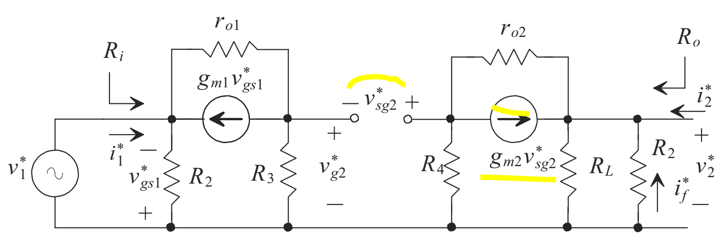

Figure 31.21 Open-loop small-signal model of Fig. 31.19.

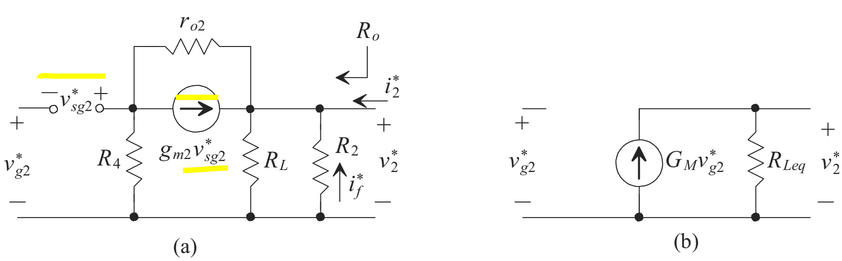

Figure 31.22 (a) Solving a portion of Fig. 31.21, including the drain-to-source resistance, and (b) the equivalent transconductance model.

The value of R_{Leq} can easily be found as

$$R_{Leq} = R_L || R_2 || R_{inD2} \tag{31.49}$$

where R_{inD2} is the resistance seen looking into the drain of M2. From Ch. 20, we know that this resistance is

$$R_{inD2} = [(1 + g_{m2}R_4)r_{o2} + R_4] \tag{31.50}$$

The value of G_M is the short-circuit transconductance and is defined as

$$G_M = \frac{i_o^*}{v_{g2}^*}\ (R_{Leq} = 0) \tag{31.51}$$

which means that the effective transconductance can be found by shorting the equivalent load resistance, in this case $R_L \| R_2$, and finding the gain from the short-circuit current to the input voltage. As seen in Fig. 31.23, the equations used to find G_M are

$$i_o^* = g_{m2}v_{sg2}^* + \frac{v_{s2}^*}{r_{o2}} \tag{31.52}$$

$$v_{s2}^* = -i_o^* R_4 \tag{31.53}$$

$$v_{s2}^* = v_{sg2}^* + v_{g2}^* \tag{31.54}$$

and solving Eqs. (31.52) - (31.54) yields

$$G_M = \frac{i_o^*}{v_{g2}^*} = \frac{-g_{m2}}{1 + g_{m2}R_4 + \frac{R_4}{r_{o2}}} \tag{31.55}$$

the gain, $\frac{v_2^*}{v_{g2}^*}$, becomes

$$\frac{v_2^*}{v_{g2}^*} = \frac{-g_{m2}(R_L || R_2 || [(1 + g_{m2}R_4)r_{o2} + R_4)]}{1 + g_{m2}R_4 + \frac{R_4}{r_{o2}}} \tag{31.56}$$

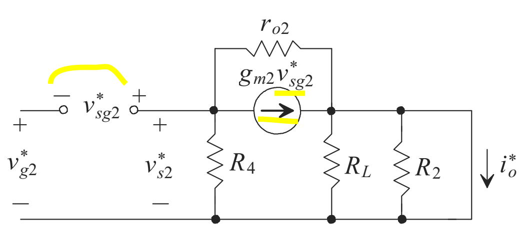

Figure 31.23 Circuit used to determine the equivalent transconductance.

Referring back to Eq. (31.47), the second factor, $\frac{v_{g2}^*}{v_1^*}$, can be found by analyzing Fig. 31.21 as

Since the output and the feedback are connected to two separate terminals of the output device, the output variable is a current, sampling i_o. The small-signal model for this circuit is shown in Fig. 31.31 with the open-loop, small-signal model shown in Fig. 31.32. Since the output sampling is a current, loading of the β network will be slightly different from that of the series-shunt example. The input utilizes series mixing; therefore the loading of the β network on the output will be identical to the series-shunt example discussed previously ($R_{\beta o} = R_1 + R_2$). However, since the output sampling is series, the equivalent resistance, $R_{\beta i}$, will be the resistance seen looking into the β network from the input, with the output device taken "out-of-socket" and $R_{\beta i} = R_2 + R_5$.

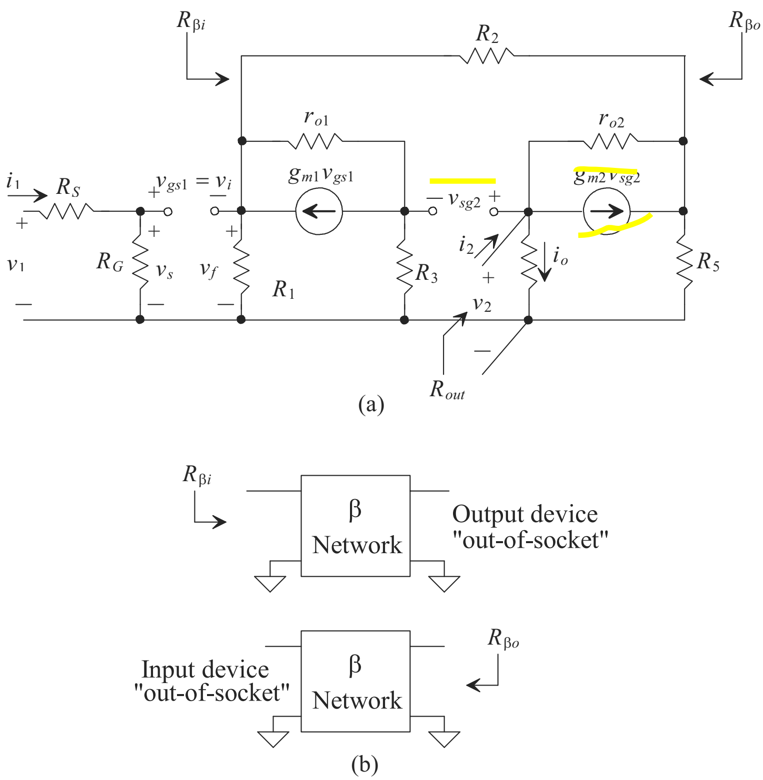

Figure 31.31 (a) Closed-loop small-signal model of Fig. 31.30 and (b) method for determining feedback loading.

Once the open-loop model has been constructed, A_{OL} can be calculated as

$$A_{OL} = \frac{i_o^*}{v_s^*} = \frac{i_o^*}{v_{g2}^*} \cdot \frac{v_{g2}^*}{v_s^*} \tag{31.80}$$

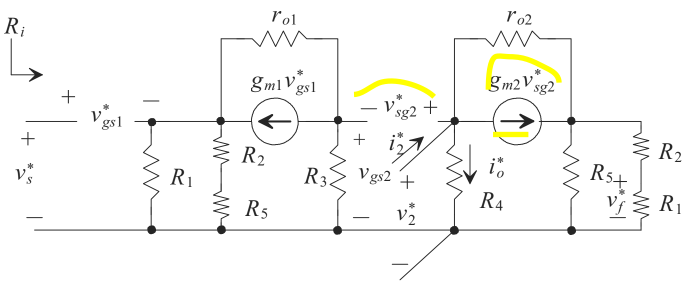

Figure 31.32 Open-loop small-signal model of Fig. 31.30.

where the term, $\frac{i_o^*}{v_{g2}^*}$, can be determined by using straightforward circuit analysis to solve $\frac{v_2^*}{v_{g2}^*}$ and then dividing the result by R_4 ,

$$\frac{i_o^*}{v_{g2}^*} = \frac{g_{m2}}{1 + g_{m2}R_4 + \frac{R_4 + R_5||(R_2+R_1)}{r_{o2}}} \tag{31.81}$$

The term, $\frac{v_{g2}^*}{v_s^*}$, is found by using the G_M method presented in the previous section on shunt-shunt feedback and is

$$\frac{v_{g2}}{v_s^*} = \frac{-g_{m1}(R_3||[(1+g_{m1}R_A)r_{o1} + R_A)]}{1 + g_{m1}R_A + \frac{R_A}{r_{o1}}} \text{ mhos} \tag{31.82}$$

where $R_A = R_1||(R_2+R_5)$. The feedback factor, β, is

$$\beta = \frac{v_f^*}{i_o^*} \approx \frac{-R_5R_1}{R_5 + R_1 + R_2}\Omega \tag{31.83}$$

And the closed-loop gain is simply

$$A_{CL} = \frac{i_o}{v_s} = \frac{A_{OL}}{1 + A_{OL}\beta} \text{mhos} \tag{31.84}$$

The value of R_i is obviously infinite, resulting in an identical value of R_{inf}. Therefore, $R_{in} = R_{inf}||R_G = R_G$.

Calculating R_o for a series output requires some explanation. Examine Fig. 31.33. The value of R_o is the value seen looking in series with the load resistor. In this case, the value of R_o becomes

$$R_o = R_4 + \frac{\frac{R_B}{r_{o2}} + 1}{\frac{1}{r_{o2}} + g_{m2}} \approx R_4 + \frac{1}{g_{m2}} \tag{31.85}$$

where $R_B = R_5||(R_1 + R_2)$ and the closed-loop value becomes

$$R_{of} = R_o(1 + A_{OL}\beta) \tag{31.86}$$

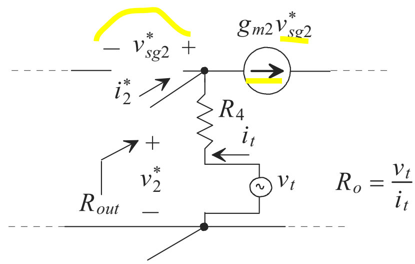

Figure 31.33 Calculation of the output impedance for the circuit in Fig. 31.30.

Notice, however, that R_{of} is not the same as R_{out}, in this case. Typically, R_{out} is designated as the resistance in parallel with the load. Taking the resistance in series with the load is not a practical specification. Therefore, the resistance R_{out} can be described as seen in Fig. 31.34. In part (a), it can be seen that $R_{of} = R_o(1+A_{OL}\beta)$ and that $R'_{of} = R_{of} - R_4$. If we want to find a value for R_{out}, using Fig. 31.34b, R_{out} is simply

$$R_{out} = R_4 || R'_{of} = R_4 || (R_{of} - R_4) \tag{31.87}$$

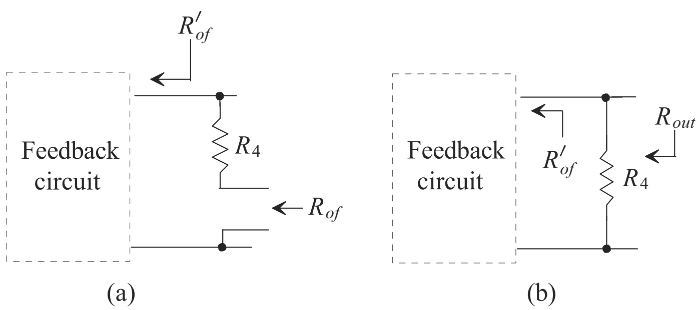

Figure 31.34 Determining the output resistance of a series sampling circuit.

31.7 The Current Amplifier (Shunt-Series Feedback)

The last feedback topology to be discussed is the shunt-series feedback amplifier, also known as a current amplifier. As can be expected, both A_{OL} and β have units of I/I, and we can expect the input impedance to be very low and the output impedance very high. Figure 31.35 illustrates the ideal shunt-series amplifier with open-loop values included. Based on past derivations, we can expect that

$$R_{inf} = \frac{R_i}{(1 + A_{OL}\beta)} \tag{31.88}$$

and R_{of} to be

$$R_{of} = R_o(1 + A_{OL}\beta) \tag{31.89}$$

The derivations of this topology will be left to the reader in the Problems section.