Chapter

31

Feedback Amplifiers

Back in Sec. 30.2 we compared op-amps in series-shunt and shunt-shunt feedback topologies without going into the details of what these terms mean. The goals of this chapter are to further develop feedback theory and its application to integrated circuit design by discussing the design of feedback amplifiers.

Feedback, in general, is a very powerful concept that has numerous applications. Examples of feedback systems abound in everyday life. For example, the thermostat on an air conditioner unit uses feedback to maintain a comfortable temperature within a room. Similarly, our bodies use feedback to increase antibodies to fight off an infection. By definition, feedback is the process of combining the output of a system with its input. In the air conditioner example, the temperature of the room is considered to be the output of the system, and the temperature set on the thermostat, the system input. The thermostat subtracts the value of the room temperature from the input value. If the temperature in the room is higher than the temperature set on the thermostat, the air conditioner turns on until the temperature in the room is less than or equal to the desired set value.

Thus, it can be said that feedback stabilizes a system. However, not all types of feedback have this property. There are two types of feedback: positive and negative. Negative feedback stabilizes, while positive feedback has the opposite effect. A good example of positive feedback is when a microphone is held too closely to the speaker of a public address (PA) system. More than likely, most people have heard the loud ringing that occurs. The positive feedback causes the system to become unstable, and the undesired effect is clearly recognizable.

Positive feedback occurs when the system output is added to the system input, whereas negative feedback occurs when the system output is subtracted from its input. For our purposes in this chapter, only negative feedback will be considered. While positive feedback can be useful if controlled, its applications are discussed elsewhere in the book. We will, however, discuss methods for minimizing the effects of positive feedback.

31.1 The Feedback Equation

Consider the feedback system shown in Fig. 31.1. The variables used in this diagram are labeled as an x value because they may be either a voltage or a current. The input signal, x_s, and the output signal x_o, may be either a voltage or a current. However, the feedback signal, x_f, must be the same type of signal as the input signal. If the input signal is a voltage, then the feedback signal must also be a voltage.

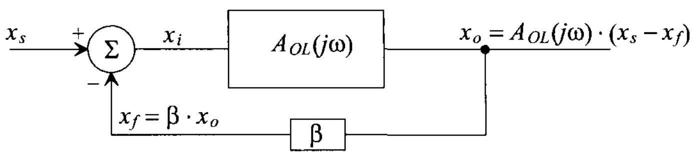


Figure 31.1 Basic block diagram to illustrate the feedback concept.

The block A_{OL} represents an amplifier's open-loop gain and is frequency dependent. The input to the amplifier, x_i, is the difference in the input source signal and the feedback signal, or

$$x_i = x_s - x_f \tag{31.1}$$

and the output is given by

$$x_o = A_{OL}(j\omega) \cdot (x_s - x_f) \tag{31.2}$$

In the following discussion, we will assume that all components of the system are ideal, meaning that the feedback network, β, will not load the amplifier. The specifications for the system can then be defined as follows:

$$A_{OL} = \frac{x_o}{x_i} \tag{31.3}$$

The *feedback factor*, β, is defined as

$$\beta = \frac{x_f}{x_o} \tag{31.4}$$

and the closed-loop gain, A_{CL}, is

$$A_{CL} = \frac{x_o}{x_s} \tag{31.5}$$

Realizing that

$$x_f = \beta \cdot x_o \tag{31.6}$$

and plugging Eq. (31.6) into (31.2) and solving for the closed-loop gain, we find that A_{CL} becomes

$$A_{CL} = \frac{x_o}{x_s} = \frac{A_{OL}}{1 + A_{OL}\beta} \tag{31.7}$$

where the frequency dependence of A_{OL} has not been shown.

Note the dependence of the A_{CL} on the value of A_{OL}. If the value of A_{OL} becomes large (A_{OL} approaches infinity), then the value of A_{CL} approaches $\frac{1}{\beta}$. This illustrates the need for having a high-gain amplifier. If A_{OL} is very large, then the closed-loop gain becomes highly dependent on the feedback components.

The term $A_{OL}\beta$ is often referred to as the *loop gain* and will be used later sections to determine overall amplifier stability (see, also, Sec. 24.1)

31.2 Properties of Negative Feedback on Amplifier Design

When used in the design of amplifiers, feedback can provide a number of advantages. These advantages include desensitizing the gain to process parameter variation, reducing nonlinear distortion, reducing the effects of noise, extending the useful bandwidth of the amplifier, and controlling the input and output impedance levels.

31.2.1 Gain Desensitivity

Since the value of the open-loop gain, A_{OL}, is large, its value may change significantly with temperature, mismatch of devices, and other parameter variations. However, negative feedback desensitizes the closed-loop gain from changes in the open-loop gain. The following derivation illustrates this property.

Differentiating both sides of Eq. (31.7) yields

$$\frac{dA_{CL}}{dA_{OL}} = \frac{1}{(1+A_{OL}\beta)^2} \text{ or } dA_{CL} = \frac{dA_{OL}}{(1+A_{OL}\beta)^2} \tag{31.8}$$

Dividing each side of Eq. (31.8) by the corresponding factors in Eq. (31.7),

$$\frac{dA_{CL}}{A_{CL}} = \frac{1}{(1+A_{OL}\beta)} \cdot \frac{dA_{OL}}{A_{OL}} \tag{31.9}$$

where $\frac{dA_{CL}}{A_{CL}}$ represents the fractional change in A_{CL} for a given fractional change in $\frac{dA_{OL}}{A_{OL}}$.

Using Eq. (31.9), if A_{OL} = 10,000 V/V (and assuming a voltage amplifier) and β = 1/10 V/V, it can be seen that if $\frac{dA_{OL}}{A_{OL}}$ = 10 %, then the change in the closed-loop gain, $\frac{dA_{CL}}{A_{CL}}$ = 0.01 %! This can easily be verified by using Eq. (31.7) for A_{OL}=10,000 and 9,000, keeping β = 1/10 V/V and solving for A_{CL} for each case. The resulting difference proves the immunity of the closed-loop gain to changes in A_{OL}.

31.2.2 Bandwidth Extension

Negative feedback also increases the usable bandwidth of an amplifier. Assume that an amplifier has the frequency response given by

$$A_{OLH}(s) = \frac{A_{OL} \cdot \omega_H}{s + \omega_H} = A_{OL}\frac{1}{\frac{s}{\omega_H} + 1} \tag{31.10}$$

$A_{OLH}(s)$ is simply the frequency dependent version of A_{OL} and is approximated using a first-order pole at ω_H. Plugging Eq. (31.10) into Eq. (31.7) yields the high-frequency dependent closed-loop version of Eq. (31.7)

$$A_{CLH}(s) = \frac{A_{OLH}(s)}{1 + A_{OLH}(s) \cdot \beta} = \frac{A_{OL} \cdot \omega_H}{s + \omega_H \cdot (1 + A_{OL}\beta)} \tag{31.11}$$

which can be rewritten as

$$A_{CLH}(s) = \frac{A_{OL}}{(1 + A_{OL}\beta)} \cdot \frac{1}{\frac{s}{\omega_H(1+A_{OL}\beta)} + 1} = \frac{A_{OL}}{(1 + A_{OL}\beta)} \cdot \frac{1}{\frac{s}{\omega_{HF}} + 1} \tag{31.12}$$

Note that the resulting equation is composed of two parts. The first factor should be recognized as Eq. (31.7) and is simply the expression for the closed-loop gain at midband frequencies. The second factor is the frequency dependent term. It is interesting to observe that the original –3 dB frequency ω_H in Eq. (31.10) is now multiplied by $(1+A_{OL}\beta)$. Figure 31.2 illustrates the bandwidth extension (at the cost of gain) from using negative feedback. The original open-loop frequency response is drawn in the solid line, while the closed-loop frequency response is illustrated in the dashed line. The closed-loop response shows a decrease in the gain and at the same time an increase in bandwidth.

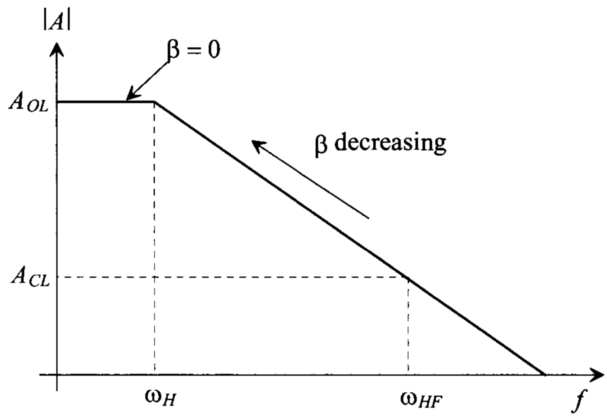


Figure 31.2 Extension of the high-frequency pole by using feedback.

If one were to decrease the value of β, from Eq. (31.11), it can be seen that two interesting effects occur. First, the closed-loop value of the gain will increase, since the first factor in Eq. (31.12) increases with a decrease in β. Second, the frequency response decreases since ω_H is multiplied by $(1 + A_{OL}\beta)$ and β is decreasing. As β decreases towards 0 the overall curve follows the original open-loop curve. It should be seen that when using feedback one trades gain for bandwidth.

The same analysis can be applied to the low-frequency response of an amplifier. Equation (31.13) approximates the low-frequency response of an amplifier with a single first-order low-frequency pole

$$A_{OLL}(s) = A_{OL} \cdot \frac{s}{s + \omega_L} \tag{31.13}$$

If we plug Eq. (31.13) into Eq. (31.7), the closed-loop low-frequency response becomes

$$A_{CLL}(s) = \frac{A_{OLL}(s)}{1 + A_{OLL}(s)\beta} = \frac{A_{OL}\frac{s}{s+\omega_L}}{1 + A_{OL}\frac{s}{s+\omega_L}\beta} = \frac{A_{OL}}{1 + A_{OL}\beta} \cdot \frac{s}{s + \frac{\omega_L}{1+A_{OL}\beta}} \tag{31.14}$$

The result in Eq. (31.14) is comprised of the standard closed-loop gain at midband and the frequency dependent term. Note however, that with feedback, the original low-frequency pole, ω_L, in Eq. (31.13) is now divided by $1+A_{OL}\beta$ in Eq. (31.14). Figure 31.3 illustrates the effect of the feedback on the low-frequency response of an amplifier.

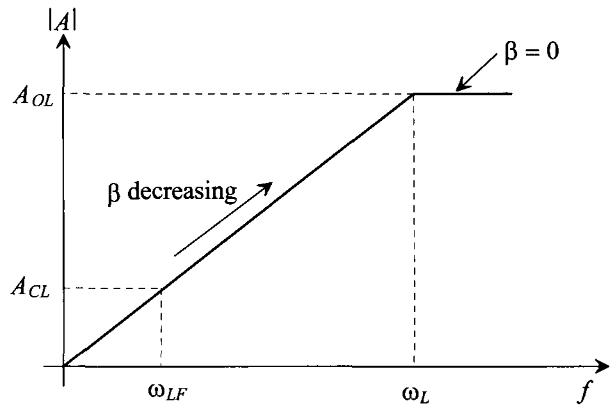


Figure 31.3 Extension of the low-frequency bandwidth by using feedback.

31.2.3 Reduction in Nonlinear Distortion

Negative feedback can also improve the nonlinear behavior of an amplifier. Examine the transfer curve of the voltage amplifier shown in Fig. 31.4 without feedback. Ideally, the amplifier should have a straight line from $-2\ \text{V} < V_{in} < 2\ \text{V}$. However, nonlinear behavior in the amplifier causes a different slope to occur when $V_{in} > 1\text{V}$ and $V_{in} < -1\ \text{V}$. Note that the output voltage is limited by the value of the power supply (+ 15 V).

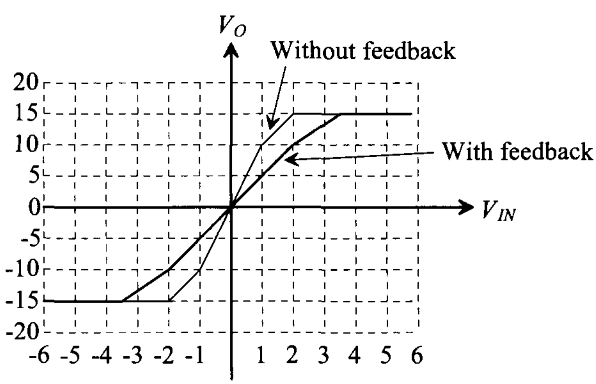


Figure 31.4 Using feedback to improve amplifier linearity.

Since the gain of the amplifier can be defined as the slope of the transfer curve, we can deduce that

$$A_{OL} = \frac{\Delta V_O}{\Delta V_{IN}} = 10\,\text{V/V} \quad \text{for} -10\text{ V} \le V_O \le 10\text{ V} \tag{31.15}$$

and

$$A_{OL} = \frac{\Delta V_O}{\Delta V_{IN}} = 5\text{ V/V} \quad \text{for } V_O > 10 \text{ and } V_O < -10\text{ V} \tag{31.16}$$

Now suppose that feedback is applied around the amplifier with $\beta = 0.1$ V/V and the transfer curve is redrawn. The gain of the amplifier with feedback becomes

$$A_{CL} = \frac{A_{OL}}{1 + A_{OL}\beta} = \frac{10}{1 + 10(0.1)} = 5\text{ V/V} \quad \text{for} -10\text{ V} < V_O < 10\text{ V} \tag{31.17}$$

and

$$A_{CL} = \frac{5}{1 + 5(0.1)} = \frac{10}{3} \quad V_O > 10\text{ V and } V_O < -10\text{ V} \tag{31.18}$$

The resulting transfer curve is seen in Fig. 31.4. Note that with feedback, the overall transfer curve is much more linear, resulting in an amplifier that has less nonlinear distortion than the original amplifier shown in Fig. 31.4.

31.2.4 Input and Output Impedance Control

The input and output impedances of an amplifier can also be controlled using negative feedback. As seen in Fig. 31.5, R_i is the small-signal input impedance looking into an amplifier without feedback, and R_{inf} is the input impedance with feedback applied. Similarly, R_{of} and R_o are the output impedances with and without feedback, respectively. Feedback allows us either to increase or decrease both R_{inf} and R_{of} by a factor of $(1 + A_{OL}\beta)$. Although an example of this property is difficult to derive in a general way, proofs of this property will be illustrated with actual circuit examples in a later section. For the time being, it will be sufficient to summarize the impedance control properties of feedback with Table 31.1. Note that the closed-loop impedances are completely dependent on the type of variable that is used (voltage or current) at the input and the output.

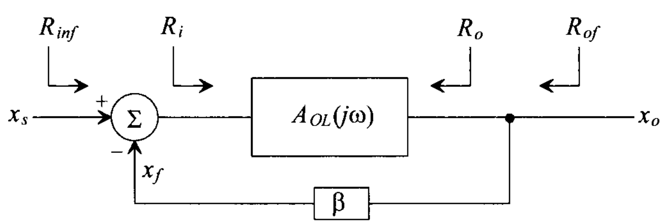


Figure 31.5 Determining the input and output impedances with and without feedback.

If the input variable, x_s, is a voltage, the closed-loop input resistance, R_{inf}, is equal to the open-loop value of the input resistance, R_{in}, multiplied by the value of $1 + A_{OL}\beta$. If the input variable is a current, R_{inf} is divided by the same factor. Alternatively, if the output variable, x_o, is a voltage, then the closed-loop output impedance is the open-loop output impedance divided by $(1 + A_{OL}\beta)$. And if the output variable is a current, the open-loop output impedance is multiplied by the same factor.

Obviously, this concept is a powerful one. We can adjust both input and output impedances in a larger or smaller direction simply by choosing the type of feedback used in the circuit.

Table 31.1 Summary of impedances with feedback.

Input Variable, x_s	Output Variable, x_o	R_{inf}	R_{of}
V	V	$R_i \cdot (1 + A_{OL}\beta)$	$R_o/(1 + A_{OL}\beta)$
V	I	$R_i \cdot (1 + A_{OL}\beta)$	$R_o \cdot (1 + A_{OL}\beta)$
I	V	$R_i/(1 + A_{OL}\beta)$	$R_o/(1 + A_{OL}\beta)$
I	I	$R_i/(1 + A_{OL}\beta)$	$R_o \cdot (1 + A_{OL}\beta)$

31.3 Recognizing Feedback Topologies

Examine the general single-loop feedback circuit in Fig. 31.6. Remember that x_s, x_i, and x_f must all either be currents or all be voltages, and that the output variable, x_o may be either a voltage or a current. As a result of these restrictions, a total of four types of feedback are possible (Table 31.2).

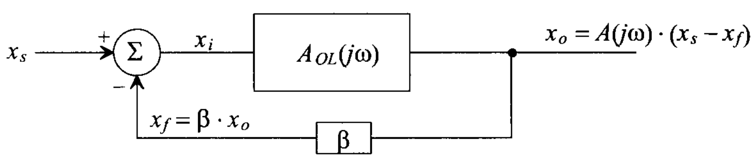


Figure 31.6 Basic block diagram to illustrate the feedback concept.

The input summation is often referred to as input mixing. If the input variables, x_s, x_i, and x_f can be written as voltages, the mixing is referred to as *series* or voltage mixing. If the input variables can be written as currents, it is referred to as *shunt* or current mixing. The type of variable used at the output determines the second term, known as sampling. If the output variable is a voltage, the sampling is then referred to as *shunt* or voltage sampling. If the output variable is a current, the sampling is known as *series* or current sampling.

Table 31.2 Feedback type as a function of system variables.

x_s, x_i, x_f	x_o	Feedback Type (mixing-sampling)
Voltage	Voltage	Series-shunt
Voltage	Current	Series-series
Current	Current	Shunt-series
Current	Voltage	Shunt-shunt

In the analysis of feedback amplifiers, the following terminology will apply. The term *basic amplifier* will correspond to the gain circuit, A_{OL}, in the block level feedback diagrams. When determining the value of A_{OL}, it is very important to determine the loading due to the β network and any source and load resistance. The β network is also known as the feedback network. The overall circuit that includes both A_{OL} and β will be referred to as the *feedback amplifier*.

31.3.1 Input Mixing

Figures 31.7a and b illustrate the inputs to two generic feedback amplifiers. The basic amplifier in 31.7a is used in a series mixing configuration since the equation, $x_i = x_s - x_f$ can be written in terms of the voltages. Note also that the basic amplifier and the β network are in series with each other. In Fig. 31.7b, however, the basic amplifier is used in a shunt mixing configuration. Here, the summation at the input can only be written in terms of currents such that $i_i = i_s - i_f$. The feedback circuit and the basic amplifier circuit are both in parallel, or it can be said that the feedback shunts the basic amplifier.

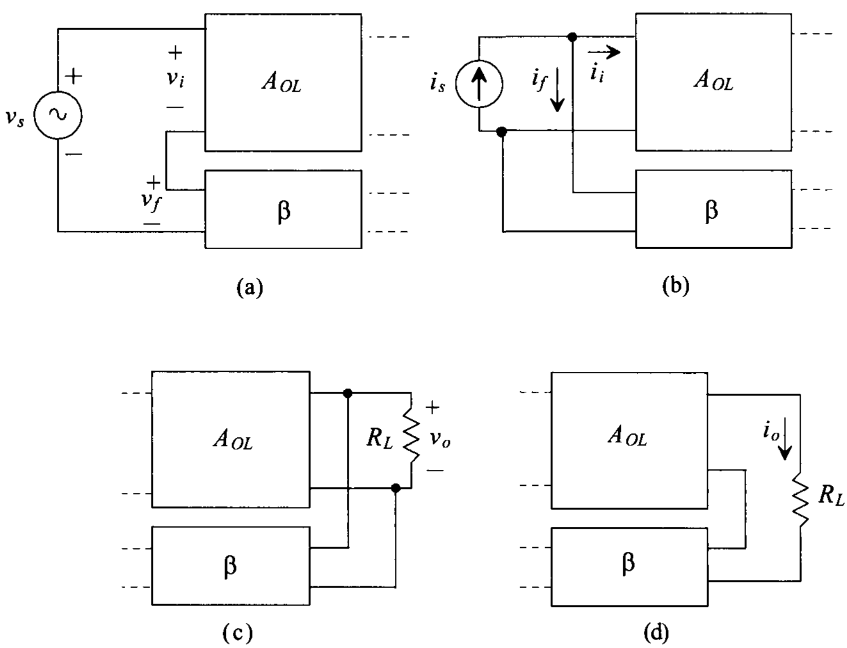


Figure 31.7 Generic block diagrams for input circuits (a) series mixing and (b) shunt mixing and output circuits, (c) shunt sampling, and (d) series sampling.

31.3.2 Output Sampling

Determining the type of variable used at the output is not an obvious endeavor. However, a general model followed by several examples will help clarify the proper procedure. Figures 31.7c and d show the outputs of two generic feedback amplifiers. Shunt sampling is shown in Fig. 31.7c. Note that the feedback circuit is shunting, or in parallel with the basic amplifier circuit. The output variable is considered a voltage because the feedback circuit "senses" or samples the voltage across R_L. Series sampling is seen in Fig. 31.7d. Here, the feedback network is in series with the basic amplifier circuit. The output variable is a current since the feedback circuit now senses the current through the load resistor. Two rules that help distinguish between the two types of output sampling are:

Rule 1: If one terminal of the output active device (drain or source) is driving the load and the other terminal is attached to the feedback network, the output sampling is series.

Rule 2: If the load and the feedback network are connected to the same node, the output sampling is shunt.

31.3.3 The Feedback Network

When analyzing feedback circuits, it would be wise first to recognize the basic amplifier circuit, A_{OL} , and the β network by tracing the small-signal path from the input to the output and back through the feedback. The path from the input through the A_{OL} circuit to the output will be denoted as the forward path and the path from the output back through the β network, as the feedback path. Determining which path the signal takes is not obvious when there are several devices from which to choose. Some rules that will aid in the analysis are as follows:

- The *forward path* through the basic amplifier circuit will always take the path that has the highest gain.
- The AC small-signal will always enter either a gate or a source and will always exit either a drain or a source. The gain from drain to source is very small (at least for linear applications) and is considered to be negligible for most of our applications. A small-signal will never exit through the gate.
- The feedback signal must subtract from the input signal. To ensure that this is the case, one must count the number of inversions around the loop. Every time a signal crosses a gate-to-drain junction (common source amplification), an inversion occurs. Examples of this will be discussed next.

Examine Fig. 31.8a. Note that the signal path in the forward direction progresses through the path with the highest gain—from the gate of M1 to the drain of M1, into the gate of M2 and out the drain of M2. The feedback path consists of the resistors R_1 and R_2. The feedback variable consists of the voltage that appears across R_1.

An Important Assumption

One might attempt to draw the forward path from the gate of M1 to the source of M1, through the feedback resistor, R_2 , and to the output. In actuality, this would be a legitimate forward path because in some cases the feedback network is actually bidirectional. However, the signal will have very little gain since the gain from gate to source of any MOS device is a maximum of one (common drain amplification) and the gain from v_f to v_2 in the forward direction will be less than one because of the voltage divider relationship between R_5 and R_2 . *Thus, for all of the analysis performed in this*

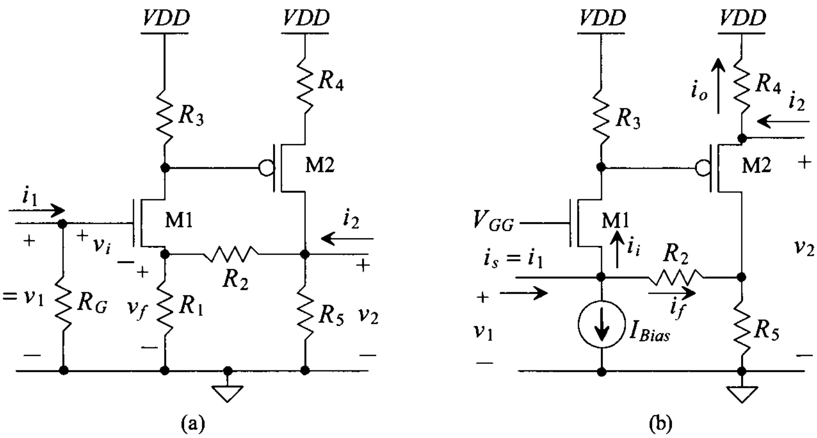


Figure 31.8 Feedback topologies (a) series-shunt and (b) shunt-series.

chapter, the forward path through the amplifier circuit dominates the expression for the total forward gain from input to output, while the forward path through the feedback is assumed to be negligible. This important assumption greatly simplifies the analysis and is a good one, since the amplifier will have very poor performance if there is not a good deal of gain through the basic amplifier. (Ideally, the value of A_{OL} is infinite.) If an active device is used in the feedback network, the forward gain through the β network is minimized.

Figure 31.8b has the same basic transistor topology, except it uses current as its input. The forward path consists of the source of M1, out the drain of M1 into the gate of M2 and out the source of M2. The feedback path consists of R_5 and R_2, and the feedback variable is the current, i_f. Note that the forward path has the highest possible gain path, since M1 acts as a common gate amplifier. The forward path could progress from the input through R_2 and into the drain of M2. However, since the small-signal gain from the drain to the source of M2 is extremely small, it will be neglected.

Counting Inversions Around the Loop

It is also necessary to check the circuit to ensure that the feedback is indeed negative. Counting the number of inversions around the loop in Fig. 31.8a may initially give the wrong impression, because the number of gate-to-drain junctions encountered by the signal around the loop is two. This implies positive feedback since the feedback variable must be inverted with respect to the input signal. However, the final mixing between v_1 and v_f will provide the additional negative behavior needed to ensure the proper feedback. Notice the relationship that exists between v_1 and v_f. Since $v_i = v_1 - v_f$ the voltage v_f has a subtractive effect on v_i. Thus, a positive change in the signal entering the gate of M1 will result in a positive change in the source of M1 and a smaller, stable voltage v_i. If v_f had been negative with respect to v_1, then v_i would be, $v_i = v_s + v_f$ and positive feedback would occur.

Now examine Fig. 31.8b. Again, if we count the number of inversions around the loop from input back to the mixing variable, i_f, the number of inversions is odd, which is exactly what is needed, since KCL tells us that $i_i = i_1 - i_f$. The direction of the feedback current from the output is opposite the direction of i_f; however, the signal is inverted, making it equivalent to i_f, as illustrated from Fig. 31.9.

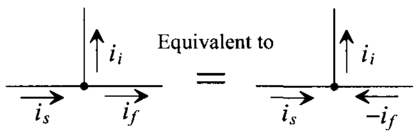


Figure 31.9 Shunt mixing illustrated with an odd number of inversions.

Examples of Recognizing Feedback Topologies

Now that the method for recognizing feedback has been discussed, examples will be presented to solidify the concepts. Again examine Fig. 31.8. Here, two types of feedback are illustrated. Transferring the block diagram concepts to transistor-level circuits can be a difficult task. However, several other rules can be applied which will reveal the type of sampling used in the circuit. The term *input active device* is used to denote the transistor that is being driven by the input source. The term *output active device* is used to denote the transistor used to drive the load.

In Fig. 31.8a, the input variables can only be written in terms of voltages such that $v_i = v_s - v_f$. We could attempt to sum currents at the node on the gate. However, the current flowing into the resistor, R_G, will not be the result of any feedback, thus making current mixing impossible. The output sampling is of the shunt type since the β network and the basic amplifier are in parallel (the β network is connected to the same node as the output) and v_o is the voltage being sampled. This amplifier employs series mixing and shunt sampling, and is referred to as a series-shunt feedback amplifier configuration.

Figure 31.8b illustrates shunt-series feedback. The DC current source is viewed as an AC open circuit, and the DC voltage source, V_{GG}, is considered as an AC short circuit. Note how the input variables can only be written in terms of currents forming the expression, $i_i = i_s - i_f$. Series sampling is used at the output since the feedback network is in series with the basic amplifier. This amplifier configuration also follows rule 2, mentioned previously; thus, the proper small-signal output variable is the current, i_o. Note that the direction of the small-signal output current, i_o, is consistent with the direction of the small-signal model, since *VDD* is considered a small-signal ground.

The variables v_1, i_1, v_2, and i_2 may or may not correspond to the defined input and output variables of the feedback circuit. For example, in Fig. 31.8b, the correct output variable is defined as i_o. However, the gain of the feedback amplifier can be ascertained in terms of v_1, i_1, v_2, and i_2, since $v_2 = i_o \cdot R_4$. Also note that v_1/i_1 and v_2/i_2 correspond to the input and output impedances, respectively, of the feedback amplifier.

Series-series feedback is illustrated in Fig. 31.10a. Although the input variables are written in terms of voltages, the output variable is considered to be a current since the

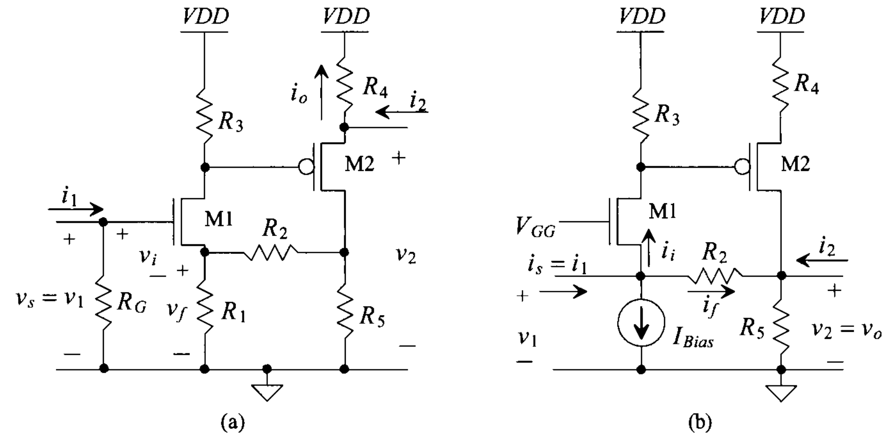


Figure 31.10 Feedback topologies: (a) series-series and (b) shunt-shunt.

feedback network is in series with the amplifier output. Note that the only difference between Figs. 31.10a and 31.8a is in placement of the output terminal. The same holds true for Figs. 31.10b and 31.8b. Lastly, Fig. 31.10b illustrates shunt-shunt feedback. Current summation occurs at the input which typifies shunt or current mixing, and the β network made up of R_5 and R_2 shunts the output.

31.3.4 Calculating Open-Loop Parameters

Once the feedback topology can be recognized, the analysis of the circuit can begin. We will perform two types of analysis: open-loop and closed-loop. Open-loop analysis entails approximately 80 percent of the circuit analysis required to solve a feedback problem. Open-loop values are then used to calculate the closed-loop values, so extreme care must be taken when analyzing the open-loop circuit. We will then "close the loop" and calculate the feedback characteristics of the overall feedback amplifier

The method of analysis for feedback amplifiers can be confusing if we do not distinguish between open- and closed-loop notation. We will denote the open-loop voltage and current variables by use of the * notation. The open-loop parameters include A_{OL}, R_i, R_o, and β and are defined as follows.

$$A_{OL} = \frac{x_o^*}{x_s^*} \text{ and is the open-loop gain of the basic amplifier} \tag{31.19}$$

$$R_i = \frac{v_1^*}{i_1^*} \text{ and is the open-loop input impedance} \tag{31.20}$$

$$R_o = \frac{v_2^*}{i_2^*} \text{ and is the open-loop output impedance} \tag{31.21}$$

$$\beta = \frac{x_f^*}{x_o^*} \text{ and is the gain through the feedback network} \tag{31.22}$$

In our discussion of general feedback principles in Sec. 31.1, it was assumed that the β network was ideal, meaning that the impedance of the β network did not load the amplifier circuit. However, in real-life applications, the β network can cause loading effects on both the input source and the output of the amplifier circuit. The type of feedback used will determine how the β network loading is calculated. As we progress in the discussion of each feedback topology, the method of determining the β network loading will be presented.

All variables in the following analysis are considered to be small-signal AC voltages or currents. We will perform the open-loop analysis using the following steps:

1. Replace input source with Norton or Thevenin equivalent circuit and calculate open-loop gain. The amplifier circuit will produce a different type of gain for each type of feedback. For example, the open-loop gain for an amplifier used in a series-shunt configuration will have units of V/V. This is a standard voltage amplifier since the output variable is a voltage and the input mixing sums voltages. An amplifier using series-series feedback will have a gain with units of I/V, since the output variable is now a current and the input mixing uses voltages. This type of amplifier is also known as a transconductance amplifier since the units of the gain, I/V, are equivalent to a conductance. Similarly, an amplifier used in the shunt-series configuration has a gain with units of I/I (a current amplifier), and an amplifier used in a shunt-shunt configuration will have a gain with units of V/I and is known as a transimpedance amplifier.

2. When calculating the open-loop gain of the circuit, consider the β network loading. An equivalent resistance, $R_{\beta i}$ and $R_{\beta o}$, as seen in Fig. 31.11 will represent the total input and output resistance of the β network. The method for calculating $R_{\beta i}$ involves the following steps.

   - If the output sampling is shunt (voltage), short the output node to ground.
   - If the output sampling is series (current), remove the device driving the output load as if it were taken "out-of-socket."
   - Calculate $R_{\beta i}$ as the impedance looking from the input into the β network.

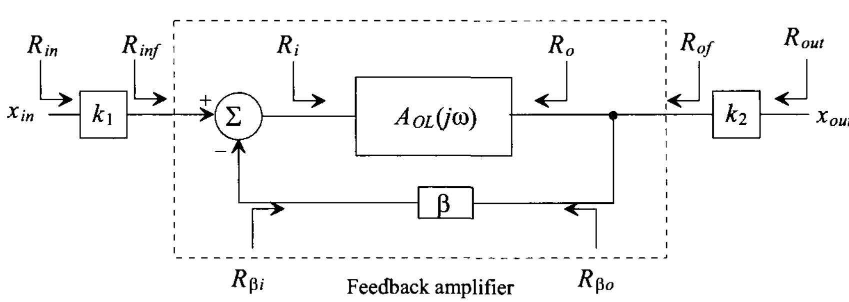


Figure 31.11 Block diagram of a generic feedback amplifier that distinguishes between open- and closed-loop impedances.

The method for calculating $R_{\beta o}$ involves similar methodology:

- If the input mixing is shunt (current), short the input node to ground.
- If the input mixing is series (voltage), remove the input active device as if it were taken "out-of-socket."
- Calculate $R_{\beta o}$ as the impedance looking from the output into the β network.

The method for determining the loading of the β network is rooted in basic two-port theory where each type of feedback configuration corresponds to one of the four basic two-port topologies. For further information regarding two-port theory, the reader is advised to consult [1]. One method used to easily remember the above rules is to remember that if mixing or sampling is shunt, then "short" the input or output, respectively. If the mixing or sampling is series, then "sever" the input or output device, respectively, by taking it "out-of-socket."

3. Calculate feedback factor, β, for the amplifier. The open-loop circuit is analyzed to determine the gain from the output back to the point where the feedback variable mixes with the input.
4. Determine the open-loop input and output impedances, R_i and R_o. These impedances can be calculated using standard circuit analysis techniques.

31.3.5 Calculating Closed-Loop Parameters

Once all the open-loop parameters are obtained, the closed-loop values are easily calculated. The closed-loop gain is

$$A_{CL} = \frac{x_o}{x_s} = \frac{A_{OL}}{1 + A_{OL} \cdot \beta} \tag{31.23}$$

for all four topologies. The closed-loop input impedances represent only the input and output impedances of the feedback amplifier. The input impedance from the input source, R_{in}, may correspond to R_{inf} if there is no gain from the source, x_{in}, to the input of the feedback amplifier, x_s. Similarly, the value of the output resistance, R_{out}, may or may not correspond to R_{of}, depending on the type of sampling used. Figure 31.11 illustrates this distinction. Calculating the input and output impedances of the feedback amplifier is dependent on the type of mixing and sampling circuits used, respectively.

$$\text{For series input mixing, } R_{inf} = R_i(1 + A_{OL}\beta) \tag{31.24}$$

$$\text{For shunt input mixing, } R_{inf} = \frac{R_i}{(1 + A_{OL}\beta)} \tag{31.25}$$

The value of the closed-loop output impedances will be as follows:

$$\text{For series output sampling, } R_{of} = R_o(1 + A_{OL}\beta) \tag{31.26}$$

$$\text{For shunt output sampling, } R_{of} = \frac{R_o}{(1 + A_{OL}\beta)} \tag{31.27}$$

Equations (31.24)-(31.27) will be derived with each specific topology. However, it is important to note the effectiveness of using feedback to control input and output impedances. Note that if series mixing or series sampling is used, the open-loop value of the input or output impedance is multiplied by $(1 + A_{OL}\beta)$; and if shunt mixing or shunt sampling is used, the open-loop values of impedances are divided by the same factor.

Now that the general methodology has been described, a detailed examination of the four feedback topologies will be presented. The analysis of the four basic feedback topologies begins at the discrete level and progresses into more complex integrated circuits throughout the section.

31.4 The Voltage Amp (Series-Shunt Feedback)

Consider the ideal voltage feedback amplifier shown in Fig. 31.12, with open-loop values, A_{OL}, β, R_i, and R_o already given. The basic amplifier is a voltage amplifier with gain given V/V. Since this is an ideal feedback amplifier, the β network does not load the basic amplifier, meaning that $R_{\beta o} = \infty$ and $R_{\beta i} = 0$.

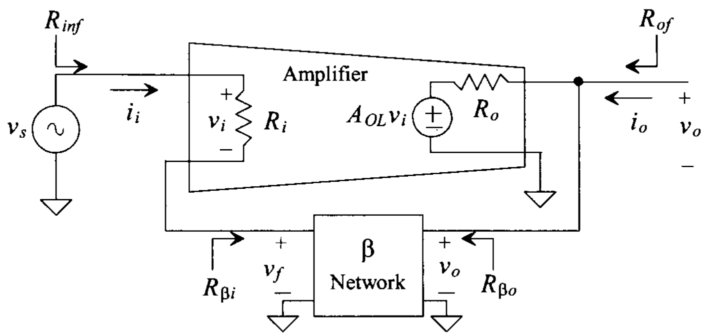


Figure 31.12 An ideal voltage amplifier (series-shunt).

To determine what type of feedback is present in the circuit, examine the input variables. Since input variables x_s, x_i, and x_f correspond to the voltages v_s, v_i, and v_f, such that the equation $v_i = v_s - v_f$ can be written the input mixing is series. The output mixing can be determined using the previous rules. Since the output of the amplifier, A_{OL}, and the β network are attached in parallel (both being connected to the load, R_L), the output mixing is considered to be shunt.

From Sec. 31.1, we already know that the closed-loop gain of the amplifier is

$$A_{CL} = \frac{v_o}{v_s} = \frac{A_{OL}}{1 + A_{OL} \cdot \beta} \tag{31.28}$$

Notice that as A_{OL} approaches infinity, Eq. (31.28) approximates to

$$A_{CL} \approx \frac{1}{\beta} \tag{31.29}$$

Equation (31.29) is important because it illustrates another powerful feedback concept. The entire gain of the feedback amplifier can be approximated as the inverse of β as the basic amplifier gain increases to higher and higher values.

We can calculate how the feedback affects the input impedance of the amplifier by applying a test voltage directly to the input of the feedback amp shown in Fig. 31.12, so that $v_s = v_{test}$ and $i_i = i_{test}$. Writing a voltage loop at the input and assuming that the feedback network does not load the amplifier gives

$$v_{test} = i_{test} \cdot R_i + \beta \cdot v_o = i_{test} \cdot R_i + \beta \cdot \frac{A_{OL}}{1 + A_{OL}\beta} \cdot v_{test} \tag{31.30}$$

Equation 31.30 simplifies to

$$R_{inf} = \frac{v_{test}}{i_{test}} = R_i \cdot (1 + A_{OL}\beta) \tag{31.31}$$

The gain of the open-loop amplifier was decreased by $1 + A_{OL}\beta$, while the input impedance, the impedance the source sees, was increased by $1 + A_{OL}\beta$.

Calculation of the output impedance proceeds in the same way as the calculation of the input resistance except that the test voltage is applied to the output of the amplifier with the input shorted to ground with $v_o = v_{test}$ and $i_o = i_{test}$. Writing a loop equation at the output gives

$$v_{test} = i_{test} \cdot R_o + A_{OL} \cdot v_i = i_{test} \cdot R_o + A_{OL} \cdot (-\beta \cdot v_o) \tag{31.32}$$

since the input is shorted to ground and $v_i = -v_f = -\beta \cdot v_o$. The output impedance is given by

$$R_{of} = \frac{v_{test}}{i_{test}} = \frac{R_o}{1 + A_{OL}\beta} \tag{31.33}$$

or the output resistance is reduced by $1 + A_{OL}\beta$. Ideally, a voltage amplifier has infinite input resistance and zero output resistance. Adding feedback to a voltage amplifier with finite input resistance and nonzero output resistance help make the amplifier closer to the ideal.

Now that the ideal series-shunt amplifier has been examined from a block level point of view, a discussion of the nonideal series-shunt feedback amplifier at the transistor level with loading effects will be considered. The amplifier seen in Fig. 31.13 was analyzed previously and was determined to use series-shunt feedback.

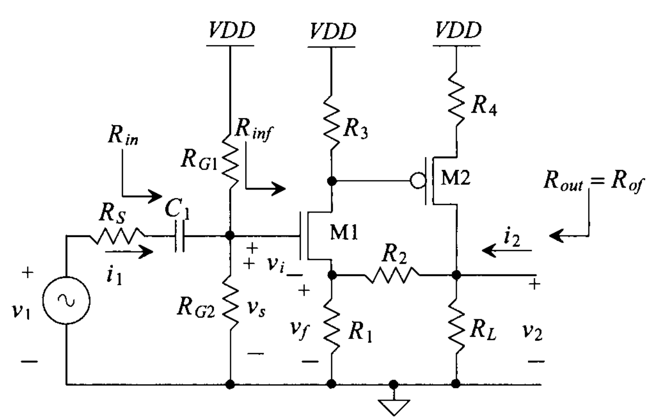


Figure 31.13 Transistor-level series-shunt feedback amplifier.

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.

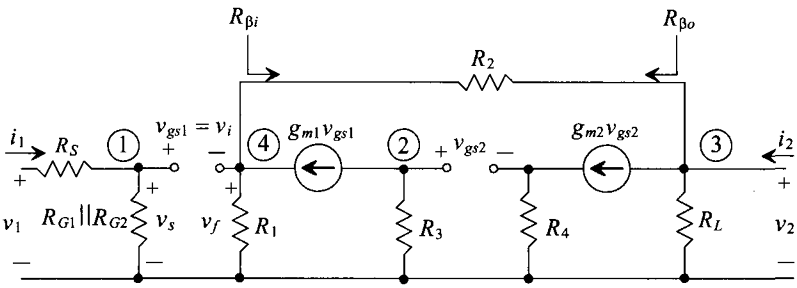


Figure 31.14 Closed-loop small-signal model of Fig. 31.13.

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

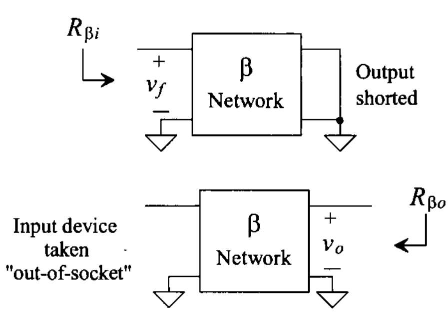


Figure 31.15 Determining the loading due to the feedback network for a series-shunt amplifier.

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.

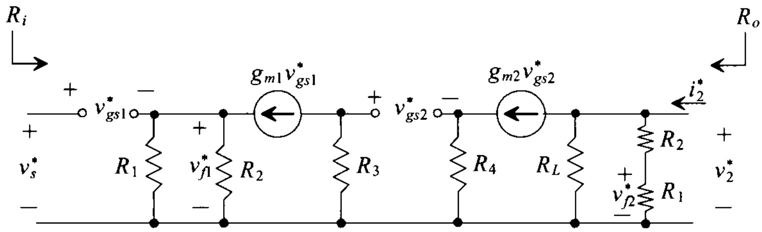


Figure 31.16 Open-loop small-signal model of Fig. 31.13.

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

$$A_{OL} = \frac{v^*_2}{v^*_s} = \left(\frac{v^*_2}{v^*_{gs2}}\right)\left(\frac{v^*_{gs2}}{v^*_{gs1}}\right)\left(\frac{v^*_{gs1}}{v^*_s}\right) = [-g_{m2}R_L||(R_2+R_1)]\left[-\frac{g_{m1}R_3}{1+g_{m2}R_4}\right]\left[\frac{1}{1+g_{m1}(R_1||R_2)}\right] \tag{31.35}$$

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 be obvious that the input resistance to the open-loop circuit is ∞. The output resistance, however, can be calculated shorting the gate of M1 to ground and applying a test voltage to the output. Since the input is grounded, $v_{g2} = 0$, and R_o is simply the parallel combination of resistances at the output (assuming that r_{o2} is very large).

$$R_o = \frac{v_o^*}{i_2^*} = R_L || (R_1 + R_2) \tag{31.37}$$

Once the open-loop values are calculated, the closed-loop values are easily attained. The closed-loop values are

$$A_{CL} = \frac{v_2}{v_s} = \frac{A_{OL}}{1 + A_{OL}\beta} \tag{31.38}$$

$$R_{inf} = \frac{v_s}{i_s} = R_i(1 + A_{OL}\beta) \tag{31.39}$$

$$R_{out} = R_{of} = \frac{v_2}{i_2} = \frac{R_o}{1 + A_{OL}\beta} \tag{31.40}$$

Analysis of this problem has not yet been completed. Notice that we initially neglected the source resistance and the biasing resistors, R_{G1} and R_{G2}, since they played no part in the feedback analysis of the amplifier. However, they will have an effect in the overall gain and need to be considered. The total gain of the entire circuit is

$$\frac{v_2}{v_1} = \frac{v_s}{v_1} \cdot \frac{v_2}{v_s} = \frac{R_{G1} || R_{G2}}{R_{G1} || R_{G2} + R_S} \cdot A_{CL} \tag{31.41}$$

and the value of R_{in} as seen by the signal source is

$$R_{in} = \frac{v_1}{i_1} = R_{G1} || R_{G2} \tag{31.42}$$

and the analysis is now complete. Notice that the value of R_{in} is not the same as R_{inf}.

Example 31.1

For the series-shunt circuit shown in Fig. 31.17a, draw the closed-loop small-signal model and identify the forward and feedback paths, draw the open-loop model, calculate A_{OL}, β, R_i, and R_o, and calculate the closed-loop parameters v_2/v_1, v_1/i_1, and v_2/i_2. You may assume that $r_o = \infty$ for both devices and that DC analysis has already been performed with $g_{m1} = g_{m2} = 1$ mA/V.

The closed-loop small-signal model can be seen in Fig. 31.17b with the forward and feedback paths drawn. The open-loop model with loading effects is seen in Fig. 31.17c. Since the output uses shunt sampling, the value of $R_{\beta i}$ is found by shorting the output node and looking at the load resistance resulting from the feedback network. In this case, $R_{\beta i}$ is equal to R_2. Since the input mixing is series, the input device, M1, is "severed," so that only $R_1 + R_2$ is seen looking into the feedback loop from the output. Thus, $R_{\beta o} = R_1 + R_2$.

The values of r_o are considered to be infinite, so the values of R_i and R_o can be found by inspection, $R_i = \infty$, and $R_o = R_1 + R_2 = 11$ kΩ. Notice that when solving the open-loop circuit (Fig. 31.17c) v_1 and R_G are not included since we are only interested in solving the feedback portion of the circuit.

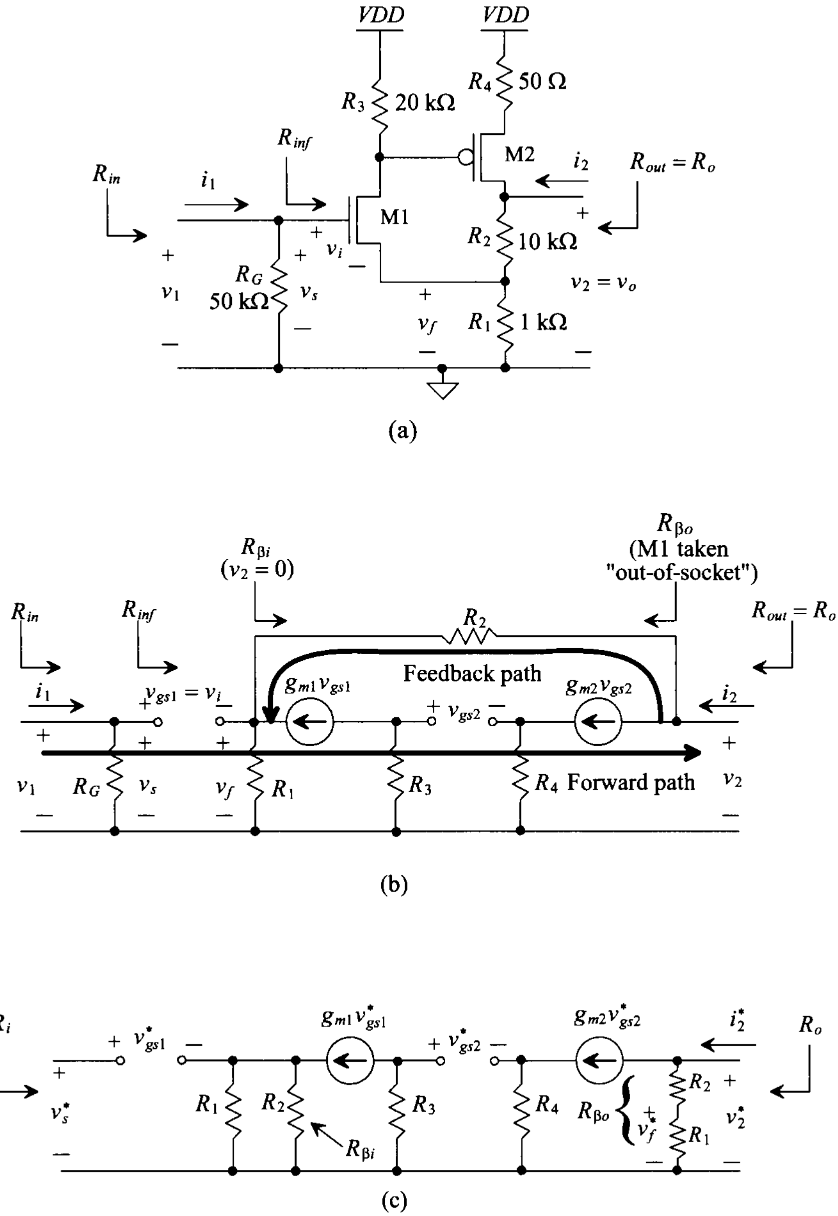


Figure 31.17 (a) Series-shunt circuit used in Ex. 31.1; (b) its closed-loop small-signal model; and (c) the resulting open-loop model.

The open-loop gain, A_{OL}, can be found simply as the gain of two common source amplifiers with source resistance,

$$A_{OL} = \frac{v_2^*}{v_s^*} = \frac{v_2^*}{v_{g2}^*} \cdot \frac{v_{g2}^*}{v_s^*} = \left[\frac{-g_{m2}(R_2 + R_1)}{1 + g_{m2}R_4}\right]\left[\frac{-g_{m1}R_3}{1 + g_{m1}(R_1 || R_2)}\right] = 109.8 \text{ V/V}$$

The value of β is always the gain from the output to the feedback variable. In this case, a simple voltage divider relationship exists such that

$$\beta = \frac{v_f^*}{v_2^*} = \frac{R_1}{R_1 + R_2} = 0.0909 \text{ V/V}$$

Again, notice that the product of $A_{OL}\beta$ is positive and unitless. Now that the open-loop values are calculated, the closed-loop values can be found by using Eqs. (31.38) - (31.40).

$$A_{CL} = \frac{v_2}{v_s} = \frac{A_{OL}}{1 + A_{OL}\beta} = \frac{109.8}{1 + (109.8 \cdot 0.0909)} = 10 \text{ V/V}$$

A simple check will verify that the solution is correct if $A_{CL} \approx \frac{1}{\beta}$, which is the case.

Notice in Fig. 31.17a and Fig. 31.17b that closed-loop output impedance, R_{of} is equal to the value of R_{out}. However, the value of the closed-loop input impedance, R_{inf}, is not equal to R_{in}, since the feedback amplifier itself excludes the input source and the gate resistor, R_G. The closed-loop values, R_{inf}, and R_{of}, are also easily attained, that is,

$$R_{inf} = \infty \text{ (since } R_i = \infty)$$

and

$$R_{of} = R_{out} = \frac{v_2}{i_2} = \frac{R_o}{(1 + A_{OL}\beta)} = \frac{11 \text{ k}\Omega}{10.98} = 1.002 \ \Omega$$

The last step involves finding R_{in} and v_2/v_1. The value of R_{in} can be found by examining Fig. 31.17a as,

$$R_{in} = \frac{v_1}{i_1} = R_G || R_{inf} = 50 \text{ k}\Omega$$

The overall gain, v_2/v_1, will be equal to the value of A_{CL} since the input voltage, v_1 is equal to v_s. If we had included a value for a source resistance, then an additional voltage divider relationship would have been needed to formulate v_2/v_1 as seen in Eq. (31.41)

$$\frac{v_2}{v_1} = A_{CL} = 10 \text{ V/V} \ \blacksquare$$

31.5 The Transimpedance Amp (Shunt-Shunt Feedback)

Shunt-shunt feedback mixes current at its input and samples voltage at the output. Consider the ideal shunt-shunt feedback amplifier shown with open-loop values, A_{OL}, β, R_i, and R_o, in Fig. 31.18. In the ideal case, $R_{\beta i}$ is infinite and $R_{\beta o}$ is zero. Notice that the basic amplifier and the β network are in parallel at both the input and the output. The β

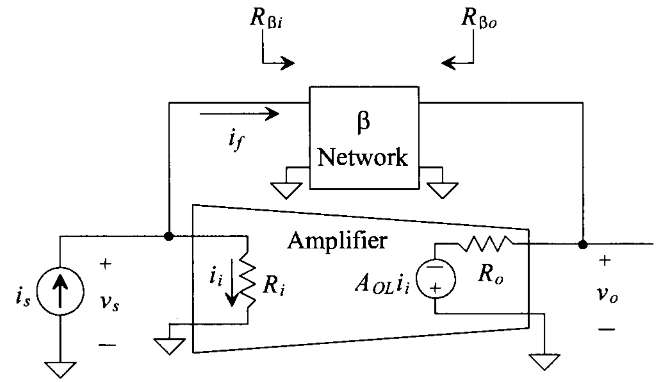


Figure 31.18 An ideal transimpedance (shunt-shunt) amplifier.

network shunts the basic amplifier; therefore, the input variable is a current. Looking into the output of the amplifier, we see that the feedback path is in parallel with or shunts the output signal. Therefore, the output signal is a voltage. Also note that since the input variable is a current and the output variable is a voltage, the basic amplifier has units of V/I, also known as a transimpedance amplifier. Since $A_{OL}\beta$ should always be unitless and positive, β will have units of I/V (mhos).

The closed-loop gain is

$$A_{CL} = \frac{v_o}{i_s} = \frac{A_{OL}}{1 + A_{OL}\beta} \tag{31.43}$$

Notice that the basic amplifier circuit is inverting, corresponding to an ideal op-amp circuit. Since A_{OL} is negative, then β must also be negative to ensure that negative feedback exists.

To calculate the input impedance of the transimpedance amplifier we apply a test current source to the input of the amplifier, such that $i_s = i_{test}$ and $v_s = v_{test}$. The test current and the input impedance are determined by

$$i_{test} = \frac{v_{test}}{R_i} + i_f = \frac{v_{test}}{R_i} + \beta v_o = \frac{v_{test}}{R_i} + \beta A_{OL} i_i = \frac{(1 + A_{OL}\beta)\cdot v_{test}}{R_i} \tag{31.44}$$

Assuming that the β network does not load the basic amplifier, we note that the closed-loop input impedance becomes

$$R_{inf} = \frac{v_{test}}{i_{test}} = \frac{R_i}{1 + A_{OL}\beta} \tag{31.45}$$

A similar analysis of the output impedance of the transimpedance amplifier shows that

$$R_{of} = \frac{R_o}{(1 + A_{OL}\beta)} \tag{31.46}$$

The ideal transimpedance amplifier has zero input resistance and zero output resistance. We can see from Eqs. (31.45) and (31.46) that the addition of the feedback helps to make the basic amplifier appear closer to the ideal.

Now examine the transistor level shunt-shunt feedback circuit shown in Fig. 31.19. The closed-loop and open-loop small-signal models are seen in Fig. 31.20 and Fig. 31.21, respectively. The gate is attached to AC ground since it is attached to a DC source and the DC current source I_{SS} is an AC open circuit. Interestingly, since the values of r_{o1} and r_{o2} have been included, there is a feedback path through the basic amplifier. However, this gain is very small compared to the feedback path through the β network and is typically assumed to be negligible.

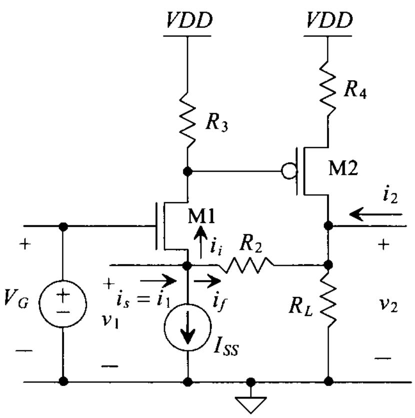


Figure 31.19 Shunt-shunt feedback amplifier.

The open-loop analysis begins by evaluating the effects of the β network on the basic amplifier using the rules stated earlier. The resistance, $R_{\beta i}$, will be the equivalent resistance seen looking into the β network from the input with the output node shorted to ground. The equivalent resistance looking into the β network from the output, $R_{\beta o}$, will be calculated with the input shorted to ground.

Since we are using shunt-shunt feedback, A_{OL} will be defined as

$$A_{OL} = \frac{v_2^*}{i_s^*} = \frac{v_2^*}{v_{g2}^*} \cdot \frac{v_{g2}^*}{v_1^*} \cdot \frac{v_1^*}{i_s^*} \tag{31.47}$$

Solving the first term in Eq. (31.47), can be quite extensive if using standard circuit analysis. A circuit technique based on two-port theory will greatly simplify the analysis. For example, the equivalent circuit for the gain, $\frac{v_2^*}{v_{g2}^*}$, can be seen in Fig. 31.22a. However, the circuit seen in Fig. 31.22b shows the equivalent circuit using an equivalent transconductance, G_M, and output resistance, R_{Leq}. The gain of the equivalent circuit, and hence the actual circuit, is

$$\frac{v_2^*}{v_{g2}^*} = G_M R_{Leq} \tag{31.48}$$

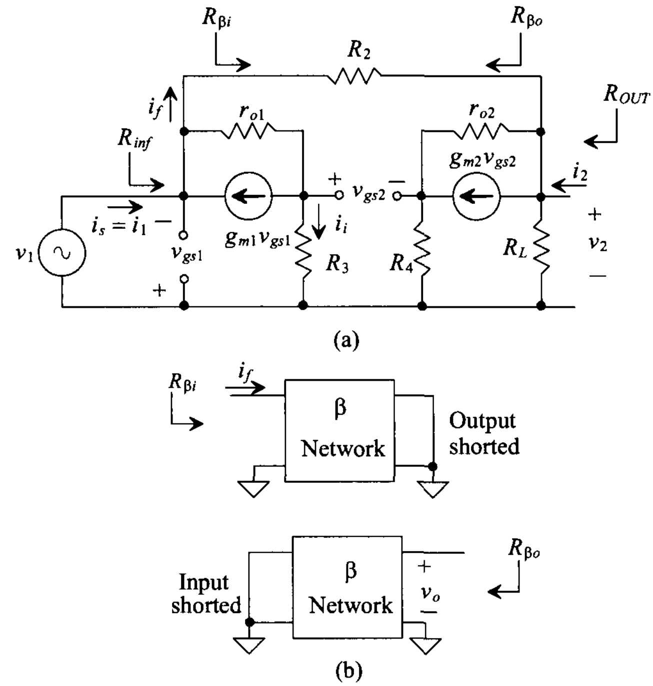


Figure 31.20 (a) Closed-loop small-signal model of Fig. 31.19 and (b) method for determining the feedback network loading.

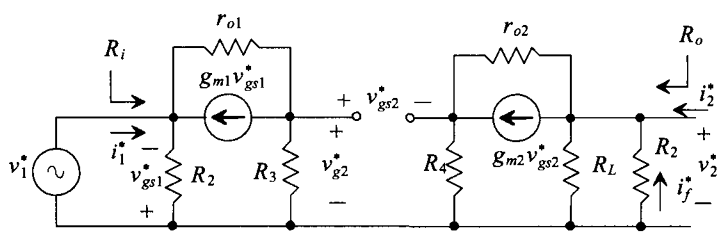


Figure 31.21 Open-loop small-signal model of Fig. 31.19.

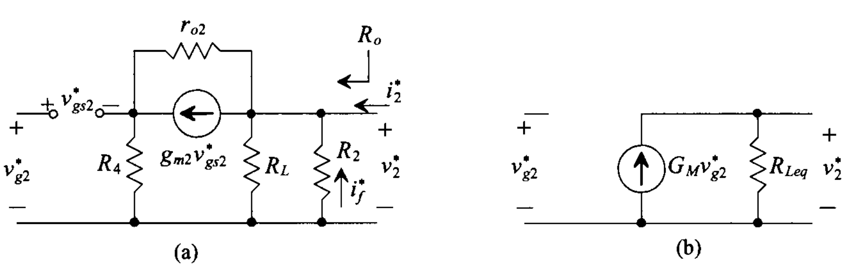


Figure 31.22 (a) Solving a portion of Fig. 31.21, including the drain-to-source resistance, and (b) the equivalent transconductance model.

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

$$i_o^* = -g_{m2}v_{gs2} + \frac{v_{s2}}{r_{o2}} \tag{31.52}$$

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

$$v_{s2}^* + v_{gs2}^* = v_{g2}^* \tag{31.54}$$

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

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

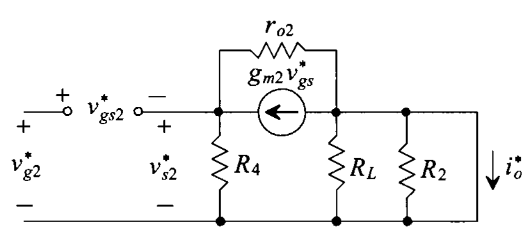


Figure 31.23 Circuit used to determine the equivalent transconductance.

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

$$\frac{v_{g2}^*}{v_1^*} = \frac{g_{m1}R_3 + \frac{R_3}{r_{o1}}}{1 + \frac{R_3}{r_{o1}}} \tag{31.57}$$

The last term in Eq. (31.47) is simply the input resistance, R_i, of the open-loop circuit shown. Using a test source, we can determine this value to be

$$R_i = \frac{v_1^*}{i_s^*} = \frac{v_t}{i_t}||R_2 = \left(\frac{1 + \frac{R_3}{r_{o1}}}{g_{m1} + \frac{1}{r_{o1}}}\right)||R_2 \tag{31.58}$$

Therefore, the entire expression for the open-loop gain, A_{OL}, becomes

$$A_{OL} = \frac{v_2^*}{i_s^*} = \frac{-g_{m2}(R_L||R_2||((1 + g_{m2}R_4)r_{o2} + R_4))}{1 + g_{m2}R_4 + \frac{R_4}{r_{o2}}} \cdot \frac{g_{m1}R_3 + \frac{R_3}{r_{o1}}}{1 + \frac{R_3}{r_{o1}}} \cdot \left(\frac{1 + \frac{R_3}{r_{o1}}}{g_{m1} + \frac{1}{r_{o1}}}\right)||R_2\ \Omega \tag{31.59}$$

At first glance, this equation may appear quite daunting. However, notice that if r_{o2} is much greater than R_2, R_4, and R_L and if r_{o1} is much greater than R_3, the open-loop gain simplifies to

$$A_{OL} = \text{[common source amp with source resistance][common gate amp]}[R_i]$$

$$\approx \frac{-g_{m2}(R_L||R_2)}{1 + g_{m2}R_4} \cdot g_{m1}R_3 \cdot \frac{1}{g_{m1}}||R_2\ \Omega \tag{31.60}$$

Equation (31.60) is typically used for discrete designs in which higher values of currents are used, therefore requiring lower resistor values. However, active loads can easily be used, as seen in Fig. 31.24. Here, the resistor, r_{o3} now replaces R_3. Since R_4 is a source degeneration resistor, its value will be small. The active load that replaces R_4 is a gate-drain connected device and equal to $\frac{1}{g_{m4}}||r_{o4}$. And the resistor, R_L, is now replaced by r_{o5}. Therefore, the open-loop gain of Fig. 31.24 can be written by using Eq. (31.29) and making the proper substitutions:

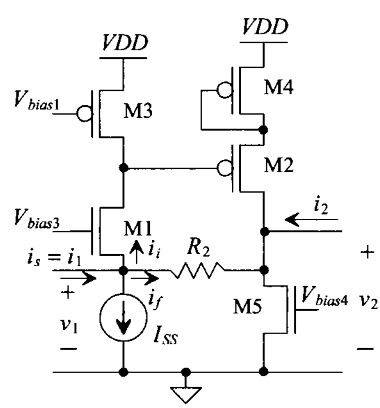


Figure 31.24 Shunt-shunt feedback amplifier using active loads.

$$A_{OL} = \frac{-g_{m2}(r_{o5}||R_2||(1+g_{m2}(\frac{1}{gm4}||r_{o4}))r_{o2}+\frac{1}{gm4}||r_{o4}))}{1+g_{m2}\left(\frac{1}{gm4}||r_{o4}\right)+\frac{\frac{1}{gm4}||r_{o4}}{r_{o2}}} \cdot \frac{g_{m1}r_{o3}+\frac{r_{o3}}{r_{o1}}}{1+\frac{r_{o3}}{r_{o1}}}\left[\frac{1+\frac{r_{o3}}{r_{o1}}}{g_{m1}+\frac{1}{r_{o1}}}||R_2\right] \tag{31.61}$$

which can be approximated as

$$A_{OL} = \frac{v_2^*}{i_s^*} \approx \frac{-g_{m2}(R_2)}{1+g_{m2}\left(\frac{1}{gm4}\right)} \cdot \frac{1+g_{m1}r_{o3}}{2} \cdot \left(\frac{1}{g_{m1}}||R_2\right)\ \Omega \tag{31.62}$$

if it is assumed that $r_{o1} \approx r_{o3}$ and that the discrete resistor, R_2, is relatively small compared to the impedances of the active loads. An active device could have been used to substitute even the resistor, R_2. This will be discussed further later in the chapter.

Next, the value of β is calculated as

$$\beta = \frac{i_f^*}{v_2^*} = -\frac{1}{R_2}\ \text{mhos} \tag{31.63}$$

Again, note that $A_{OL}\beta$ is unitless and overall positive.

Now that all of the open-loop parameters, A_{OL}, β, R_i, and R_o, have been found, the closed-loop values are easily calculated. The value for the closed-loop gain is

$$A_{CL} = \frac{v_2}{i_s} = \frac{A_{OL}}{1+A_{OL}\beta} \tag{31.64}$$

The closed-loop input impedance is

$$R_{inf} = \frac{R_i}{1+A_{OL}\beta} \tag{31.65}$$

and the value for the closed-loop output impedance is

$$R_{of} = R_{out} = \frac{R_o}{1+A_{OL}\beta} \tag{31.66}$$

We should be able to see a trend in the feedback effects. Shunt mixing and shunt sampling cause the input and output impedances to decrease by the factor $(1 + A_{OL}\beta)$. Similarly, series input mixing and series sampling cause the input and output impedances to increase by the factor $(1 + A_{OL}\beta)$.

In many cases, the overall gain is expressed in terms of a voltage gain, $\frac{v_2}{v_1}$. Since we have calculated the transfer function in terms of the current, i_s, we can easily express this in terms of v_1 by

$$\frac{v_2}{v_1} = \frac{v_2}{i_s(R_{inf})} = A_{CL}\left(\frac{1}{R_{inf}}\right) \tag{31.67}$$

31.5.1 Simple Feedback Using a Gate-Drain Resistor

One of the most popular, and simplest, examples of shunt-shunt feedback is seen in Fig. 31.25 and consists of a simple inverting amplifier with a resistor connecting the gate and the drain. The feedback resistor, R_2, is usually a large value and serves several important

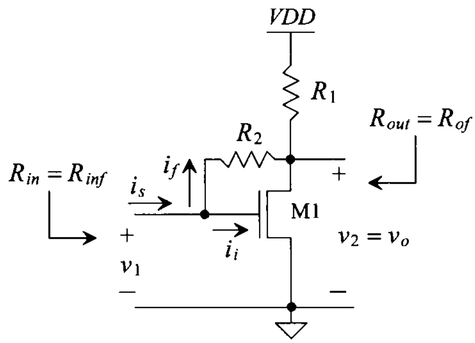


Figure 31.25 Shunt-shunt feedback using a simple gate-drain connected device.

functions. When analyzing the DC characteristics of this circuit, the voltage at the drain will be equal to the voltage on the gate, since there is no DC current flowing through R_2 (when the input is AC coupled). This ensures that the device is always in saturation and provides biasing with no other components needed. We will examine the effects of R_2 on the AC response of the amplifier and soon discover that it has little effect on the gain (if it is large) of the amplifier as well.

Figure 31.26a and b shows the small-signal models for the closed and open-loop circuits, respectively. Since the feedback resistor sums current at the gate of M1, the mixing circuit is shunt. And since the feedback is taken off of the same node as the output, the sampling circuit is shunt. Therefore, the value of $R_{\beta i}$ for the open-loop model is found by shorting the output and looking into the feedback loop from the input, and the value of $R_{\beta o}$ is found by shorting the input and looking into the feedback loop from the output.

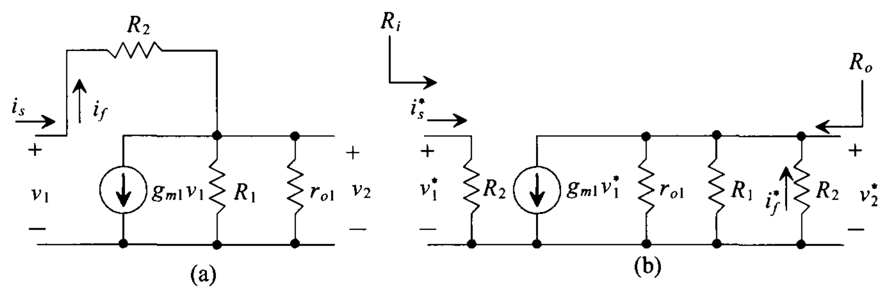


Figure 31.26 Small-signal model of Fig. 31.25: (a) the closed-loop model and (b) the open-loop model.

The open-loop values can be calculated as

$$A_{OL} = \frac{v_2^*}{i_s^*} = \frac{v_2^*}{v_1^*} \cdot \frac{v_1^*}{i_s^*} = [-g_{m1}(R_1 || R_2 || r_{o1})][R_2] \text{ V/A} \tag{31.68}$$

$$R_i = R_2 \tag{31.69}$$

$$R_o = R_1 || R_2 || r_{o1} \tag{31.70}$$

and

$$\beta = \frac{i_f^*}{v_2^*} = -\frac{1}{R_2} \tag{31.71}$$

Using Eqs. (31.43) - (31.46), we find that the closed-loop values become

$$A_{CL} = \frac{v_2}{i_s} = \frac{A_{OL}}{1 + A_{OL}\beta} = \frac{-g_{m1}R_oR_2}{1 + g_{m1}R_oR_2\frac{1}{R_2}} \tag{31.72}$$

$$R_{inf} = \frac{v_1}{i_s} = \frac{R_2}{1 + g_{m1}R_oR_2\frac{1}{R_2}} \text{ and } R_{of} = \frac{R_o}{1 + g_{m1}R_oR_2\frac{1}{R_2}} \tag{31.73}$$

The value of R_{inf} is also equal to the value of R_{in} since no source resistance is associated with v_1. Notice that the value of the closed-loop gain is dependent on R_2. However, most applications of this amplifier use voltage as the input variable. Therefore, the value of the overall voltage gain becomes

$$\frac{v_2}{v_1} = \frac{v_2}{i_s} \cdot \frac{i_s}{v_1} = A_{CL} \cdot \frac{1}{R_{inf}} = -g_{m1}R_o = -g_{m1}(R_1 || R_2 || r_{o1}) \tag{31.74}$$

and if the value of R_2 is chosen to be much larger than R_1 then its effect on the AC midband gain is minimized.

Example 31.2

Calculate the gain, $\frac{v_o}{v_s}$, of the shunt-shunt amplifier in Fig. 31.27 assuming that A = 500,000 V/A, R_i = 10 Ω, and R_o = 10 Ω. Notice that the circuit is similar to a simple inverting op-amp.

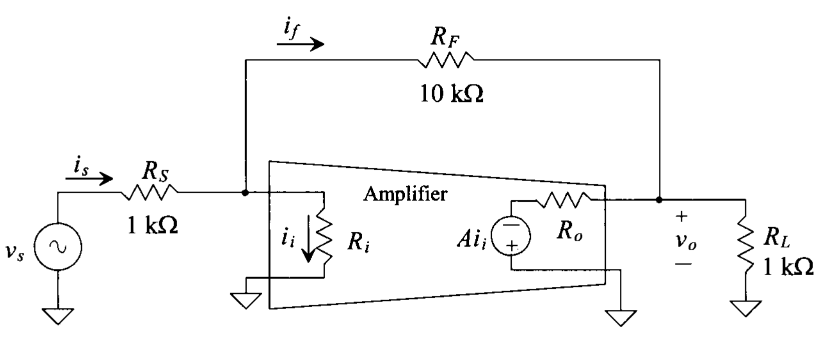


Figure 31.27 A transimpedance amplifier example.

First, we will source transform the voltage source to a current source, since the input mixing is shunt. The model of the amplifier is in terms of A, not A_{OL}, since we must now include the loading of the feedback resistor, and R_L and R_S in the calculation of A_{OL}. The basic amplifier circuit, a transimpedance amplifier with units of V/I, has an ideal input and output impedance of 0 Ω. Next, we will determine the loading of the β network consisting of R_F on the basic amplifier as seen in Fig. 31.28. A_{OL} can be determined as

$$A_{OL} = \frac{v_o^*}{i_s^*} = -A \cdot \left(\frac{R_F || R_L}{R_F || R_L + R_o} \right) \cdot \left(\frac{R_s || R_F}{R_s || R_F + R_i} \right) \Omega$$

$$= -500,000 \cdot 0.989 \cdot 0.989 = -489,060 \text{ V/A}$$

The value of β is easily calculated as

$$\beta = \frac{i_f^*}{v_o^*} = -\frac{1}{R_F} = -0.0001 \text{ A/V}$$

And A_{CL} becomes

$$A_{CL} = \frac{v_o}{i_s} = \frac{-489,060}{1 + -489,060 \cdot -0.0001} = -9.8\,\text{k}\Omega \approx R_F$$

Since $v_s = i_s \cdot R_s$, the overall voltage gain is given by

$$\frac{v_o}{v_s} = \frac{v_o}{i_s} \cdot \frac{1}{R_s} = A_{CL} \cdot \frac{1}{R_s} = -\frac{9.8\text{k}}{1\text{k}} = -9.8 \text{ V/V} \approx -\frac{R_F}{R_s}$$

Notice that this is the gain of the standard inverting op-amp configuration. ■

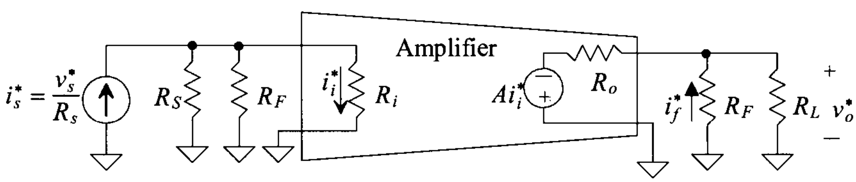


Figure 31.28 Open-loop transimpedance amplifier used in Example 31.2.

31.6 The Transconductance Amp (Series-Series Feedback)

A series-series feedback amp with open-loop values is shown in Fig. 31.29. Ideally, the values of $R_{\beta i}$ and $R_{\beta o}$ are zero. The feedback resistor R_F is in series with the input and the output of the amplifier. Therefore, the input of the amplifier is a voltage, and the output is a current. The units of A_{OL} will be I/V (a transconductance), and the units of β will be V/I (ohms). Transconductance amplifiers also have high-input and high-output impedance.

The closed-loop gain of the transconductance amplifier is

$$A_{CL} = \frac{i_o}{v_s} = \frac{A_{OL}}{1 + A_{OL}\beta} \text{ A/V} \tag{31.75}$$

The input impedance is again given by applying a test voltage to the input of the feedback amp and calculating the current that flows into the input of the amplifier. Also, the assumption that the feedback network does not load the amplifier will be used. Setting $v_{test} = v_s$, $i_{test} = i_s$ and writing a loop at the input of the amplifier give

$$v_{test} = i_{test} \cdot R_i + v_f = i_{test} \cdot R_i + \beta \cdot A_{OL} \cdot i_{test} \cdot R_i \tag{31.76}$$

and so the input resistance with feedback is now

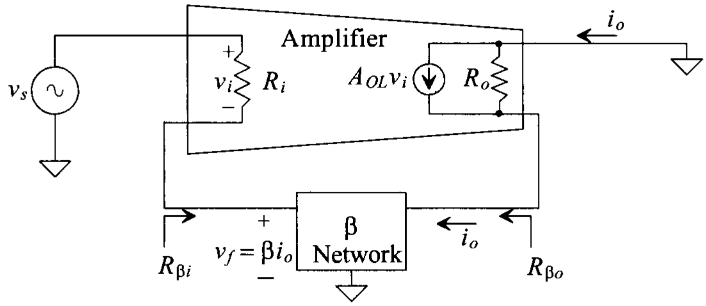


Figure 31.29 An ideal transconductance amplifier.

$$R_{inf} = \frac{v_{test}}{i_{test}} = R_i \cdot (1 + A_{OL}\beta) \text{ for } R_o \to \infty \tag{31.77}$$

keeping in mind that $i_{test} \cdot R_i = v_i$ and $A_{OL} \cdot v_i = i_o$. The output resistance is determined by applying a test current at the output with the input source shorted, and is given by

$$v_{test} = (i_{test} - A_{OL}v_i) \cdot R_o = [i_{test} - A_{OL} \cdot (-\beta i_{test})] \cdot R_o \tag{31.78}$$

since $i_{test} = i_{out}$, $v_i = -v_f$, and $R_{\beta o} = 0$. The output resistance is given by

$$R_{of} = \frac{v_{test}}{i_{test}} = R_o \cdot (1 + A_{OL}\beta) \tag{31.79}$$

The ideal transconductance amplifier has infinite output and input resistance. Again, feedback helps to make the amplifier appear closer to the ideal.

Now examine the transistor level circuit in Fig. 31.30. Notice the similarity to Fig. 31.13, the circuit used in the series-shunt example; the only difference is the location of the output connection. The feedback loops in both circuits are identical, so the feedback in Fig. 31.30 is known to be negative.

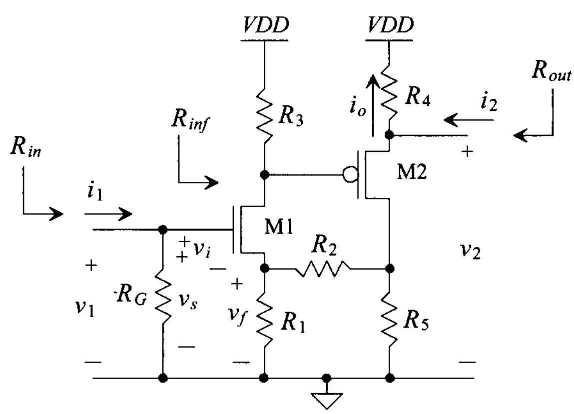


Figure 31.30 Transistor-level series-series feedback amplifier.

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 $R_{\beta i} = R_2 + R_5$.

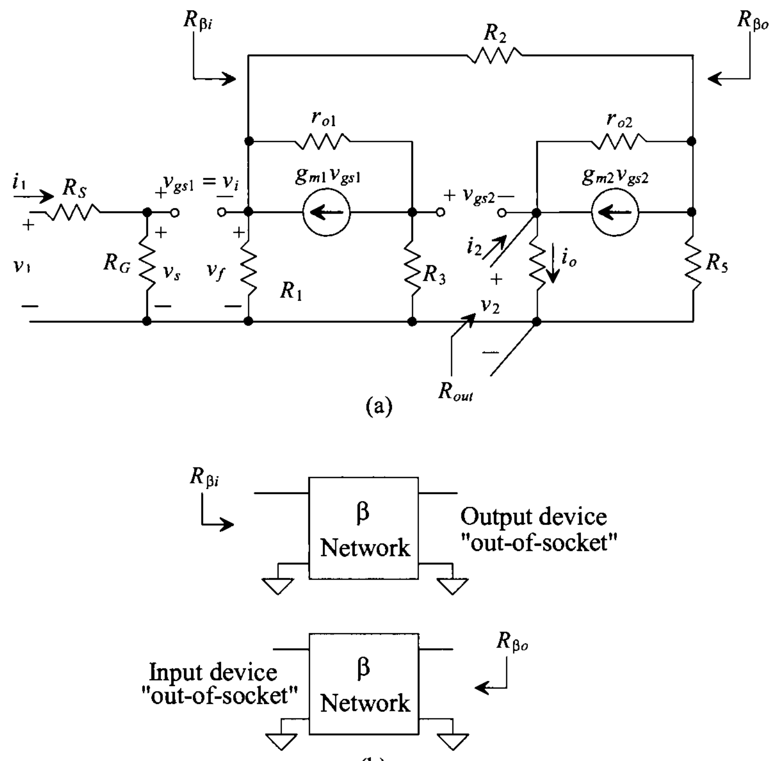


Figure 31.31 (a) Closed-loop small-signal model of Fig. 31.30 and (b) method for determining feedback loading.

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

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

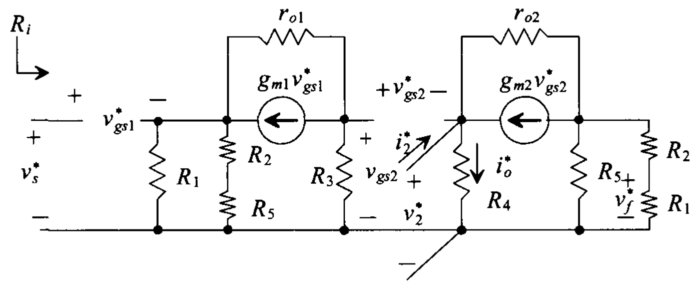


Figure 31.32 Open-loop small-signal model of Fig. 31.30.

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

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

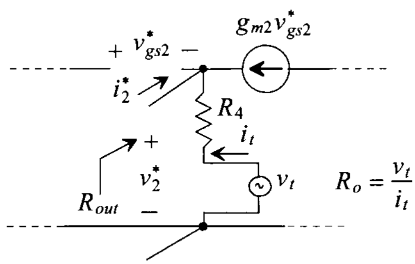


Figure 31.33 Calculation of the output impedance for the circuit in Fig. 31.30.

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

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

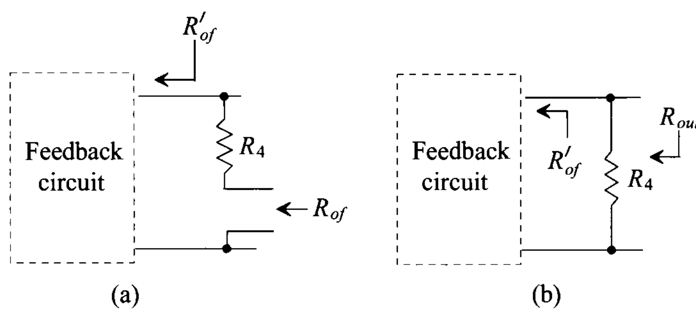


Figure 31.34 Determining the output resistance of a series sampling circuit.

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.

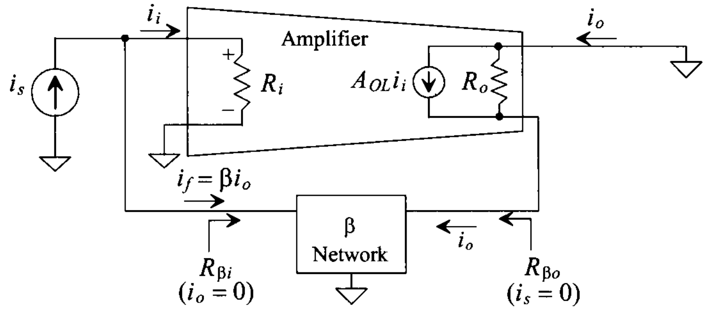


Figure 31.35 An ideal current feedback amplifier.

The transistor-level circuit shown in Fig. 31.36 is similar to the shunt-shunt topology, except for the placement of the output signal. The reader will also be asked to analyze the shunt-series circuit in the Problems section.

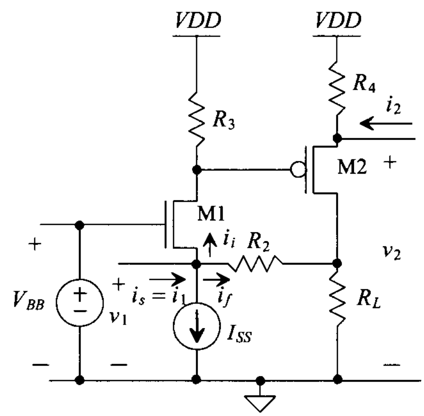


Figure 31.36 A shunt-series feedback amplifier.

Example 31.3

Examine the current amplifier seen in Fig. 31.37a. Using feedback analysis, draw open-loop small-signal models and derive values for A_{OL}, β, R_i, R_o, R_{out}, and the overall voltage gain, $\frac{v_2}{v_1}$.

Notice that although this circuit is similiar to the cascode current sink, we can also use it, though unconventionally, as a voltage op-amp. The analysis begins by identifying the feedback circuit. The current summation that occurs at the gate of M4 indicates that shunt mixing is utilized. Since the output and the feedback are taken off separate terminals of the output device, the output sampling is series. The open-loop model is seen in Fig. 31.37b. The only loading on the input due to the feedback network is r_{o2}, since M4 can be taken "out-of-socket" ($v_{gs2} = 0$). The

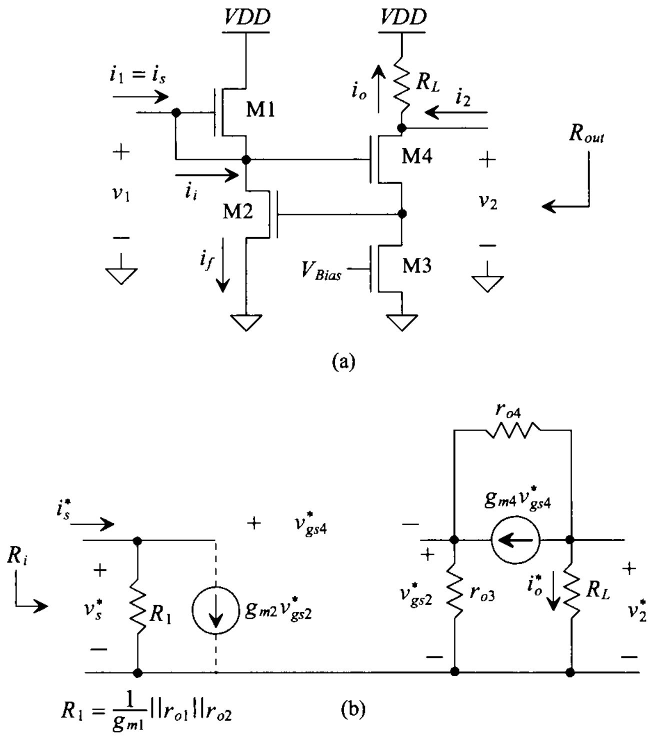


Figure 31.37 (a) Circuit used in Ex. 31.3 and (b) the open-loop model.

feedback does not load the output at all. However, we must remember to include the controlled source for determining i_f. The dependent source, $g_{m2}v_{gs}$, will not be included in calculating A_{OL} but will be needed for calculating β. Solving this open-loop circuit for A_{OL} yields (see Problem 31.28):

$$A_{OL} = \frac{i_o^*}{i_s^*} = \frac{v_2^*}{v_s^*} \cdot \frac{R_1}{R_L} = \frac{-g_{m4}(R_L||((1+g_{m4}r_{o3})r_{o4}+r_{o3}))}{1+g_{m4}r_{o3}+\frac{r_{o3}}{r_{o4}}} \cdot \frac{R_1}{R_L} \approx \frac{-g_{m4}R_1}{1+g_{m4}r_{o3}+\frac{r_{o3}}{r_{o4}}} \text{ A/A}$$

The output, R_o, is found by the same manner presented in the discussion of series-series amplifiers and is

$$R_o = R_L + (1+g_{m4}r_{o3})r_{o4} + r_{o3}$$

Similiarly, the value of R_{out} becomes

$$R_{out} = (R_o(1+A_{OL}\beta) - R_L)||R_L$$

and R_i is simply R_1, with

$$R_{if} = \frac{R_1}{1+A_{OL}\beta}$$

The feedback variable is $i_f^* \approx g_{m2} v_{gs2}^* = -g_{m2} i_o^* r_{o3}$, and the value of β can be calculated as

$$\beta = \frac{i_f^*}{i_o^*} = -g_{m2} r_{o3} \text{ A/A}$$

The overall voltage gain is

$$\frac{v_2}{v_1} = \frac{i_o \cdot R_L}{i_s \cdot R_{if}} = \frac{A_{OL}}{1 + A_{OL}\beta} \cdot \frac{R_L}{R_{if}} \text{ V/V} \blacksquare$$

31.8 Stability

The previous sections illustrated the benefits of feedback and the corresponding tradeoff in gain. However, a critical concern must be examined when applying negative feedback. Some circuits will cause a phase shift in the input signal large enough that the feedback becomes positive (the output adds to the original input), resulting in an unstable system. The occurrence of instability can be minimized with some careful analysis of both the open-loop amplifier, A_{OL}, and the feedback network, β.

The *loop gain* is defined as

$$T = A_{OL}\beta \tag{31.90}$$

Remember that the product of $A_{OL}\beta$ must itself always be positive. By examining the frequency response of the loop gain, *T*, the overall stability of the system can be determined. The stability of the system can be summarized with the following rules and is illustrated in Fig. 31.38.

Case 1: If the change in the phase of $A_{OL}\beta$ is equal to 180° and the magnitude is below 0 dB, the system will be stable.

Case 2: If the change in the phase of $A_{OL}\beta$ is equal to 180° and the magnitude equals 0 dB, the system may or may not be stable.

Case 3: If the change in the phase of $A_{OL}\beta$ is equal to 180° and the magnitude is above 0 dB, the system will be unstable.

One might wonder why the 180° phase shift and the magnitude of 0 dB (gain of 1) are such critical factors. The answer may be better understood with a qualitative rather than a quantitative explanation. In order for positive feedback to occur, the output must be added back to its original input signal. The PA system example illustrates this concept. If the loudspeaker, which represents the output of the system, is added back to the input (the microphone), the system becomes unstable because the amplifier is attempting to amplify its own output. The result is a loud, high-pitched, ringing sound, which most people have unfortunately experienced.

In negative feedback applications, the addition of the feedback signal to the original input occurs because of the additional phase shift introduced by the frequency dependent components within A_{OL} and β. Because the feedback signal is already inverted with respect to the input, an additional 180° of phase shift will cause the feedback signal to be positive with respect to the input. It is this second 180° degrees of phase shift that becomes the important concern. Thus, the frequency response of the loop gain must be

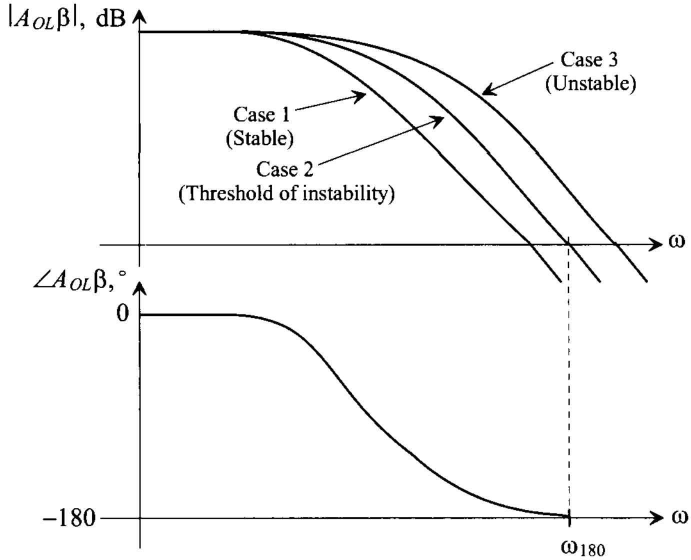


Figure 31.38 Stability analysis using the frequency response of the loop gain.

examined. The magnitude of 0 dB is also important. If the gain around the loop is less than 1, the output settles to a stable value. However, if the gain around the loop is greater than 1, the amplifier output grows and becomes unstable quickly.

Assume that A_{OL} can be described with the following frequency response:

$$A_{OL}(s) = -\frac{10}{\left(1+\frac{s}{10}\right)^2} \tag{31.91}$$

Two poles exist at $\omega = 10$ rad/s with a gain at DC equal to −10 V/V. Also assume that β is frequency dependent and has a single pole at $\omega = 10$ rad/s:

$$\beta = \frac{-1}{(\frac{s}{10}+1)} \tag{31.92}$$

The loop gain then, is

$$A_{OL}\beta = \frac{10}{\left(1+\frac{s}{10}\right)^2} \cdot \frac{1}{\left(\frac{s}{10}+1\right)} \tag{31.93}$$

The Bode plot of the loop gain can be seen in Fig. 31.39. It can be seen that since the phase plot crosses 180° slightly before the magnitude plot crosses 0 dB, the system is unstable.

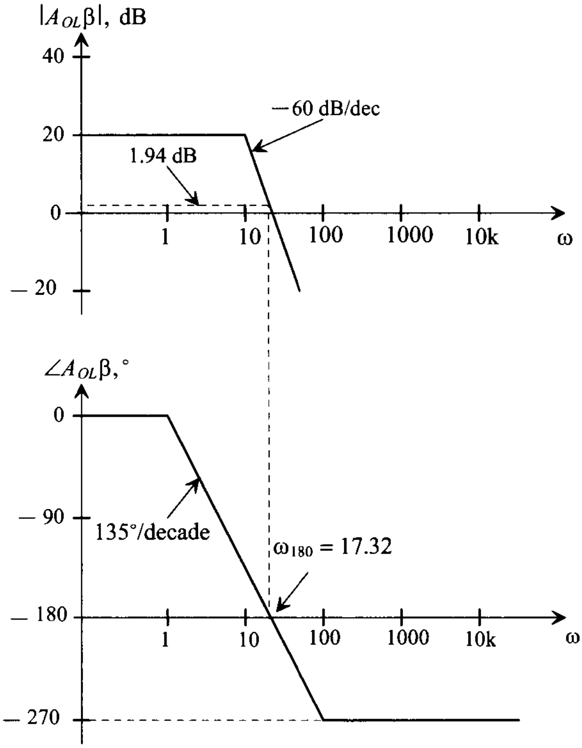


Figure 31.39 Stability analysis using the frequency response of the loop gain.

A more precise method of performing the same analysis will now be described. We can use the exact formulas to solve for the exact frequency which the phase plot crosses 180°. The phase of the loop gain, by definition, is

$$Arg[A_{OL}(j\omega)\beta(j\omega)] = \tan^{-1}\left(\frac{\omega}{z_1}\right) + \tan^{-1}\left(\frac{\omega}{z_2}\right) + \ldots + \tan^{-1}\left(\frac{\omega}{z_n}\right)$$
$$- \tan^{-1}\left(\frac{\omega}{p_1}\right) - \tan^{-1}\left(\frac{\omega}{p_2}\right) - \ldots - \tan^{-1}\left(\frac{\omega}{p_n}\right) \qquad (31.94)$$

where $z_1, z_2 \ldots z_n$ are the zeros in the system and $p_1, p_2 \ldots p_n$ are the poles. Since our example has three poles at the same frequency and no zeros, the phase response of the loop gain can be expressed as

$$Arg[A_{OL}(j\omega)\beta(j\omega)] = -3\tan^{-1}\left(\frac{\omega_{180}}{10}\right) = -180 \qquad (31.95)$$

where ω_{180} is the frequency at which the phase response is equal to −180°. Solving for ω_{180} yields

$$\omega_{180} = 17.32 \text{ rad/s} \qquad (31.96)$$

The magnitude of the loop gain is defined as,

$$20\text{Log}|A_{OL}(j\omega)\beta(j\omega)| = 20\text{Log}(A_o) + 20\text{Log}\sqrt{\left(\tfrac{\omega}{z_1}\right)^2+1} + 20\text{Log}\sqrt{\left(\tfrac{\omega}{z_2}\right)^2+1} + \ldots$$

$$+ 20\text{Log}\sqrt{\left(\tfrac{\omega}{z_n}\right)^2+1} - 20\text{Log}\sqrt{\left(\tfrac{\omega}{p_1}\right)^2+1} - \ldots - 20\text{Log}\sqrt{\left(\tfrac{\omega}{p_n}\right)^2+1} \qquad (31.97)$$

where, $z_1, z_2 \ldots z_n$ are the zeros of the system, $p_1, p_2, \ldots p_n$ are the poles, and A_o is the midband gain. Plugging in the value for ω_{180} into Eq. (31.97), we can solve for the magnitude of the loop gain at the point at which its phase is $-180°$:

$$20\text{Log}|A_{OL}(j\omega)\beta(j\omega)| = 20\text{Log}(10) - 3\left(20\text{Log}\sqrt{\left(\tfrac{17.32}{10}\right)^2+1}\right) \qquad (31.98)$$

$$|A_{OL}(j\omega_{180})\beta(j\omega_{180})| = 1.94 \text{ dB} \qquad (31.99)$$

which, using rule 3, verifies that the system is unstable.

As discussed in Ch. 24 stability is typically measured using two specifications: gain margin and phase margin. Gain margin is defined as the difference between the magnitude of $A_{OL}\beta$ at ω_{180} and unity, whereas phase margin is defined as the difference between the value of the phase at the frequency at which the magnitude of $A_{OL}\beta$ is equal to unity and ω_{180}. Figure 31.40 illustrates both definitions. It should be noted that phase margin is the typical specification used for stability. Amplifiers should be designed to have a phase margin of at least 45°, though 60° phase margin is more acceptable.

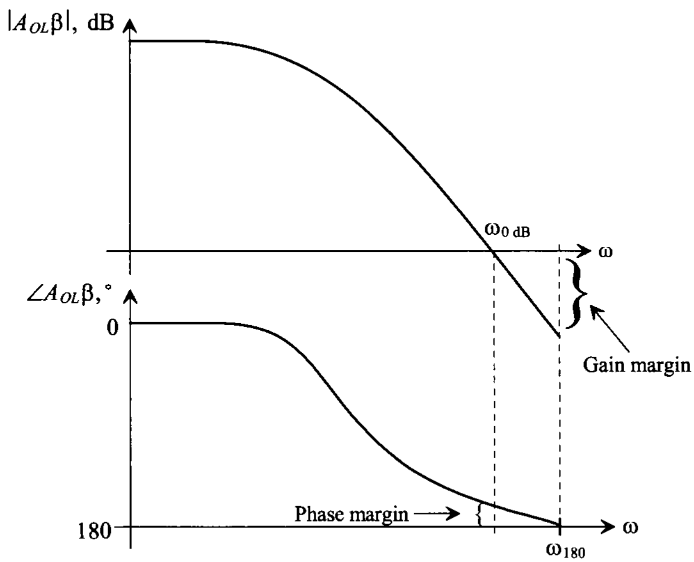


Figure 31.40 Gain margin and phase margin definitions.

In our previous discussion, only the frequency domain analysis was considered. The effect of phase margin in the time domain is related to settling time. As the phase margin increases, less time is required for the signal to settle. As the phase margin approaches 0°, the signal will oscillate indefinitely.

31.8.1 The Return Ratio

In some cases, it may be more practical to determine the loop gain from a system point of view. One method used to determine a good approximation of the loop-gain frequency response is to break the loop, input a test signal, and determine the returned value back to the point that the loop was broken. This method is called the return ratio (*RR*) method. Consider the block diagram of the single-loop structure in Fig. 31.41 with the loop "open." The gain around the loop is,

$$RR = \frac{x_f}{x_i} = -T = -A_{OL}\beta \tag{31.100}$$

This gain represents the path from the input back around through the feedback network. It should be noted that the *RR* method may vary quite extensively from the two-port method in finding the value of $A_{OL}\beta$, but for purposes of plotting the loop gain frequency response, the *RR* should be sufficient [3].

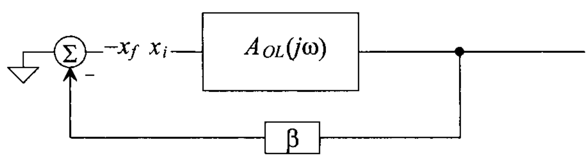


Figure 31.41 Determining the loop gain by opening the feedback loop.

The *RR* is found by first replacing all independent sources with their ideal impedances. Next, a dependent source is chosen, and the feedback loop is broken between the chosen source and the rest of the circuit. An independent test source is inserted at the node where the dependent source resided and a test signal is injected into the loop. The *RR* is then the ratio of the returned signal (which now appears across the dependent source) and the test signal, such that

$$RR = -\frac{v_r}{v_t} \tag{31.101}$$

An example will illustrate this method further.

Example 31.4

Determine the *RR* for the series-shunt op-amp circuit shown in Fig. 31.42a and compare that value to the value $A_{OL}\beta$ using the two-port method. Assume that the op-amp can be modeled as shown in Fig. 31.42b.

Since there is only one dependent source in the circuit, deciding where to break the loop is a simple endeavor. Next, the dependent source is separated from the rest of the circuit, and an independent source, v_t, is put in its original position as

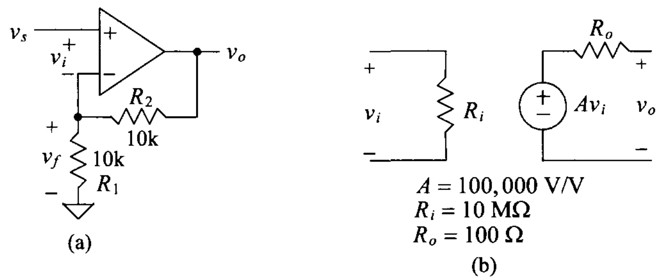


Figure 31.42 (a) A series-shunt amplifier used in Ex. 31.4 and (b) the model used for the op-amp.

seen in Fig. 31.43a. The returned signal is now across the dependent source. By inspection, the value of the RR can be found to be

$$RR = -\frac{v_r}{v_t} = A \cdot \left(\frac{R_I||R_1}{R_I||R_1 + R_2}\right) \cdot \left(\frac{R_2 + R_I||R_1}{R_2 + R_I||R_1 + R_o}\right) = 49,726$$

Next, the open-loop model for the original circuit is found using the two-port analysis presented in Sec. 31.2. The circuit, taking the β network loading effects into account, can be seen in Fig. 31.43b. The value of A_{OL} is easily calculated as

$$A_{OL} = \frac{v_o^*}{v_s^*} = A \cdot \left(\frac{R_1 + R_2}{R_o + R_1 + R_2}\right) \cdot \left(\frac{R_I}{R_1||R_2 + R_I}\right) = 99,497$$

The value of β by inspection is

$$\beta = \frac{v_f^*}{v_o^*} = \frac{R_1}{R_1 + R_2} = 0.5$$

and the value of $A_{OL}\beta$ is

$$A_{OL}\beta = (99,497)(0.5) = 49,748$$

The two methods yielded very similiar answers. ■

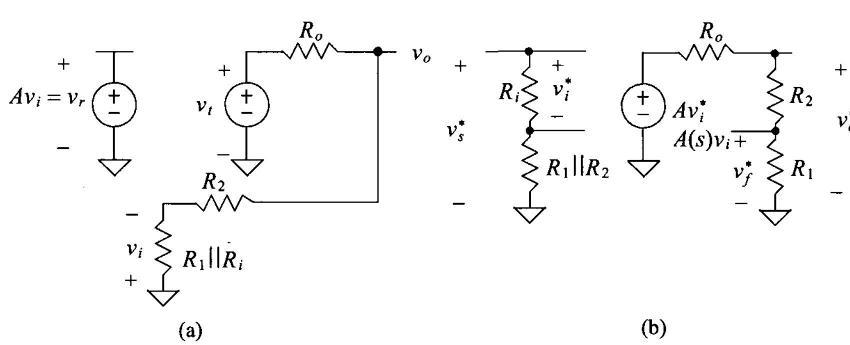


Figure 31.43 (a) The model used to calculate return ratio and (b) the model used for the two-port analysis of loop gain for Fig. 31.42.

If $A_{OL}\beta$ is frequency dependent, then the phase and gain margin of the circuit can be plotted so as to determine the phase and gain margin. Now suppose that the circuit used in Ex. 31.4 was frequency dependent as seen in Fig. 31.44. Using the same strategy presented in Ex. 31.4 and using the RR method, the value of the loop gain becomes

$$RR = -\frac{v_r}{v_t} = \frac{49,726}{(s/200+1)^3} \tag{31.102}$$

The gain and phase margin can then be analyzed to determine if the system is stable (see Problem 31.34).

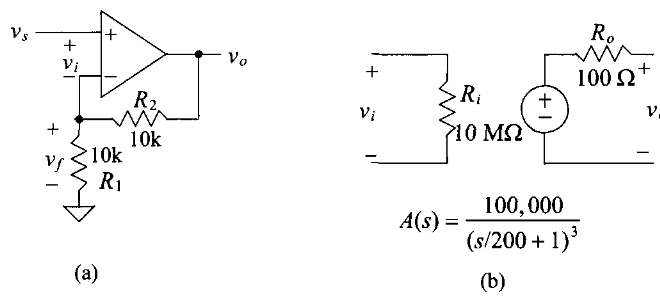


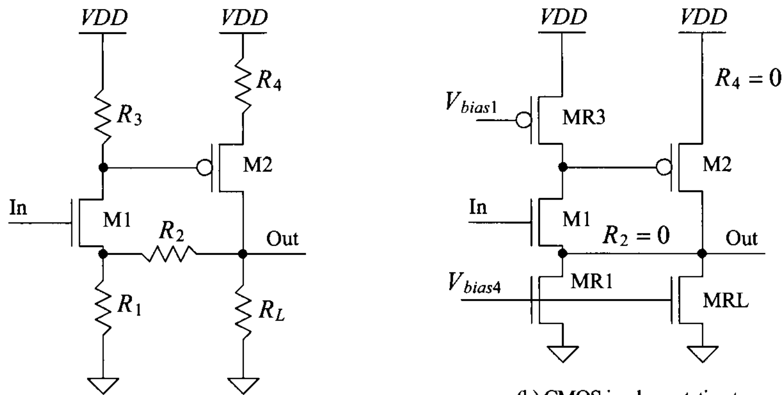
Figure 31.44 (a) A series-shunt amplifier used in Ex. 31.4 and (b) the model used for the op-amp.

31.9 Design Examples

In this section we'll provide some design examples to further demonstrate the design of feedback amplifiers in CMOS technology. For the simulations we'll use the short-channel CMOS process with the sizes and biasing conditions seen in Table 9.2.

31.9.1 Voltage Amplifiers

Figure 31.13 showed the detailed transistor-level implementation of a series-shunt feedback amplifier (voltage in and out) using an AC coupled input. Figure 31.45a shows the implementation with a DC coupled input. Reviewing the equation for A_{OL}, Eq. (31.35), corresponding to this topology we see that to maximize the gain we can replace R_L, R_1, and R_3 with transistors MRL, MR1, and MR3 (so these resistances effectively become r_{oRL}, r_{oR1}, and r_{oR3}) and make R_2 and R_4 zero ohms. The result is seen in Fig. 31.45b. With this selection β becomes 1, Eq. (31.36), while the ideal closed gain, A_{CL}, is also one, Eq. (31.38). Note that we can further increase A_{OL} by using cascode structures. However, A_{OL} isn't the limiting performance factor in this design. The body effect of M1, as we shall shortly see, is the limiting factor. This circuit is useful as a unity voltage buffer providing large input impedance (very little capacitance) and low output resistance. It can source a relatively large current via M2 but the amount of current it can sink is limited to the biasing currents of MR1 and MRL. Note that the allowable input signal range is from $V_{GS} + V_{DS,sat}$ (= 400 mV from Table 9.2) up to $VDD - V_{SD,sat} + V_{THN}$ (= 770 mV again from Table 9.2). The output can swing from $VDD - V_{SD,sat}$ to $V_{DS,sat}$.



(a) DC-coupled series-shunt (voltage amplifier) from Fig. 31.13.

(b) CMOS implementation to maximize open-loop gain. The topology has a closed-loop gain of 1.

Figure 31.45 (a) Series-shunt amplifier and (b) replacing resistors with MOSFETs.

Let's compare the unity buffer seen in Fig. 31.45b to a simple source-follower like the one seen in Fig. 21.39 each driving a 10 pF load. The simulated frequency responses of the amplifiers are seen in Fig. 31.46. Note the increase in bandwidth using the feedback amplifier. Also note the peaking in the series-shunt amplifier's response due to the output impedance becoming inductive at higher frequencies and forming a second-order response with the load capacitance. While the significant enhancement in bandwidth is important let's focus on why the closed-loop gain of the feedback amplifier isn't closer to 1 (0 dB).

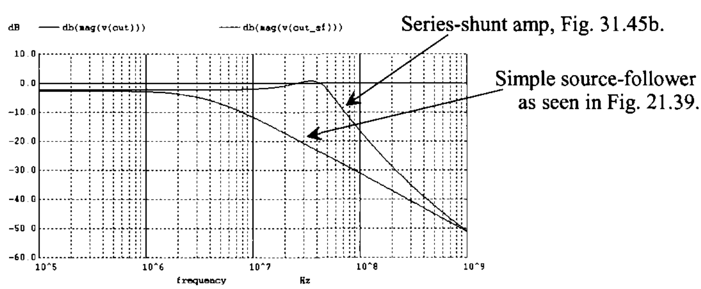
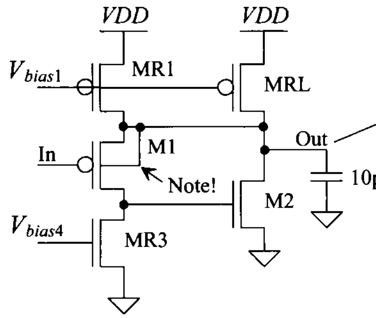
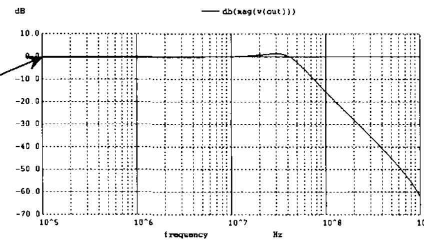


Figure 31.46 Comparing series-shunt feedback amplifier to a source-follower both driving a 10 pF load.

To start we might ask why the source-follower's gain isn't also closer to unity? We answered this question back in Sec. 21.2.4. The body effect causes the reduction in gain. It's easy to verify with simulations (connect M1's body to its source, that is, the output) that this is also the problem with the feedback amplifier. In an n-well process we can't make this connection since the p-substrate is at ground. We could try increasing R_2, say to 25k, to increase the amplifier's gain. However, the process and temperature variations will cause the gain to increase above unity in some situations. A simpler solution is to use the complement of this feedback amplifier where M1 is a PMOS device and M2 is an NMOS device, Fig. 31.47a. The simulation results are seen in Fig. 31.47b.



(a) Series-shunt amplifier with PMOS for M1 and NMOS for M2. Note M1 is in its own well.



(b) Simulating (a) with a 10 pF load.

Biasing supplied by Fig. 20.47, see Table 9.2.

Figure 31.47 (a) PMOS series-shunt amplifier and (b) simulation results.

Before moving on to another topic note that to increase the bandwidth of the amplifier we need to make the capacitive load "effectively" smaller. We know that for general design it's important to keep the overdrive voltages constant between devices by ensuring that we control the devices' biasing currents[1]. What this indicates is that we only have control over the devices' widths once an overdrive, length, and biasing current are selected for a specific transition frequency, f_T. So, to make the capacitive load "effectively" smaller we need to increase the widths of the devices (overdrive voltages remain the same while current goes up). This increases the bandwidth of the amplifier (try increasing the widths of the MOSFETs used in the simulations that generated Figs. 31.46 or 31.47 by 4 and re-simulating with the 10 pF load).

Amplifiers with Gain

Notice in the series-shunt amplifiers seen in Figs. 31.46 or 31.47 that the current flowing in M1 is provided by MR3. This is important since, as we just mentioned, we want to set the biasing current in our transistors. Further, we know the current in M2 as well since the sum of the currents flowing in M1 and M2 must be equal to the currents flowing in MR1 and MRL. In this topology it's easy, as long as we aren't driving a resistive load, to bias the transistors so they have the attributes seen in Table 9.2.

Next, consider the series-shunt amplifier seen in Fig. 31.48a. Here, as before, we've set R_4 to zero and replaced R_3 and R_L with transistors MRL and MR3. The use of transistor loads maximizes A_{OL} as discussed at the beginning of this section (again, A_{OL} can be increased further using cascoding to increase the output resistance of MRL and

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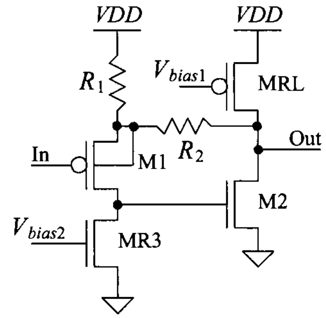
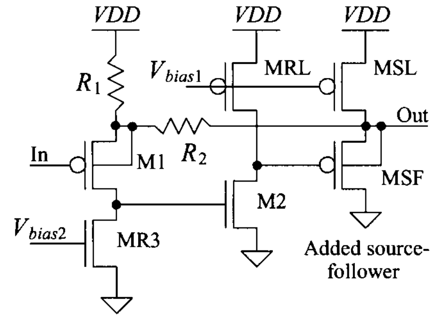
[1] Contradictions to this comment occur when we increased the widths of the diff-pair back in Ch. 26 to increase their g_m, or the widths of a current mirror load to reduce input-referred offset or noise, or we are driving a resistive load and there is no way to keep the driving device's current constant (among other contradictions).



(a) Series-shunt amplifier with gain.



(b) Improved series-shunt amplifier with gain.

Figure 31.48 Using a source-follower to ensure good biasing.

MR3). The current flowing in M1 is still set by MR3. However, now the current flowing in M2 depends on the input signal and the size of the resistors R_1 and R_2. This is bad because we aren't controlling the M2's operating point (e.g., g_m). To avoid this situation consider adding the source follower to the output of the circuit as seen in Fig. 31.48b. Since the source follower's gain is near one we can continue using the equations we derived earlier. Further, now M2's current is set by MRL so its behavior is well-controlled. MSF can sink a current greater than the current supplied by MSL and the resistors since its gate is able to move freely. This is why we must use a PMOS source-follower and not an NMOS source follower, with a constant current sink load, on the output of this circuit. Since MSF is operated as a source follower its g_m and speed performance aren't as important to the overall feedback amplifier's performance.

As a design example let's set, in Fig. 31.48b, R_1 = 1k and R_2 = 9k. Also, to ensure the source-follower has plenty of drive, let's use a multiplication factor, M, of 4 in this stage. The means the widths of MSL and MSF are increased to 400 and their bias currents increase to 40 μA. The overdrive voltages are unchanged from the values listed in Table 9.2. Before simulating let's calculate the allowable input signal range for proper operation. We expect, since the gain of the amplifier is 10, that this swing should be less than *VDD*/10 or 100 mV. The minimum output voltage is set by the source-gate voltage of MSF and the minimum voltage across M2 so that

$$V_{out,\min} = V_{SG} + V_{DS,sat} = 400\ mV$$

Note that MSF's source-gate voltage can actually drop lower than 350 mV since we aren't controlling the current in this device as discussed above. Referring the minimum output voltage back to the source of M1 through the voltage divider

$$\text{Source of M1} = VDD - 400\ mV \cdot \frac{1k}{1k+9k} = 960\ mV$$

Since the V_{SG} of M1 is 350 mV, the input voltage is 610 mV or an input signal swing from (roughly) 560 mV to 660 mV (where we centered the swing on 610 mV since MSF's V_{SG} can be less than 350 mV). Setting the proper input DC operating voltage over process, temperature, and power supply variations can be a significant design concern (AC coupling can help). Figure 31.49 shows the simulation results again with a 10 pF load.

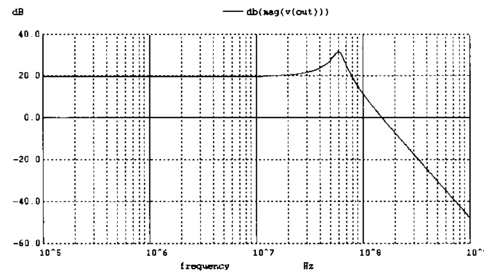


Figure 31.49 Simulating the amplifier in Fig. 31.48b in a gain of 10 configuration.

Finally, notice the increased peaking in the frequency response seen in Fig. 31.49. This additional peaking can be attributed to the added source follower. As discussed on pages 694 and 695 the impedance looking into the output of the source follower (the source of MSF) becomes inductive when it's driven with a resistive load. To remove, or reduce, the peaking note that the drains of M1 and M2 are high-impedance nodes (e.g., nodes 1 and 2 in Fig. 21.25). The compensation procedures given in Chs. 21 and 24 (e.g. using split-length devices as seen in Fig. 24.21) can be used to reduce the bandwidth and cause the amplifier's response to roll-off at – 20 dB/decade.

31.9.2 A Transimpedance Amplifier

The transimpedance, or transresistance, (shunt-shunt) amplifier was discussed back in Sec. 31.5 (and Ex. 8.17). One application of this feedback amplifier is seen in Fig. 31.50. The feedback resistor seen in Fig. 8.34 is removed to increase the gain (to lower the input-referred noise current). *Reset* goes low to enable sensing the diode's current, i_d. The RMS output noise, if C_F is small, is approximately $\sqrt{kT/C_F}$. This topology may not be useful in a communication receiver application where data are continually present on the input of the amplifier; however, it can be very useful in imaging applications. We should also point out why we want the input resistance of the transimpedance amplifier (TIA) to be zero, that is, so that all of the reverse-biased photodiode's current is input to the TIA. Both sides of the diode are at AC ground so the diode's capacitance doesn't steal some of the diode's current (the capacitance doesn't discharge). Note that ideally

$$|A_{OL}| = |-R_m| \to \infty \tag{31.103}$$

The (ideal) closed-loop gain is

$$A_{CL} = \frac{1}{\beta} = \frac{-1}{j\omega C_F} = \frac{v_{out}}{-i_d} \tag{31.104}$$

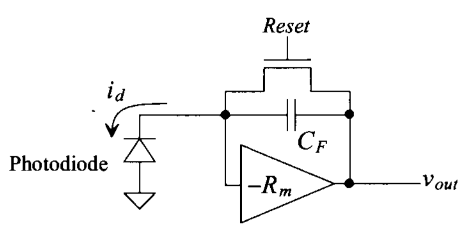


Figure 31.50 Using a transimpedance amplifier (TIA) to sense a photodiode's current.

Let's get an idea for the size of C_F for a particular application. Suppose the diode generates one electron every 100 ns. This is equivalent to an average diode current of $1.6 \times 10^{-19} C/100\, ns$ or 1.6 pA. If we set C_F to 10 fF then the output of the TIA will appear (ideally, neglecting noise and photodiode dark current, i.e., leakage current) as steps, when the electron is generated, with heights of

$$V_{step} = 1.6 \times 10^{-19}/C_F = 16\ \mu V \tag{31.105}$$

Clearly increasing C_F reduces the sensitivity of the TIA. Why not reduce C_F then? The answer is that the device and interconnect parasitics are comparable to 10 fF (if careful) so reducing C_F results in non-ideal behavior. Realistically, 10 fF is questionable if care isn't exercised when designing the TIA.

Returning to the calculations, the output of the TIA rises, when *Reset* is low, at a rate of

$$\frac{dv_{out}}{dt} = \frac{16\ \mu V}{100\ ns} = \frac{i_d}{C_F} = \frac{1.6\ pA}{10\ fF} = 160\ mV/ms \tag{31.106}$$

For an imaging application at 30 images per second we may have upwards of 30 ms to sense the diode's current before resetting the diode. In this case this slow-rate of ascent doesn't appear to be a problem (but it depends on the application). We'll return to these calculations in a moment.

Next we need to select a shunt-shunt feedback amplifier topology. It's important to note that when *Reset* goes high the TIA is in the unity-follower configuration and so negative feedback must be employed (the gain from the input to the output has to be negative). This means that we can't use the topology shown in Fig. 31.19. We can, however, use the topology seen in Fig. 31.25. If we replace R_2 in Eqs. (31.68) - (31.73) with $1/j\omega C_F\ (= Z_2)$ and assume Z_2 is much smaller than R_1 or r_{o1} (and $g_{m1}Z_2 >> 1$) then both R_{inf} and R_{of} approach $1/g_{m1}$ (a relatively small value which is what we want). Also note that by replacing R_1 with a PMOS cascode structure and cascoding M1 we can increase A_{OL} with little effect on the noise performance of the amplifier (see Fig. 21.44). The resulting amplifier is seen in Fig. 31.51. Note that the 10 pF load doesn't affect the performance of the TIA (much) since we are sensing for very long periods of time and the transistors M1-M5 are biased with 10 μA. Also, the reset transistor, M5, is made as small as possible so that the charge injection and capacitive feedthrough (when M5 shuts off) are both minimized.

Figure 31.52 shows the simulation results when the input current, i_d, is 10 pA. We expect the output of the TIA to rise at 160 mV/ms as calculated above. However, we see that the output rises at a rate of, roughly, 500 mV/25 μs considerably faster than our calculated value. *What current is the TIA integrating that could cause this error?* Looking at the simulation results we see that M5 is leaking around 13 pA (source of M5 to ground) into the TIA's input. However, the big problem is the current flowing into the gate of M1. This current is roughly 190 pA! The direct tunneling gate current in this book's short-channel CMOS process, from Table 9.2, is around 5 A/cm^2 (generally measured with *VDD* on the gate and the source/drain/substrate grounded). We aren't seeing this much gate current since the gate voltage of M1 is only around 350 mV (see page 475 and the associated discussion for additional information). The question is how do we modify the TIA's design to reduce this error? The simple solution is to reduce the

Biasing supplied by Fig. 20.47, see Table 9.2.

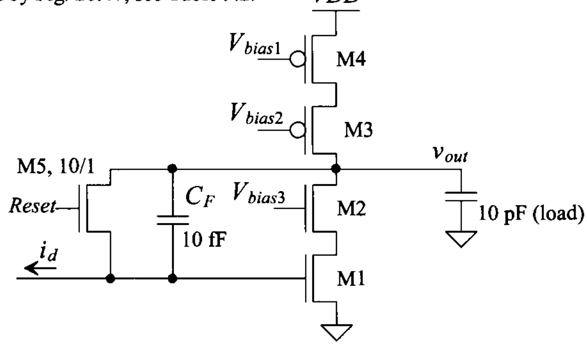


Figure 31.51 A transimpedance amplifier.

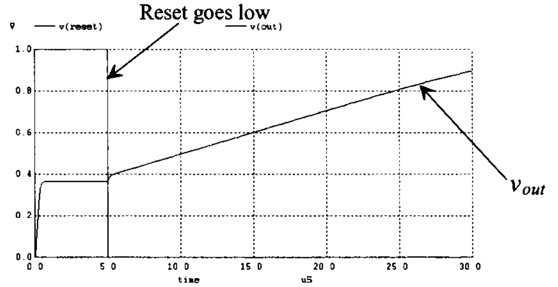


Figure 31.52 Simulating the TIA seen in Fig. 31.51 with 10 pA input current.

widths of M1-M4. Let's drop the NMOSs' widths to 10 and the PMOSs' widths to 20. The actual sizes of the NMOS and PMOS are now 500n/100n and 1μ/100n respectively. Using the same bias circuit, Fig. 20.47, our overdrives remain unchanged and the drain current drops to (roughly) 1 μA. This change also reduces the input-referred noise by increasing A_{OL} and reducing the size of the flicker and thermal drain noise currents. The simulation results, again with a 10 pA input current and 10 pF load, are seen in Fig. 31.53. The leakage current from M5, i_{S5}, is still around 13 pA while the leakage current, i_{G1}, from the gate of M1 dropped to 36 pA. The net current integrated by the TIA is then the 49 pA leakage current from M1 and M5 added to the 10 pA of signal current for an output change of

$$\frac{dv_{out}}{dt} = \frac{i_d + i_{S5} + i_{G1}}{C_F} = \frac{59\,pA}{10\,fF} = 5.9\ V/ms = 59\ mV/10\ \mu s \tag{31.107}$$

which matches fairly well with the simulation results seen in Fig. 31.53. Note that one might be tempted to compensate for M1/M5's leakage current by adding another, opposite direction, leakage path (another MOSFET connected to the input). This is generally not the best solution since this approach increases the TIA's input capacitance and noise. Further the leakage varies with process shifts and VDD. A better solution is to use an older CMOS process, if possible/available, with considerably less gate leakage current.

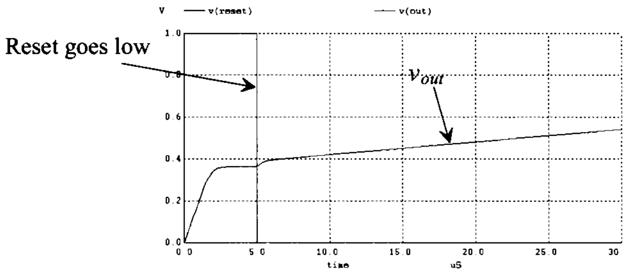


Figure 31.53 Reducing the widths of the MOSFETs in the TIA to lower the integrated gate current from M1.

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[1] A. S. Sedra and K. C. Smith, *Microelectronic Circuits, Sixth Edition,* Oxford University Press, 2010. ISBN 978-0-19-532303-0.

[2] W. M. C. Sansen, *Analog Design Essentials,* Springer, 2006. ISBN 978-0-387-25746-4.

[3] P. J. Hurst, "Exact Simulation of Feedback Circuit Parameters," *IEEE Transactions on Circuits and Systems*, Vol. 38, No. 11, pp. 1382-1389, November 1991.

PROBLEMS

31.1 An op-amp is designed so that the open-loop gain is guaranteed to be 150,000 ± 10 percent V/V. If the amplifier is to be used in a closed-loop configuration with β = 0.1 V/V, determine the tolerance of the closed-loop gain.

31.2 What is the maximum possible value of β using resistors in the feedback loop of a noninverting op-amp circuit? Sketch this op-amp circuit when $\beta = 1/2$.

31.3 Examine the feedback loop in Fig. 31.54. A noise source, v_n, is injected in the system between two amplifier stages. (a) Determine an expression for v_o which includes both the noise and the input signal, v_s. (b) Repeat (a) for the case where there is no feedback ($\beta = 0$). (c) If $A_1 = A_2 = 200$, and feedback is again applied around the circuit, what value of β will be required to reduce the noise by one-half as compared to the case stated in (b)?

31.4 An amplifier can be characterized as follows:

$$A(s) = 10,000 \cdot \frac{100}{s+100} \text{ V/V}$$

A series of these amplifiers are connected in cascade, and feedback is used around each amplifier. Determine the number of stages needed to produce an overall gain of 1,000 with a high-frequency rolloff (at – 20 dB/decade) occurring at 100,000 rad/sec. Assume that the first stage produces the desired high-frequency pole and that the remaining stages are designed so that their high-frequency poles are at least a factor of four greater.

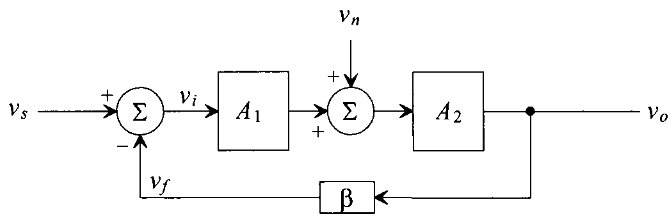

Figure 31.54 Problem 31.3, noise injected into a feedback system.

31.5 An amplifier can be characterized as follows:

$$A(s) = 1{,}000 \cdot \frac{s}{s+100} \text{ V/V}$$

and is connected in a feedback loop with a variable β. Determine the value of β for which the low-frequency rolloff is 50 rad/sec. What is the value of the closed-loop gain at that point?

31.6 Make a table summarizing the four feedback topologies according to the following categories: input variable, output variable, units of A_{OL}, units of β, method to calculate $R_{\beta i}$ and $R_{\beta o}$, and expressions for A_{CL}, R_{if}, and R_{of}.

31.7 Using the two n-channel common source amplifiers shown in Fig. 31.55a and the addition of a single resistor, draw (a) a series-shunt feedback amplifier, (b) a series-series feedback amplifier, (c) a shunt-shunt feedback amplifier, and (d) a shunt-series amplifier. For each case, identify the forward and feedback paths, ensure that the feedback is negative by counting the inversions around the loop, and label the input variable, the feedback variable, and the output variable. Assume that the input voltage has a DC component that biases M1.

31.8 Repeat problem 31.7 using the two-transistor circuit shown in Fig. 31.55b.

31.9 Repeat problem 31.7 using Fig. 31.55c.

31.10 Repeat problem 31.7 using Fig. 31.55d.

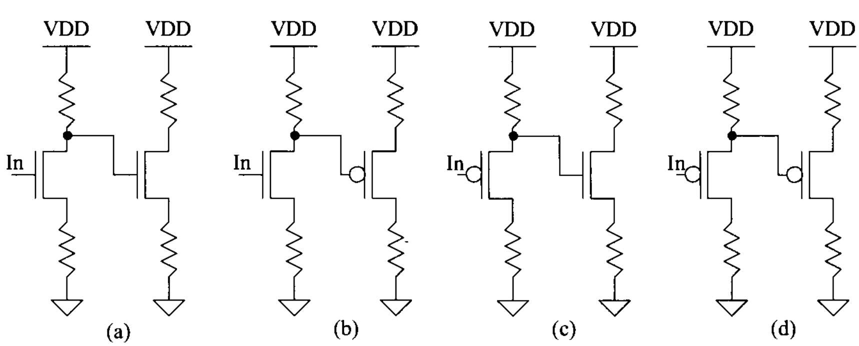

Figure 31.55 Two-transistor feedback topologies.

For each of the following feedback analysis problems, assume that the circuit has been properly DC biased and that MOSFETs have been characterized. The n-channel devices have $g_m = 0.06$ A/V and $r_o = 70$ kΩ. The p-channel devices have $g_m = 0.04$ A/V and $r_o = 50$ kΩ.

31.11 Using the series-shunt amplifier shown in Fig. 31.56, (a) identify the feedback topology by labeling the mixing variables and output variable, (b) verify that negative feedback is employed, (c) draw the closed-loop small-signal model, and (d) find the expression for the resistors $R_{\beta i}$ and $R_{\beta o}$.

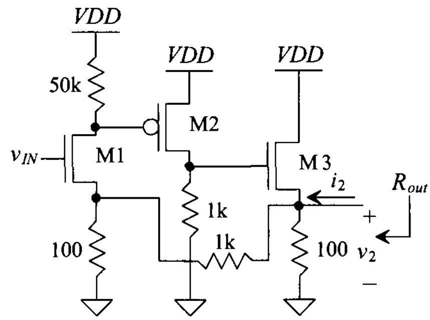

Figure 31.56 A series-shunt feedback amplifier with source-follower output buffer.

31.12 Using Fig. 31.56 and the results from problem 31.11, (a) draw the small-signal open-loop model for the circuit and (b) find the expressions for the open-loop parameters, A_{OL}, β, R_i, and R_o and (c) the closed-loop parameters, A_{CL} and R_{out}. Note that finding R_{in} is a trivial matter since the signal is input into the gate of M1.

31.13 Using the series-shunt amplifier shown in Fig. 31.57, (a) verify the feedback topology by labeling the mixing variables and the output variable closed-loop small-signal model and (b) find the values of $R_{\beta i}$ and $R_{\beta o}$.

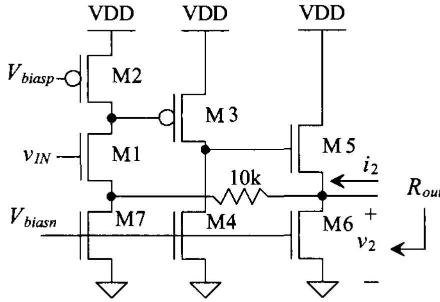

Figure 31.57 A series-shunt amplifier with source-follower output buffer.

31.14 Using the series-shunt amplifier shown in Fig. 31.57 and the results from problem 31.13, (a) draw the small-signal open-loop model for the circuit and (b) calculate the open-loop parameters, A_{OL}, β, R_i, and R_o and (c) the closed-loop parameters, A_{CL}, and R_{out}. Note that Fig. 31.57 is identical to Fig. 31.56 except that the resistors have been replaced with active loads.

31.15 Using the principles of feedback analysis, find the value of the voltage gain, $\frac{v_2}{v_{in}}$ and $\frac{v_2}{i_2}$ for the series-shunt circuit shown in Fig. 31.58.

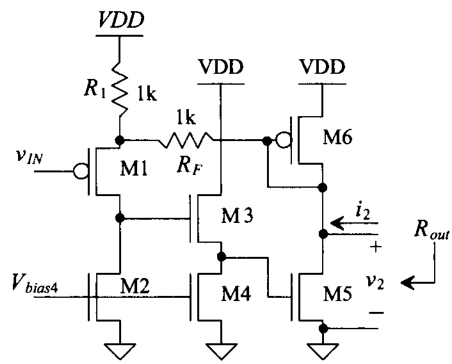

Figure 31.58 Feedback amplifier used in problem 31.15.

31.16 A shunt-shunt feedback amplifier is shown in Fig. 31.59. (a) Identify the feedback topology by labeling the input mixing variables and the output variables, (b) verify that negative feedback is employed, (c) draw the closed-loop small-signal model, and (d) find the values of $R_{\beta i}$ and $R_{\beta o}$.

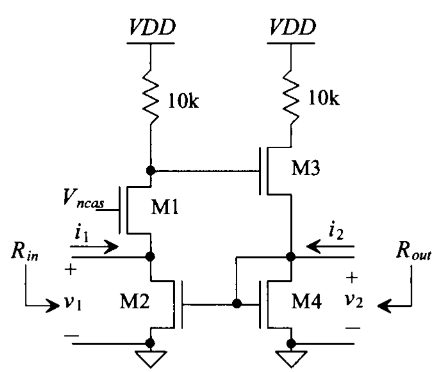

Figure 31.59 A shunt-shunt feedback amplifier.

31.17 Using the shunt-shunt amplifier shown in Fig. 31.59 and the results from problem 31.16, (a) draw the small-signal open-loop model for the circuit and (b) calculate expressions for the open-loop parameters, A_{OL}, β, R_i, and R_o and (c) the closed-loop parameters, A_{CL}, R_{in}, and R_{out}.

31.18 Using the principles of feedback analysis, find the value of the voltage gain, $\frac{v_2}{v_1}$, $\frac{v_1}{i_1}$, and $\frac{v_2}{i_2}$ for the shunt-shunt circuit shown in Fig. 31.60.

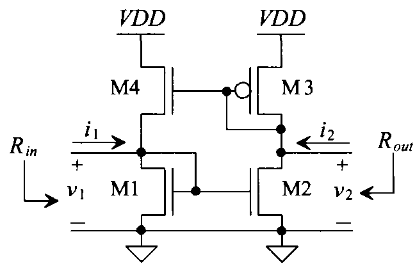

Figure 31.60 A shunt-shunt feedback amplifier, see problem 31.18.

31.19 Using the series-series feedback amplifier shown in Fig. 31.61, (a) identify the feedback topology, (b) verify that negative feedback is employed, (c) draw the closed-loop small-signal model, and (d) find the values of $R_{\beta i}$ and $R_{\beta o}$.

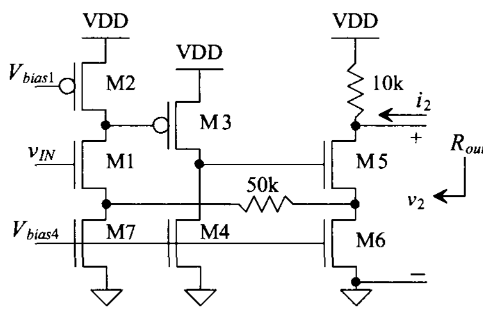

Figure 31.61 Series-series feedback amplifier with source-follower output buffer.

31.20 Using the series-series amplifier shown in Fig. 31.61 and the results from problem 31.19, (a) draw the small-signal open-loop model for the circuit and (b) calculate the open-loop parameters, A_{OL}, β, R_i, and R_o and (c) the closed-loop parameters, A_{CL}, R_{in}, and R_{out}.

31.21 Using the shunt-series amplifier in Fig. 31.35, derive the expressions for A_{OL}, R_{if}, and R_{of}.

31.22 Convert the shunt-shunt amplifier shown in Fig. 31.59 into a shunt-series feedback amplifier without adding any components. (a) Identify the feedback topology, (b) verify that negative feedback is employed, (c) draw the closed-loop small-signal model, and (d) find the values of $R_{\beta i}$ and $R_{\beta o}$.

31.23 Using the shunt-series amplifier from problem 31.22, (a) draw the small-signal open-loop model for the circuit and (b) calculate the open-loop parameters, A_{OL}, β, R_i, and R_o and (c) the closed-loop parameters, A_{CL}, R_{in}, and R_{out}.

31.24 A feedback amplifier is shown in Fig. 31.62. Identify the feedback topology and determine the value of the voltage gain, $\frac{v_2}{v_1}$, R_{in}, and R_{out}.

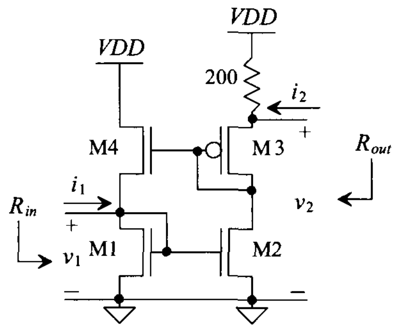

Figure 31.62 Feedback amplifier used in problem 31.24.

31.25 Notice that the amplifier shown in Fig. 31.63 is a simple common source amplifier with source resistance. Explain how this is actually a very simple feedback amplifier and determine the type of feedback used. Determine A_{OL} and β.

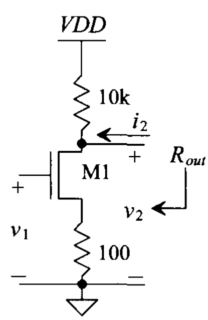

Figure 31.63 Common-source amplifier with source degeneration.

31.26 A feedback amplifier is shown in Fig. 31.64. Identify the feedback topology and determine the value of the voltage gain, $\frac{v_2}{v_1}$, R_{in}, and R_{out}.

31.27 A feedback amplifier is shown in Fig. 31.65. Identify the feedback topology and determine the value of the voltage gain, $\frac{v_2}{v_1}$ and R_{out}.

31.28 Prove that the expression for the open-loop gain derived in Ex. 31.3 is correct.

31.29 Determine if the amplifier seen in Fig. 31.44 is stable.

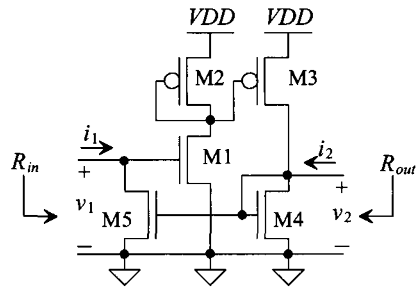

Figure 31.64 Feedback amplifier used in problem 31.26.

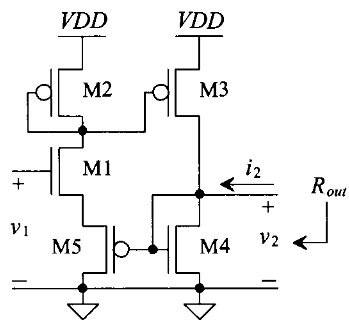

Figure 31.65 Feedback amplifier used in problem 31.27.

31.30 The op-amp shown in Fig. 31.66a can be modeled with the circuit of Fig. 31.66b. With a feedback factor, $\beta = 1$, determine if the op-amp is stable (and the corresponding phase and gain margins) for the following transfer function and $\omega_2 = 10^5$, 10^6, 10^7, and 5×10^6 rad/sec.

$$A_{OL}(j\omega) = \frac{10,000}{\left(1 + j\frac{\omega}{100}\right)\left(1 + j\frac{\omega}{\omega_2}\right)}$$

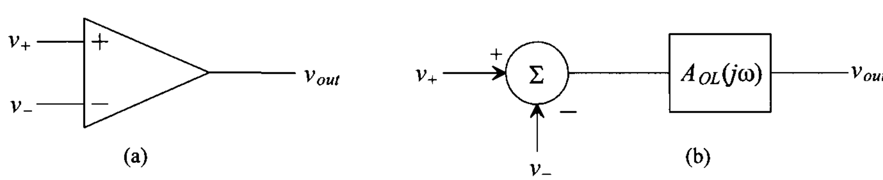

Figure 31.66 Modeling the op-amp.

31.31 The phase plot of an amplifier is shown in Fig. 31.67. The amplifier has a midband gain of −1,000 and 3 zeros at $\omega = \infty$ and three other unspecified poles. If the amplifier is configured in a feedback configuration and β is frequency independent, what is the exact value of β that would be necessary to cause the amplifier to oscillate?

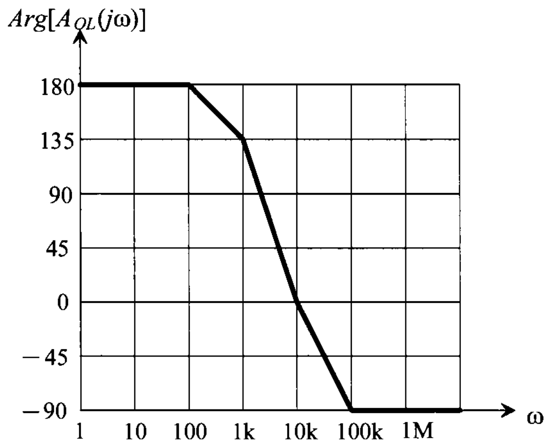

Figure 31.67 Phase response used in problem 31.31.

31.32 You have just measured the gain of the op-amp circuit shown in Fig. 31.68. You know from basic op-amp theory that the gain of the circuit should be $-R_2/R_1$ V/V. However, your measurements with R_2 = 10 kΩ and R_1 = 1 kΩ revealed that the gain was only −5 V/V. What is the open-loop gain of the op-amp?

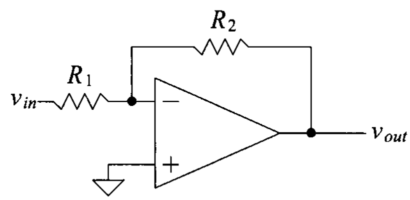

Figure 31.68 How finite open-loop gain effects closed-loop gain, problem 31.32.

31.33 Using the circuit shown in Fig. 31.69 and the *RR* method, find a value of R_1 and A_o which will cause the phase margin to equal 45° at ω = 8,000 rad/sec. The amplifier can be modeled as having an infinite input impedance and zero output resistance and has a frequency response of

$$A(s) = \frac{-A_o}{(s/200+1)(s/10{,}000+1)}$$

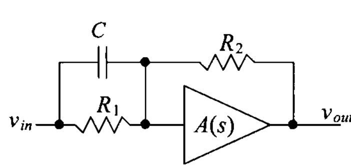

Figure 31.69 Amplifier used in problem 31.33.

31.34 Determine if the system with a return ratio as described by Eq. (31.102) is stable.

31.35 Redesign the transimpedance amplifier seen in Fig. 31.51 using the long-channel CMOS process discussed in this book.

31.36 How would the input-referred noise for the TIA design presented in Sec. 31.9.2 be reduced?