Design of Analog CMOS Integrated Circuits

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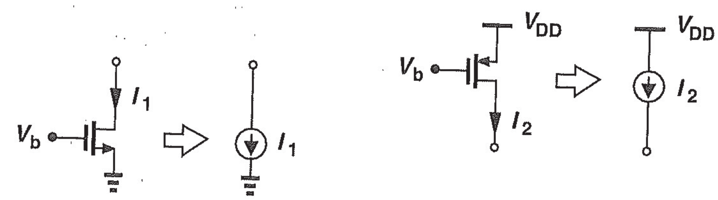

Figure 2.17 Saturated MOSFETs operating as current sources.

With the approximation $L \approx L'$, a saturated MOSFET can be used as a current source connected between the drain and the source (Fig. 2.17), an important component in analog design. Note that the current sources inject current into ground or draw current from V_{DD}. In other words, only one terminal of each current source is "floating."

Since a MOSFET operating in saturation produces a current in response to its gate-source overdrive voltage, we may define a figure of merit that indicates how well a device converts a voltage to a current. More specifically, since in processing signals we deal with the *changes* in voltages and currents, we define the figure of merit as the change in the drain current divided by the change in the gate-source voltage. Called the "transconductance" and denoted by g_m, this quantity is expressed as:

$$g_m = \left.\frac{\partial I_D}{\partial V_{GS}}\right|_{V_{DS},\text{const.}} \tag{2.16}$$

$$= \mu_n C_{ox}\frac{W}{L}(V_{GS} - V_{TH}). \tag{2.17}$$

In a sense, g_m represents the sensitivity of the device: for a high g_m, a small change in V_{GS} results in a large change in I_D. Interestingly, g_m in the saturation region is equal to the inverse of R_{on} in deep triode region.

The reader can prove that g_m can also be expressed as

$$g_m = \sqrt{2\mu_n C_{ox}\frac{W}{L}I_D} \tag{2.18}$$

$$= \frac{2I_D}{V_{GS} - V_{TH}}. \tag{2.19}$$

Plotted in Fig. 2.18, each of the above expressions proves useful in studying the behavior of g_m as a function of one parameter while other parameters remain constant. For example, (2.17) suggests that g_m increases with the overdrive if W/L is constant whereas (2.19) implies that g_m decreases with the overdrive if I_D is constant. The concept of transconductance

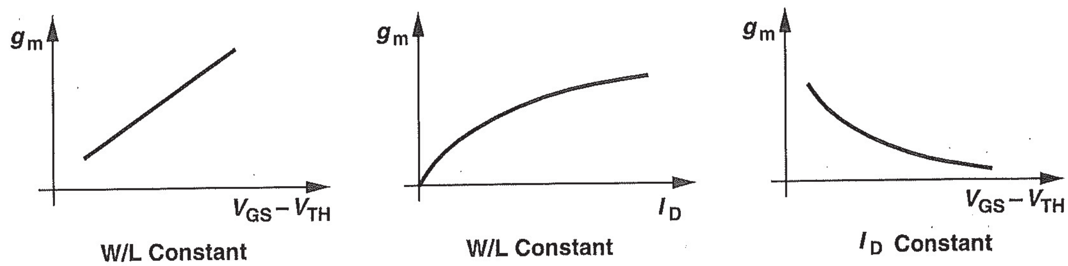

Figure 2.18 MOS transconductance as a function of overdrive and drain current.

can also be applied to a device operating in the triode region, as illustrated in the following example.

Example 2.2

For the arrangement shown in Fig. 2.19, plot the transconductance as a function of V_{DS}.

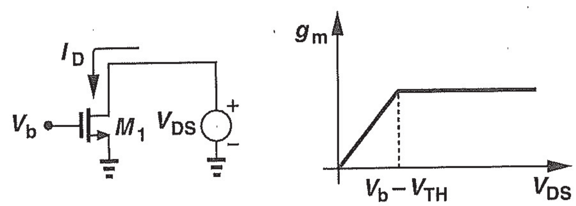

Figure 2.19

Solution

It is simpler to study g_m as V_{DS} decreases from infinity. So long as $V_{DS} \geq V_b - V_{TH}$, M_1 is in saturation, I_D is relatively constant, and, from (2.18), so is g_m. For $V_{DS} < V_b - V_{TH}$, M_1 is in the triode region and:

$$g_m = \frac{\partial}{\partial V_{GS}} \left\{ \frac{1}{2} \mu_n C_{ox} \frac{W}{L} \left[2(V_{GS} - V_{TH}) V_{DS} - V_{DS}^2 \right] \right\} \tag{2.20}$$

$$= \mu_n C_{ox} \frac{W}{L} V_{DS}. \tag{2.21}$$

Thus, as plotted in Fig. 2.19, the transconductance drops if the device enters the triode region. For amplification, therefore, we usually employ MOSFETs in saturation.

The distinction between saturation and triode regions can be confusing, especially for PMOS devices. Intuitively, we note that the channel is pinched off if the difference between the gate and drain voltages is not sufficient to create an inversion layer. As depicted conceptually in Fig. 2.20, as $V_G - V_D$ of an NFET drops below V_{TH}, pinch-off occurs. Similarly, We now re-examine Eq. (2.18) for the transconductance of a MOS device operating in the subthreshold region. Is it possible to achieve an arbitrarily high transconductance by increasing W while maintaining I_D constant? Is it possible to obtain a *higher* transconductance than that of a bipolar transistor (I_C/V_T) biased at the same current? Equation (2.18) was derived from the square-law characteristics $I_D = (1/2)\mu_n C_{ox}(W/L)(V_{GS} - V_{TH})^2$. However, if W increases while I_D remains constant, then $V_{GS} \to V_{TH}$ and the device enters the subthreshold region. As a result, the transconductance is calculated from (2.30) to be $g_m = I_D/(\zeta V_T)$, revealing that MOSFETs are inferior to bipolar transistors in this respect.

The exponential dependence of I_D upon V_{GS} in subthreshold operation may suggest the use of MOS devices in this regime so as to achieve a higher gain. However, since such conditions are met by only a large device width or low drain current, the speed of subthreshold circuits is severely limited.

Voltage Limitations MOSFETs experience various breakdown effects if their terminal voltage differences exceed certain limits. At high gate-source voltages, the gate oxide breaks down irreversibly, damaging the transistor. In short-channel devices, an excessively large drain-source voltage widens the depletion region around the drain so much that it touches that around the source, creating a very large drain current. (This effect is called "punchthrough.") Other limitations relate to "hot electron effects" and are described in Chapter 16.

2.4 MOS Device Models

2.4.1 MOS Device Layout

For the developments in subsequent sections, it is beneficial to have some understanding of the layout of a MOSFET. We describe only a simple view here, deferring the fabrication details and structural subtleties to Chapters 17 and 18.

The layout of a MOSFET is determined by both the electrical properties required of the device in the circuit and the "design rules" imposed by the technology. For example, W/L is chosen to set the transconductance or other circuit parameters, while the minimum L is dictated by the process. In addition to the gate, the source and drain areas must be defined properly as well.

Shown in Fig. 2.28 are the "bird eye's view" and the top view of a MOSFET. The gate polysilicon and the source and drain terminals are typically tied to metal (aluminum) wires that serve as interconnects with low resistance and capacitance. To accomplish this, one or more "contact windows" must be opened in each region, filled with metal, and connected to the upper metal wires. Note that the gate poly extends beyond the channel area by some amount to ensure reliable definition of the "edge" of the transistor.

The source and drain junctions play an important role in the performance. To minimize the capacitance of S and D, the total area of each junction must be minimized. We see from Fig. 2.28 that one dimension of the junctions is equal to W. The other dimension must be large enough to accommodate the contact windows and is specified by the technology design rules.[7]

[7]This dimension is typically three to four times the minimum allowable channel length.

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Chapter 3

Single-Stage Amplifiers

Amplification is an essential function in most analog (and many digital) circuits. We amplify an analog or digital signal because it may be too small to drive a load, overcome the noise of a subsequent stage, or provide logical levels to a digital circuit. Amplification also plays a critical role in feedback systems (Chapter 8).

In this chapter, we study the low-frequency behavior of single-stage CMOS amplifiers. Analyzing both the large-signal and the small-signal characteristics of each circuit, we develop intuitive techniques and models that prove useful in understanding more complex systems. An important part of a designer's job is to use proper approximations so as to create a simple mental picture of a complicated circuit. The intuition thus gained makes it possible to formulate the behavior of most circuits by inspection rather than by lengthy calculations.

Following a brief review of basic concepts, we describe in this chapter four types of amplifiers: common-source and common-gate topologies, source followers, and cascode configurations. In each case, we begin with a simple model and gradually add second-order phenomena such as channel-length modulation and body effect.

3.1 Basic Concepts

The input-output characteristic of an amplifier is generally a nonlinear function (Fig. 3.1) that can be approximated by a polynomial over some signal range:

$$y(t) \approx \alpha_0 + \alpha_1 x(t) + \alpha_2 x^2(t) + \cdots + \alpha_n x^n(t) \quad x_1 \leq x \leq x_2. \tag{3.1}$$

The input and output may be current or voltage quantities. For a sufficiently narrow range of x,

$$y(t) \approx \alpha_0 + \alpha_1 x(t), \tag{3.2}$$

where α_0 can be considered the operating (bias) point and α_1 the small-signal gain. So long as $\alpha_1 x(t) \ll \alpha_0$, the bias point is disturbed negligibly, (3.2) provides a reasonable

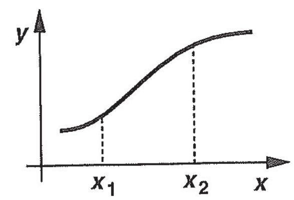

Figure 3.1 Input-output characteristic of a nonlinear system.

approximation, and higher order terms are insignificant. In other words, $\Delta y = \alpha_1 \Delta x$, indicating a linear relationship between the *increments* at the input and output. As $x(t)$ increases in magnitude, higher order terms manifest themselves, leading to nonlinearity and necessitating large-signal analysis. From another point of view, if the slope of the characteristic (the incremental gain) *varies* with the signal level, then the system is nonlinear. These concepts are described in detail in Chapter 13.

What aspects of the performance of an amplifier are important? In addition to gain and speed, such parameters as power dissipation, supply voltage, linearity, noise, or maximum voltage swings may be important. Furthermore, the input and output impedances determine how the circuit interacts with preceding and subsequent stages. In practice, most of these parameters trade with each other, making the design a multi-dimensional optimization problem. Illustrated in the "analog design octagon" of Fig. 3.2, such trade-offs present many challenges in the design of high-performance amplifiers, requiring intuition and experience to arrive at an acceptable compromise.

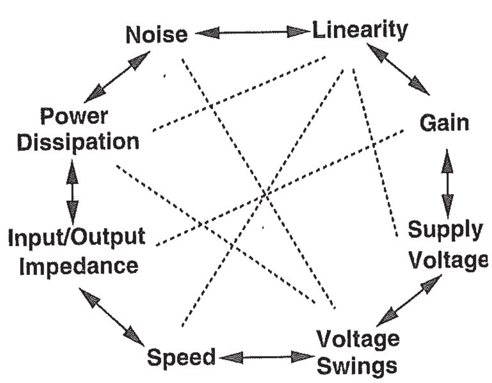

Figure 3.2 Analog design octagon.

3.2 Common-Source Stage

3.2.1 Common-Source Stage with Resistive Load

By virtue of its transconductance, a MOSFET converts variations in its gate-source voltage to a small-signal drain current, which can pass through a resistor to generate an output voltage. Shown in Fig. 3.3(a), the common-source (CS) stage performs such an operation.

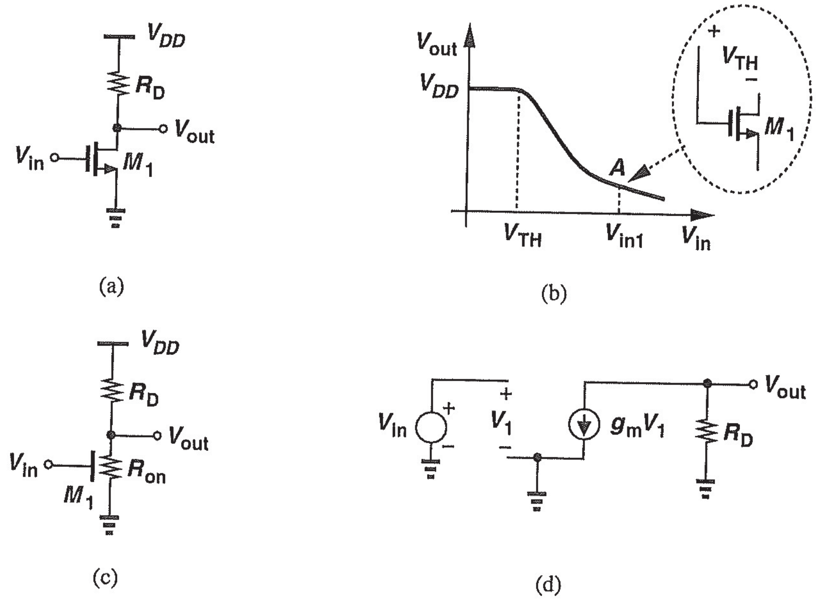

Figure 3.3 (a) Common-source stage, (b) input-output characteristic, (c) equivalent circuit in deep triode region, (d) small-signal model for the saturation region.

We study both the large-signal and the small-signal behavior of the circuit. Note that the input impedance of the circuit is very high at low frequencies.

If the input voltage increases from zero, M_1 is off and $V_{out} = V_{DD}$ [Fig. 3.3(b)]. As V_{in} approaches V_{TH}, M_1 begins to turn on, drawing current from R_D and lowering V_{out}. If V_{DD} is not excessively low, M_1 turns on in saturation, and we have

$$V_{out} = V_{DD} - R_D \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in} - V_{TH})^2, \tag{3.3}$$

where channel-length modulation is neglected. With further increase in V_{in}, V_{out} drops more and the transistor continues to operate in saturation until V_{in} exceeds V_{out} by V_{TH} [point A in Fig. 3.3(b)]. At this point,

$$V_{in1} - V_{TH} = V_{DD} - R_D \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{in1} - V_{TH})^2, \tag{3.4}$$

from which $V_{in1} - V_{TH}$ and hence V_{out} can be calculated.

For $V_{in} > V_{in1}$, M_1 is in the triode region:

$$V_{out} = V_{DD} - R_D \frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left[2(V_{in} - V_{TH})V_{out} - V_{out}^2\right]. \tag{3.5}$$

Now suppose we apply feedback to the amplifier such that the gain and bandwidth are modified to 10 and 100 MHz, respectively. Placing two of these amplifiers in a cascade [Fig. 8.10(b)], we obtain a much faster response with an overall gain of 100. Of course, the cascade consumes twice as much power, but it would be quite difficult to achieve this performance by the original amplifier even if its power dissipation were doubled.

Nonlinearity Reduction A very important property of negative feedback is the suppression of nonlinearity in analog circuits. We defer the study of this effect to Chapter 13.

8.1.2 Types of Amplifiers

Most of the circuits studied thus far can be considered "voltage amplifiers" because they sense a voltage at the input and produce a voltage at the output. However, three other types of amplifiers can also be constructed such that they sense or produce currents. Shown in Fig. 8.11, the four configurations have quite different properties: (1) circuits sensing

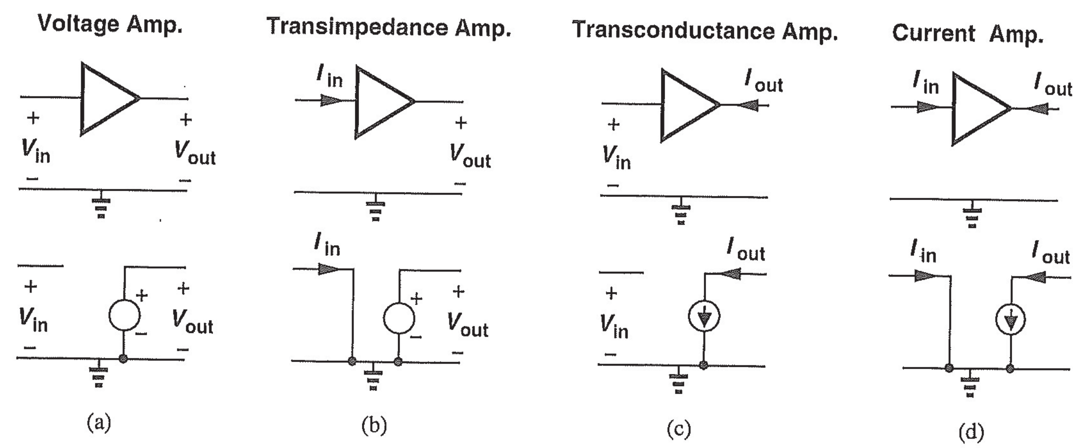

Figure 8.11 Types of amplifiers along with their idealized models.

a voltage must exhibit a high input impedance (as a voltmeter) whereas those sensing a current must provide a low input impedance (as a current meter); (2) circuits generating a voltage must exhibit a low output impedance (as a voltage source) while those generating a current must provide a high output impedance (as a current source). Note that the gains of transimpedance and transconductance[4] amplifiers have a dimension of resistance and conductance, respectively. For example, a transimpedance amplifier may have a gain of 2 kΩ, which means it produces a 2-V output in response to a 1-mA input. Also, we use the sign conventions depicted in Fig. 8.11, for example, the transimpedance $R_0 = V_{out}/I_{in}$ if I_{in} flows *into* the amplifier.

[4]This terminology is standard but not consistent. One should use either transimpedance and transadmittance or transresistance and transconductance.

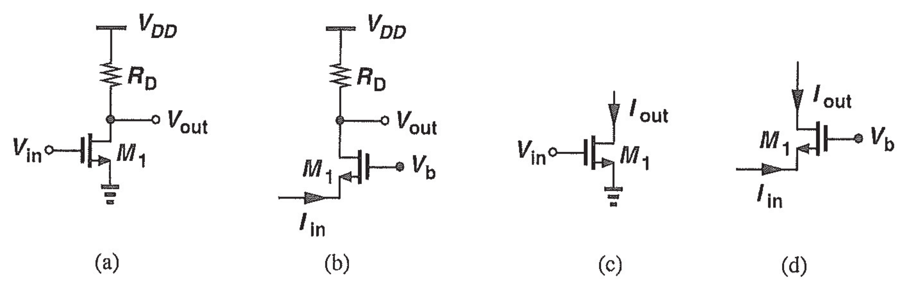

Figure 8.12 Simple implementations of four types of amplifiers.

Figure 8.12 illustrates simple implementations of each amplifier. In Fig. 8.12(a), a common-source stage senses and produces voltages and in Fig. 8.12(b), a common-gate circuit serves as a transimpedance amplifier, converting the source current to a voltage at the drain. In Fig. 8.12(c), a common-source transistor operates as a transconductance amplifier, generating an output current in response to an input voltage, and in Fig. 8.12(d), a common-gate device senses and produces currents.

The circuits of Fig. 8.12 may not provide adequate performance in many applications. For example, the circuits of Figs. 8.12(a) and (b) suffer from a relatively high output impedance. Fig. 8.13 depicts modifications that alter the output impedance or increase the gain.

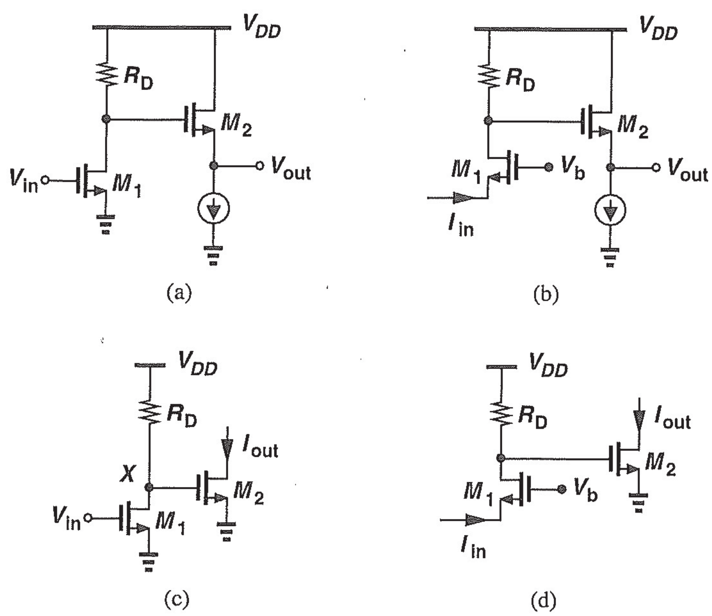

Figure 8.13 Four types of amplifiers with improved performance.

Example 8.1

Calculate the gain of the transconductance amplifier shown in Fig. 8.13(c).

Solution

The gain in this case is defined as $G_m = I_{out}/V_{in}$. That is,

$$G_m = \frac{V_X}{V_{in}} \cdot \frac{I_{out}}{V_X} \tag{8.22}$$

$$= -g_{m1}(r_{O1} \| R_D) \cdot g_{m2}. \tag{8.23}$$

While most familiar amplifiers are of voltage-voltage type, the other three configurations do find usage. For example, transimpedance amplifiers are an integral part of optical fiber receivers because they must sense the current produced by a photodiode, eventually generating a voltage that can be processed by subsequent circuits.

8.1.3 Sense and Return Mechanisms

Placing a circuit in a feedback loop requires sensing the output signal and returning (a fraction) of the result to the summing node at the input. With voltage or current quantities as input and output signals, we can identify four types of feedback: voltage-voltage, voltage-current, current-current, and current-voltage, where the first entry in each case denotes the quantity sensed at the *output* and the second the type of signal returned to the input.[5]

It is instructive to review methods of sensing and summing voltages or currents. To sense a voltage, we place a voltmeter *in parallel* with the corresponding port [Fig. 8.14(a)], ideally introducing no loading. When used in a feedback system, this type of sensing is also called "shunt feedback."

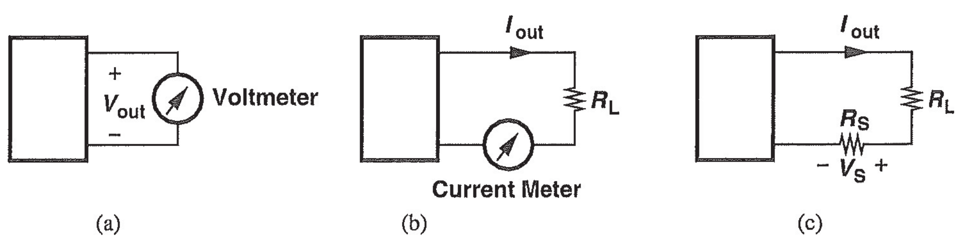

Figure 8.14 Sensing (a) a voltage by a voltmeter, (b) a current by a current meter, (c) a current by a small resistor.

To sense a current, a current meter is inserted *in series* with the signal [Fig. 8.14(b)], ideally exhibiting zero series resistance. Thus, this type of sensing is also called "series feedback." In practice, a small resistor replaces the current meter [Fig. 8.14(c)], with the voltage drop across the resistor serving as a measure of the output current.

The addition of the feedback signal and the input signal can be performed in the voltage domain or current domain. To add two quantities, we place them in series if they are

[5]Different authors use different orders or terminologies for the four types of feedback.

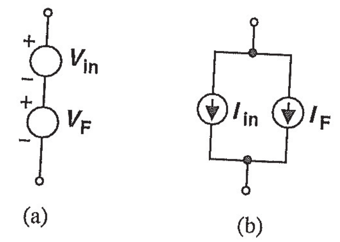

Figure 8.15 Addition of (a) voltages and (b) currents.

voltages and in parallel if they are currents (Fig. 8.15). While ideally having no influence on the operation of the open-loop amplifier itself, the feedback network in reality introduces loading effects that must be taken into account. This issue is discussed in Section 8.3.

To visualize the methods of Figs. 8.14 and 8.15, we consider a number of practical implementations. A voltage can be sensed by a resistive (or capacitive) divider in parallel with the port [Fig. 8.16(a)] and a current by placing a resistor in series with the wire and sensing

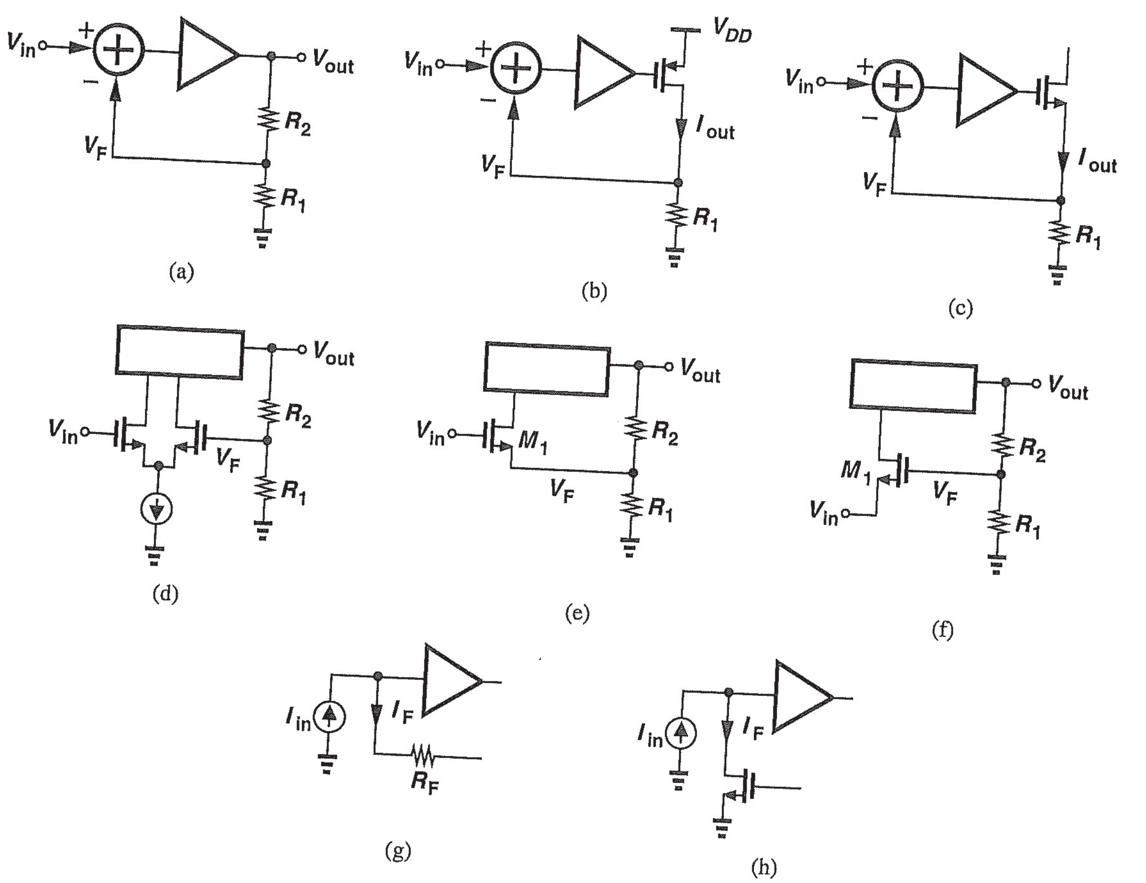

Figure 8.16 Practical means of sensing and adding voltages and currents.