SIMPLE TRANSCONDUCTANCE AMPLIFIER

The schematic for the transconductance amplifier is shown in Figure 5.3. The circuit consists of a differential pair and a single current mirror, like the one shown in Figure 3.9 (p. 40), which is used to subtract the drain currents I_1 and I_2. The current I_1 drawn out of Q3 is reflected as an equal current out of Q4; the output current is thus equal to $I_1 - I_2$, and is therefore given by Equation 5.5.

We can measure the current out of the amplifier as a function of the input voltages using the setup shown in Figure 5.4. We are using a current meter with its primary input connected to the amplifier output and its *reference input* connected to a voltage source V_{out}. A perfect current meter has zero resistance; real current meters have sophisticated feedback arrangements to make their input resistances very small. For that reason, the voltage on the output node of the amplifier will be V_{out}. For now, we will simply set V_{out} in the midrange between V_{DD} and ground. In later sections, we will investigate the effect of V_{out} on the performance of the circuit.

The current out of the simple amplifier is plotted as a function of $V_1 - V_2$ in Figure 5.5. The curve is very close to a tanh, as expected. We can determine the effective value of $kT/(q\kappa)$ by extrapolating the slope of the curve at

FIGURE 5.3 Schematic diagram of the simple transconductance amplifier. The current mirror formed by Q3 and Q4 is used to form the output current, which is equal to $I_1 - I_2$. The symbol used for the circuit is shown in the inset.

from "Analog VLSI and Neural Systems" by C. Mead

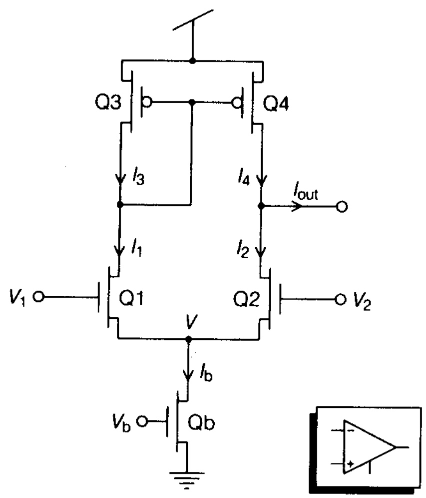
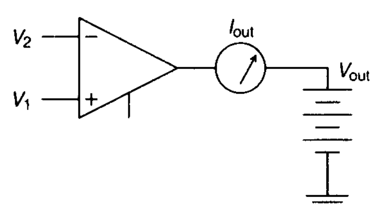


FIGURE 5.4 Arrangement for measuring the output current of the transconductance amplifier. The circuit of Figure 5.3 is represented symbolically.



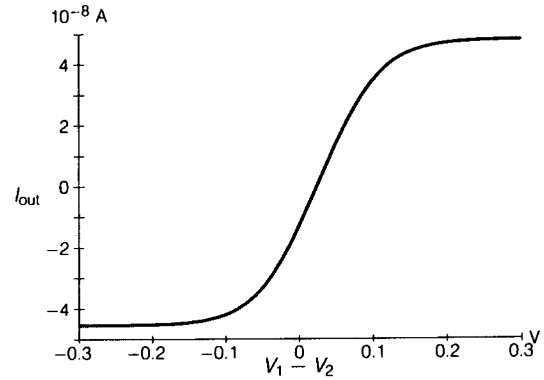


FIGURE 5.5 Output current of the transconductance amplifier as a function of differential input voltage. The mismatch between transistor characteristics can be seen in two ways. For this particular amplifier, the input offset voltage is approximately 25 millivolts, typical for a digital CMOS process. The limiting current for positive inputs is approximately 6 percent larger than that for negative inputs; a more typical variation would be 20 percent.

the origin to the two asymptotes. The difference between the positive and negative intercepts should be $4kT/(q\kappa)$. Using this procedure on Figure 5.5, we obtain the result that $kT/(q\kappa)$ is approximately 43 millivolts, giving κ as approximately 0.58. This value is in good agreement with that obtained from the voltage dependence of the saturation current (Figure 3.7 (p. 38)). The **transconductance** G_m of the amplifier is just the slope of the tanh in Equation 5.5 at the origin. For Figure 5.5, the output current changes 5.6×10^{-8} amp for a 100-millivolt change in $V_{in} = V_1 - V_2$. G_m is therefore 5.6×10^{-7} mho. In terms of the circuit variables,

$$G_m = \frac{\partial I_{out}}{\partial V_{in}} = \frac{I_b}{2kT/(q\kappa)} \tag{5.6}$$

Notice that the transconductance is proportional to the bias current I_b, a fact that will become important when the differential circuit is used to produce a voltage-type output, or as part of a multiplier.

The layout of a typical implementation of the simple amplifier is shown in Plate 7(a).

CIRCUIT LIMITATIONS

Now that we know how an ideal transconductance amplifier works, we can investigate the limitations and imperfect behavior of such circuits in the real world. Deviations from ideal behavior are of two basic sorts:

1. Mismatch between transistors

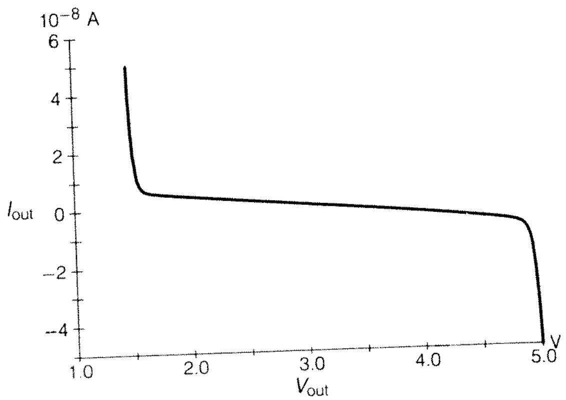


FIGURE 5.6 Dependence of output current of the simple transconductance amplifier on output voltage. The exponential dependencies at the ends are due to the high (V_{DD}) and low (V_{min}) limits discussed in the text. The slope in the midrange, due to the nonzero drain conductance of the output transistors Q2 and Q4, is the output conductance of the amplifier. The curve shown is for V_1 approximately equal to V_2; input-voltage differences cause the entire curve to shift up or down, according to the relationship shown in Figure 5.5.

is positive, approximately equal to I_4. As we lower the output voltage, all is well until V_{out} decreases to less than V, after which the output node becomes the *source* of Q2, and the V node becomes the *drain*. The interchange of source and drain of Q2 results in a reversal of current through Q2—I_2 becomes negative instead of positive. The reversal occurs when V_{out} is equal to $\kappa(V_1-V_b)$, but is not noticeable in the output current until V_{out} is approximately equal to $\kappa(V_2-V_b)$, where I_2 becomes comparable with I_1. A further decrease in output voltage results in an exponential increase in I_{out}, because the gate–source voltage of Q2 is increasing. This negative I_2 is supplied by an increase in I_1, which results in an equal increase in output current through Q4. The output current thus increases from two equal contributions of the same sign.

If V_2 is greater than V_1 by several $kT/(q\kappa)$, the same effect can be observed. The output current is negative, and V is equal to $\kappa(V_2-V_b)$. As we decrease the output voltage, we make the voltage between the source and the drain of Q2 smaller and smaller, Q2 comes out of saturation, and V begins to decrease. As both V_{out} and V decrease, the gate–source voltage of Q2 increases, causing Q2 to conduct more current. The voltage V follows V_{out} more and more closely. There is no noticeable change in output current, however, until V approaches $\kappa(V_1-V_b)$, at which point the current through Q1 becomes comparable to I_b. As we decrease the voltage at the output node further, I_1 exceeds I_b, and V does not decrease as fast as does V_{out}. Once V is greater than V_{out}, the drain and source of Q2 are interchanged, and the situation is exactly as it was for V_2 greater than V_1. Transistor Q2 starts siphoning charge away from the V node, and the output current increases exponentially.

We call the limitation on the operation of the simple transconductance amplifier imposed by this behavior the "V_{min} problem." We can express the minimum output voltage as

$$V_{min} = \kappa(\min(V_1, V_2) - V_b) \tag{5.7}$$

In other words, the amplifier will work with its output voltage up to nearly V_{DD}, and down to V_b below the lowest input signal that we have applied to it, but not lower than that.

We run into two walls, one on the top and one on the bottom. The wall on the top side is not serious; all it does is to prevent us from going right up to V_{DD}. When V_1 is greater than V_2, the current comes out of Q4—so, if we make the output node equal to V_{DD}, we will not get any current out. We cannot quite work up against the rail, but we can get very close. The upper limit on the output voltage is set by the saturation properties of Q4. As long as we stay a few $kT/(q\kappa)$ below V_{DD}, we are fine.

The bottom V_{min} limit is much more serious. It is the biggest problem with this circuit. It forms a *hard limit* below which the circuit does not work, and that limit depends on the input voltage.

VOLTAGE OUTPUT

We call these circuits "transconductance amplifiers" because that is the way in which they are usually used. They also can be used, however, to take a difference in voltage at the input and turn it into a voltage at the output. Instead of measuring I_{out} with an ammeter, we measure V_{out} with a voltmeter. The drain conductances of Q2 and Q4 are used to convert the output current into an output voltage.

The drain current of a transistor is not completely independent of its drain voltage, even in saturation. There is a finite slope of I_d versus V_d given by the **Early effect**, discussed in Appendix B. This effect is responsible for the dependence of output current on output voltage seen between the two limits in Figure 5.6.

The finite output conductance of the circuit can be used to convert the current-type output signal into a voltage-type signal. We take away the V_{out} voltage source completely, and use a voltmeter instead of an ammeter. An ideal voltmeter draws no current from the circuit it is measuring. It is an ideal open circuit, having zero input conductance (infinite input resistance). The input conductances of real voltmeters vary over many orders of magnitude. A good electrometer has an input resistance greater than 10^{12} ohm; ordinary voltmeters can have resistances many orders of magnitude less. A typical oscilloscope has an input resistance of 10^6 ohm. We must always check that the current drawn by our measuring equipment is small compared with the current flowing in the circuit. Using a good voltmeter, we can observe the voltage at which the output current is equal to zero. That is the **open-circuit output voltage** of the device

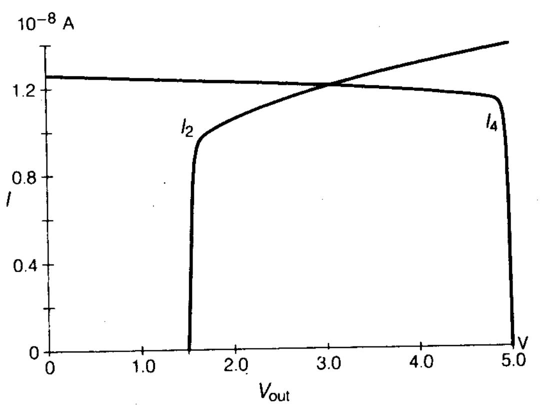


FIGURE 5.7 Current out of the n-channel (Q2) and p-channel (Q4) output transistors of a simple transconductance amplifier, as a function of output voltage, for V_1 approximately equal to V_2. The output current is the difference between the two curves; therefore, the open-circuit output voltage is given by the intersection of these curves.

Consider the combination of Q4 and Q2. In Figure 5.7, we plot the magnitude of I_4 and I_2 versus the output voltage. Current is flowing down from V_{DD} to ground—I_4 is flowing *into* the output node, and I_2 is flowing *out of* the output node. For any given output voltage, the output current is the difference between the two curves. For any particular input voltages, the value of V_{min} will be somewhat below the lowest input—for the operating conditions shown in Figure 5.7, I_2 goes to zero when V_{out} is approximately equal to V_{min}. Because Q4 is a p-channel device, its drain voltage is plotted downward from V_{DD}. The open-circuit voltage is the value at which the two currents are equal; *open-circuit* means *no current.*

We can easily see what the output voltage would do as a function of a difference in the input voltages, if the drain curves for Q2 and Q4 were absolutely flat. When V_1 is a little less than V_2, the output voltage will decrease to nearly V_{min}; when V_1 is a little bigger than V_2, the voltage will increase to almost V_{DD}. The experimental dependence of V_{out} on V_1, for several values of V_2, is shown in Figure 5.8. The output voltage stays at 5 volts until V_1 gets very close to V_2, then it drops rapidly to V_{min}. The sloping line where all curves merge is thus V_{min}.

Instead of an infinite slope in the transition region, which would correspond to infinite gain, which corresponds to output transistors that have zero drain conductance, the actual circuit has a finite slope as a result of its real transistors, which have finite drain conductance.

Voltage Gain

The **voltage gain** A is defined as $\partial V_{out}/\partial V_{in}$, where V_{in} is equal to $V_1 - V_2$. An enlargement of the steep part of the V_2 at 2.5 volt curve in Figure 5.8 is

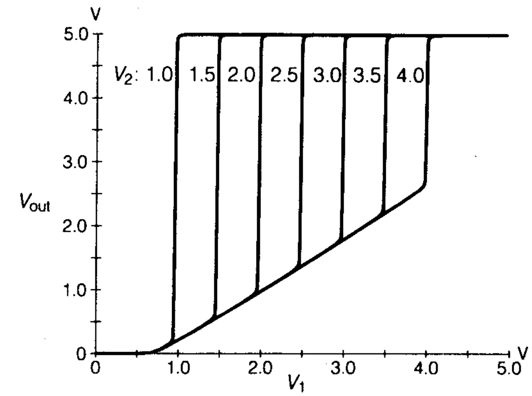


FIGURE 5.8 Open-circuit output voltage of the simple transconductance amplifier as a function of V_1, for several values of V_2. The steep part of the curves occur at V_1 approximately equal to V_2. The sloping lower limit of the output is due to the V_{min} problem.

shown in Figure 5.9. The maximum gain is approximately 143. We can easily compute what the gain of this circuit should be by considering the properties of the output transistors.

An enlargement of the intersection of the Q2 and Q4 drain curves in Figure 5.7 is shown in Figure 5.10. For a certain input-voltage difference, the curves are marked I_2 and I_4. When the input-voltage difference is increased by Δv, both curves change: I_2 decreases to I_2', and I_4 increases to I_4'. Because the bias current I_b is constant, an increase ΔI in I_4 due to a change in input voltage will result in an equal decrease ΔI in I_2, as shown. The total change in output current per unit change in input-voltage difference was defined in Equation 5.6 as the transconductance G_m of the circuit.

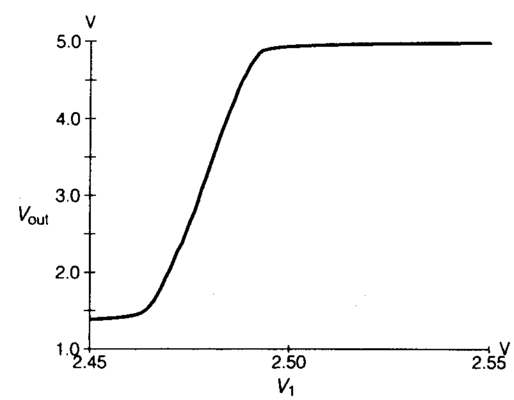


FIGURE 5.9 Expanded view of the center curve (V_2 equal to 2.5 volts) of Figure 5.8.

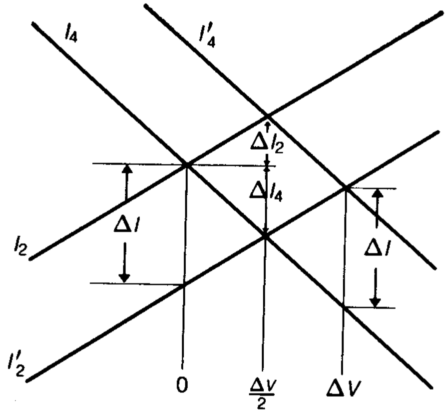


FIGURE 5.10 Expanded view of the intersection of the two curves of Figure 5.7 for two slightly different input voltages. The dependence of output voltage on transistor currents and drain conductances is obtained from the construction shown.

From Figure 5.6, we have seen that the output current *decreases* as the output voltage *increases*. The decrease in the output current per unit change in the output voltage is called the **output conductance** of the circuit.

$$G_{\text{out}} = -\frac{\partial I_{\text{out}}}{\partial V_{\text{out}}} = \frac{\partial I_2}{\partial V_{\text{out}}} - \frac{\partial I_4}{\partial V_{\text{out}}} \tag{5.8}$$

This quantity is the negative of the slope of the central region of Figure 5.6. The output conductance is just the sum of the contributions due to the two output transistors. In Figure 5.6, the output current changes 3.6×10^{-9} amp for a 1 volt change in output voltage. Thus, G_{out} is approximately 3.6×10^{-9} mho.

When the output is open-circuited, the total *increase* $2\Delta I$ in output current due to an increase Δv in input voltage difference is compensated by an equal *decrease* in output current due to the increase ΔV in the output voltage.

$$2\Delta I = \frac{\partial I_{\text{out}}}{\partial V_{\text{in}}}\Delta v = -\frac{\partial I_{\text{out}}}{\partial V_{\text{out}}}\Delta V \tag{5.9}$$

Substituting Equation 5.6 and Equation 5.8 into Equation 5.9, we obtain the open-circuit voltage gain $A = \Delta V/\Delta v$.

$$A = \frac{G_{\text{m}}}{G_{\text{out}}}$$

From the values of G_{m} and G_{out} measured on the simple transconductance circuit, the open-circuit voltage gain should be

$$\frac{5.6 \times 10^{-7}}{3.6 \times 10^{-9}} = 148$$

The result is in good agreement with the measured gain of 143.

In Appendix B, we derive the slope of the transistor drain curve in saturation. The slope of the saturated part of the drain curves of a given transistor has a slope proportional to the current level. In other words, the intercept on the V_{ds} axis occurs at a voltage V_0 that is approximately independent of the absolute current level:

$$\frac{\partial I_{\text{sat}}}{\partial V_{\text{ds}}} \approx \frac{I_{\text{sat}}}{V_0}$$

V_0 is called the **Early voltage** for the transistor. For open-circuit operation, the steep part of the curves of Figure 5.8 will occur when the current through the two output transistors is equal, and therefore is half of the bias current I_b. Thus, *the output conductance of the amplifier is proportional to the bias current* I_b. The transconductance also is proportional to the bias current (Equation 5.6).

Because both G_{m} and G_{out} are proportional to the bias current, the explicit dependence on current level cancels out and *the voltage gain is independent of bias current.* We can express the $\partial I/\partial V$ terms in Equation 5.8 in terms of the V_0 values (V_{N} for Q2 and V_{P} for Q4):

$$\frac{1}{A} = \left(\frac{1}{V_{\text{N}}} + \frac{1}{V_{\text{P}}}\right)\frac{2}{\kappa} \tag{5.10}$$

where the Vs are expressed in kT/q units.

Equation 5.10 allows us to compute the gain of any output stage composed of complementary p- and n-channel transistors. In Appendix B, we note that the V_0 of a given transistor is proportional to its length. Hence, we can make the gain arbitrarily high by committing a large silicon area to long output transistors. For typical 1988 processes, output transistors 20 microns long will give a voltage gain of approximately 2000.

WIDE-RANGE AMPLIFIERS

A simple transconductance amplifier will not generate output voltages below V_{min}, which, in turn, is dependent on the input voltages. This limitation often is a source of problems at the system level, because it is not always possible to restrict the range of input voltages. We can remove this restriction, however, by a simple addition to the circuit, as shown in Figure 5.11.

Instead of feeding the output directly, the drain of Q2 is connected to the current mirror formed by Q5 and Q6. The currents coming out of Q4 and Q6 now are just the two halves of the current in the differential pair. We then reflect the Q6 current one more time, through Q7 and Q8, and subtract it from I_4 to form the output. As in the simple circuit, the output current is just the difference between I_1 and I_2.

The major advantage of the wide-range amplifier over the simple circuit is that both input and output voltages can run almost up to V_{DD} and almost down to ground, without affecting the operation of the circuit. In other words, we have eliminated the V_{min} problem.

The other nice thing about this circuit is that the current mirrors, such as Q3 and Q5, hold the drain voltages of Q1 and Q2 very nearly constant. In

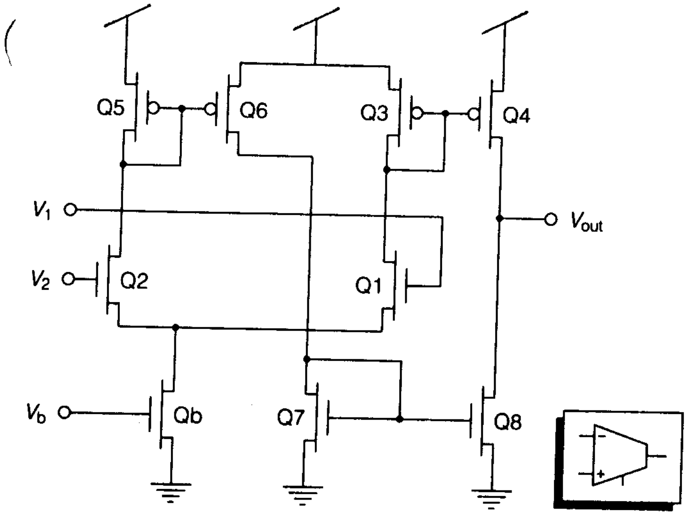


FIGURE 5.11 Schematic of the wide-range transconductance amplifier. This circuit has many advantages over the simple transconductance amplifier of Figure 5.3. The output-voltage range is not affected by the input voltages. The symbol for the circuit is shown in the inset.

diode-connected transistors, the current increases exponentially with the gate voltage, so the drain voltages never get very far below V_{DD}. For that reason, Q2 no longer has a problem associated with its drain conductance; its source–drain voltage is nearly equal to that of Q1. So the drain conductances of Q1 and Q2 are not critical in this circuit. The same thing is true of Q6: Q7 is a diode-connected transistor; it holds the drain voltage of Q6 very nearly constant. The only transistors that work over a large voltage range are Qb, Q4, and Q8, and we can make their channels long to get a low drain conductance (output current that is *nearly independent* of output voltage). Because of their low output conductance, long Q4 and Q8 transistors give the circuit a high voltage gain.

A layout of the improved circuit is shown in Plate 7(b). This new circuit is about twice the size of the simple transconductance amplifier of Plate 7(a). Such wide-range amplifiers have about 10 times the gain of the simple amplifier, and they work all the way down to ground and all the way up to V_{DD}. When we are willing to tolerate an increase of a factor of two in area, we can build a much better amplifier.

ABSTRACTION

The improved wide-range circuit can be used just like our original transconductance amplifier. When we design complex analog systems, we use an amplifier as an elementary component. We will not always distinguish between the simple and wide-range transconductance amplifiers, until we work through a complete implementation. We use the symbol shown in Figure 5.3 as an *abstraction* of the detailed circuit diagram. By convention, the minus input is shown at the top and the plus input at the bottom.

If we wish to distinguish the two circuits, we use a symbol with a wider flattened end to indicate a wide-range amplifier (see Figure 5.11). When we are not sure which circuit to use, we can think about the application and work out its operating range. Most of the time, we need to deal with only the abstraction, in which the output current is a simple function of the difference between the input voltages.

The open-circuit voltage gain of either kind of transconductance amplifier is large. The voltage gain of a simple amplifier like that shown in Plate 7(a) can be 100 to 300; that of a wide-range amplifier like that shown in Plate 7(b) is 1000 to 2000. For this reason, we will often use these amplifiers as "operational amplifiers," as the term is used in classical linear-circuit design. The gain of a classical operational amplifier usually is considerably larger than that we can achieve with the designs described in this chapter. We have chosen to use a symbol for the transconductance amplifier that is similar to that commonly used for an operational amplifier, with the addition of the transconductance control input. This convention is not as confusing as it might appear to be, at first sight, to people familiar with the conventional lore. All amplifiers have a well-defined limit to the current they can supply. Hence, the conventional operational amplifier has all the limitations described for the transconductance amplifier when the former is used in the open-circuit output mode. The classical operational amplifier, however, does not allow its user to control its output-current level. This additional degree of freedom provided by the transconductance amplifier is, as we will see in the following chapters, essential to the full range of techniques necessary for large-scale analog computation. The contrast between common usage and the convention adopted in this book can thus be viewed as follows: The usual treatment of an operational amplifier emphasizes the latter's open-circuit output properties, and treats output-current limitation as a nuisance—as a deviation from ideality. Instead, we choose to view this limitation as a virtue, and to give the designer control over it. The open-circuit behavior then can be viewed as an idealization, achievable only as a limiting case. These issues will become much more clear when we consider the response of systems to time-varying inputs.

SUMMARY

We mentioned that analog circuits can do computations that are difficult or time consuming (or both) when implemented in the conventional digital paradigm. We already have seen that the transconductance amplifier computes a tanh, which is an extremely useful function. It has a very smooth transition from linear behavior for small inputs to behavior that is saturated and does not blow up if we push the input out of limits.