1.Explain in detail about the Multipole system

Let us consider a two-pole system first. For stability considerations, we plot $|\beta H|$ and $\_\ \beta H$ as a function of the frequency. Shown in Fig. 10.7, the magnitude begins to drop at 20 dB/dec at $\omega = \omega p1$ and at 40 dB/dec at $\omega = \omega p2$. Also, the phase begins to change at $\omega = 0.1\omega p1$, reaches $-45°$ at $\omega = \omega p1$ and $-90°$ at $\omega = 10\omega p1$, begins to change again at $\omega = 0.1\omega p2$ (if $0.1\omega p2 > 10\omega p1$), reaches $-135°$ at $\omega = \omega p2$, and asymptotically approaches $-180°$. The system is therefore stable because $|\beta H|$ drops to below unity at a frequency where $\_\ \beta H < -180°$.

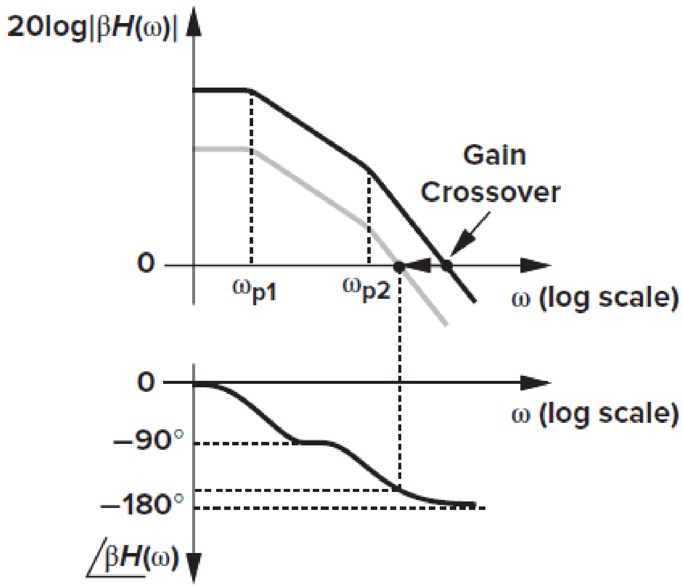


Figure 10.7 Bode plots of loop transmission for a two-pole system.

As the feedback becomes weaker, the gain crossover point moves toward the origin while the phase crossover point remains constant, resulting in a more stable system. The stability is obtained at the cost of weaker feedback.

2.Phase Margin

We have seen that to ensure stability, $|\beta H|$ must drop to unity before $\_\ \beta H$ crosses $-180°$. We may naturally ask: How far should PX be from GX? Let us first consider a "marginal" case where, as depicted in Fig. 10.10(a), GX is only slightly below PX; for example, at GX, the phase equals $-175°$. How does the closed-loop system respond in this case? Noting that at GX, $\beta H(j\omega 1) = 1\times\exp(-j175°)$, we have for the closed-loop system

$$\frac{Y}{X}(j\omega_1) = \frac{H(j\omega_1)}{1+\beta H(j\omega_1)} \tag{10.10}$$

$$= \frac{\frac{1}{\beta}\exp(-j175°)}{1+\exp(-j175°)} \tag{10.11}$$

$$= \frac{1}{\beta}\cdot\frac{-0.9962 - j0.0872}{0.0038 - j0.0872} \tag{10.12}$$

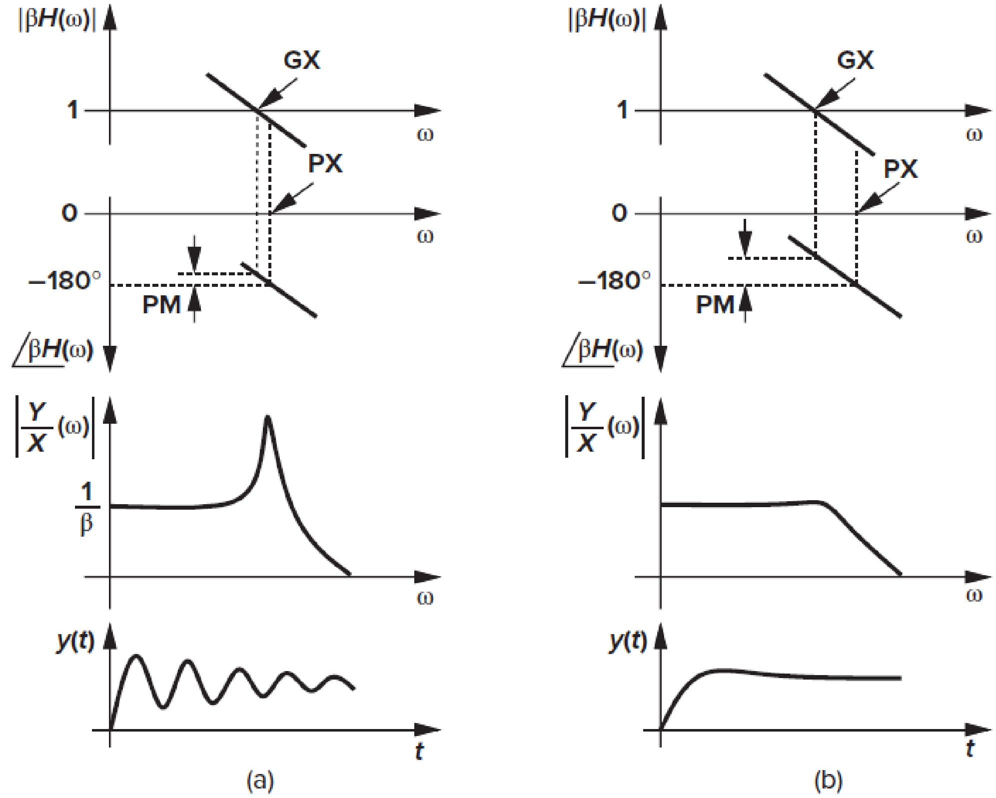


Figure 10.10 Closed-loop frequency and time response for (a) small and (b) large margin between gain and phase crossover points.

and hence

$$\left|\frac{Y}{X}(j\omega_1)\right| = \frac{1}{\beta}\cdot\frac{1}{0.0872} \quad (10.13)$$

$$\approx \frac{11.5}{\beta} \quad (10.14)$$

Since at low frequencies, $|Y/X| \approx 1/\beta$, the closed-loop frequency response exhibits a sharp peak in the vicinity of $\omega = \omega 1$. In other words, the closed-loop system is near oscillation, and its step response, *y(t)*, exhibits a very underdamped behavior. This point also reveals that a second-order system may suffer from ringing although it is stable.

Now suppose, as shown in Fig. 10.10(b), GX precedes PX by a greater margin. Then, we expect a relatively "well-behaved" closed-loop response in both the frequency domain and the time domain. It is therefore plausible to conclude that the greater the spacing between GX and PX (while GX remains below PX), the more stable the feedback system. Alternatively, the phase of βH at the gain crossover frequency can serve as a measure of stability: the smaller |_ βH| at this point, the more stable the system.

This observation leads us to the concept of "phase margin" (PM), defined as PM = 180◦ +_ $\beta H(\omega = \omega 1)$, where $\omega 1$ is the gain crossover frequency. We see that stability calls for a positive and large PM.

3.Frequency compensation Technique

Typical op amp circuits contain many poles. In a folded-cascode topology, for example, both the folding node and the output node contribute poles. For this reason, op amps must usually be "compensated," that is, their open-loop transfer function must be modified such that the closed-loop circuit is stable and the time response is well behaved.

The need for compensation arises because $|\beta H|$ does not drop to unity well before _ βH reaches

−180◦. We then postulate that stability can be achieved by (1) minimizing the overall phase shift, thus pushing the phase crossover *out* [Fig. 10.15(a)]; or (2) dropping the gain with frequency, thereby pushing the gain crossover *in* [Fig. 10.15(b)]. The first approach requires that we attempt to minimize the number of poles in the signal path by proper design. Since each additional stage contributes at least one pole, this means that the number of stages must be minimized, a remedy that yields low voltage gain and/or limited output swings (Chapter 9). The second approach, on the other hand, retains the low-frequency gain and the output swings, but it reduces the bandwidth by forcing the gain to fall at lower frequencies.

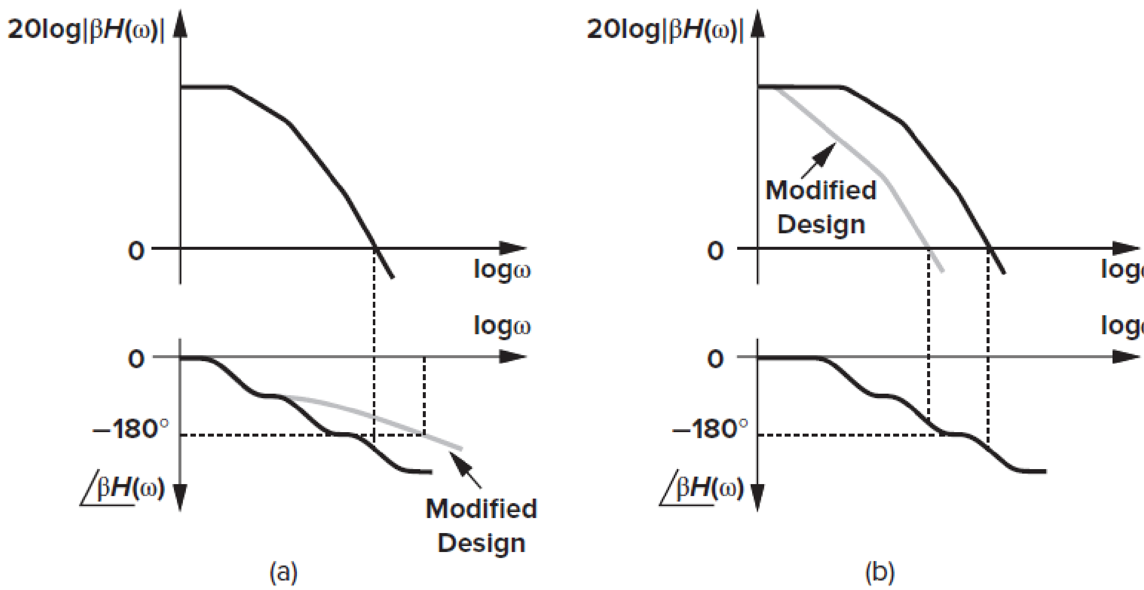


Figure 10.15 Frequency compensation by (a) moving PX out and (b) pushing GX in.

In practice, we first try to design an op amp so as to minimize the number of poles while meeting other requirements. Since the resulting circuit may still suffer from insufficient phase margin, we then compensate the op amp, i.e., modify the design so as to move the gain crossover toward the origin.

These efforts proceed with the β value chosen according to the final design requirements. For example, a closed-loop gain of 4 in some cases translates to $\beta \approx 0.25$ if the loop gain is large.3 In other words, we need not compensate the circuit for $\beta = 1$ if the closed-loop gain is always higher.

Let us apply the above concepts to the telescopic-cascode op amp shown in Fig. 10.16, where a PMOS current mirror performs differential to single-ended conversion. We identify a number of poles in the signal paths: path 1 contains a high-frequency pole at the source of $M3$, a mirror pole at node A, and another high-frequency pole at the source of $M7$, whereas path 2 contains a high-frequency pole at the source of $M4$. The two paths share a pole at the output.

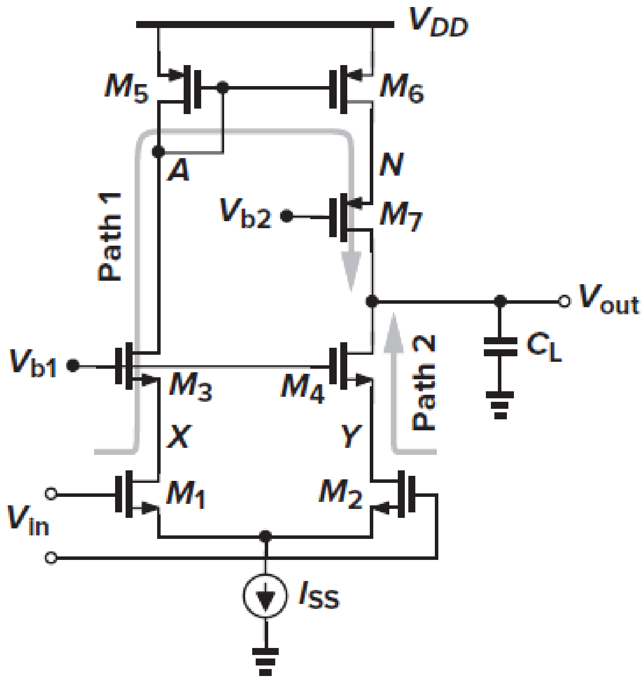


Figure 10.16 Telescopic op amp with single-ended output.

"dominant pole," $\omega_{p,out}$ usually sets the open-loop 3-dB bandwidth.

We also surmise that the first "nondominant pole," i.e., the closest pole to the origin after the dominant pole, arises at node *A*. This is because the total capacitance at this node, roughly equal to $C_{GS5} + C_{GS6} + C_{DB5} + 2C_{GD6} + C_{DB3} + C_{GD3}$, is typically quite a lot larger than that at nodes *X*, *Y*, and *N*, and the small-signal resistance of M_5, approximately $1/g_{m5}$, is also relatively large. Which node yields the next nondominant pole: *N* or *X* (and *Y*)? Recall from Chapter 9 that, to obtain a low overdrive and consume a reasonable voltage headroom, the PMOS devices in the op amp are typically wider than the NMOS transistors. Comparing M_4 and M_7 and neglecting body effect, we note that since $g_m = 2I_D/|V_{GS} - V_{TH}|$, if the two transistors are designed to have the same overdrive, they also exhibit the same transconductance. However, from square-lawcharacteristics, we have$W_4/W_7 = \mu_p/\mu_n$, which is about 1/2 to 1/3. Thus, nodes *N* and *X* (or *Y*) see roughly equal small-signal resistances toground, but node *N* suffers from much more capacitance. It is therefore plausible to assume that node *N* contributes the next nondominant pole. Figure 10.17 illustrates the results, denoting the capacitance at nodes *A*, *N*, and *X* by C_A, C_N, and C_X, respectively. The poles at nodes *X* and *Y* are nearly equal, and their corresponding terms in the transfer functions of path 1 and path 2 can be factored out. Thus, they count as one pole rather than two.

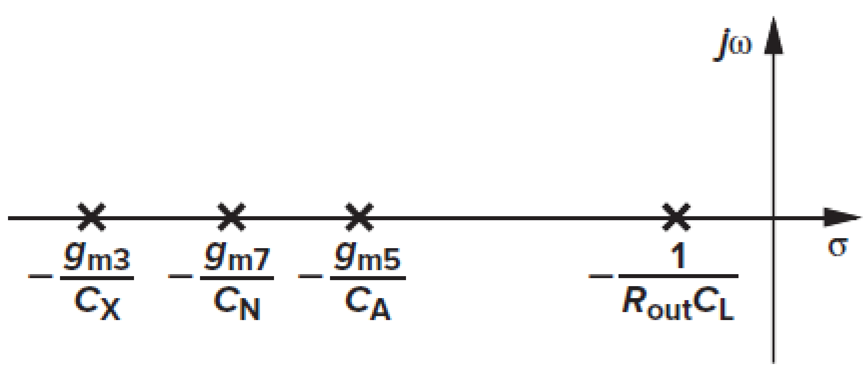


Figure 10.17 Pole locations for the op amp of Fig. 10.16.

Compensation Procedure How should we compensate the telescopic-cascode op amp? Recall that our ultimate goal is to ensure a loop gain sufficiently less than unity at the phase crossover frequency. Let us assume that the number and location of the nondominant poles and hence the phase plot at

frequencies higher than roughly $10\omega_{p,out}$, remain constant.We begin with the original response shown in Fig. 10.19, which has a negative phase margin.We must force the loop gain to drop such that the gain crossover point moves toward the origin. To accomplish this, we simply lower the frequency of the dominant pole, ω_{p1}, by increasing the load capacitance. The key point is that the phase contribution of the dominant pole in the vicinity of the gain or phase crossover point is close to 90° and relatively independent of the location of the pole. That is, as illustrated in Fig. 10.19, translating the dominant pole toward the origin affects the magnitude plot, but not the critical part of the phase plot. If ω_{p1} is lowered sufficiently, the PM reaches
an acceptable value, but at the cost of bandwidth.

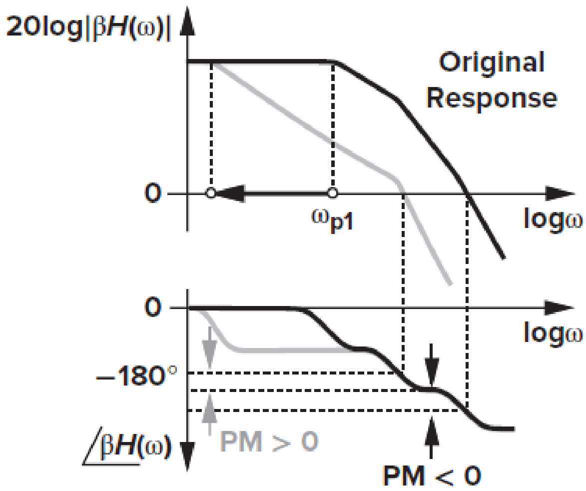
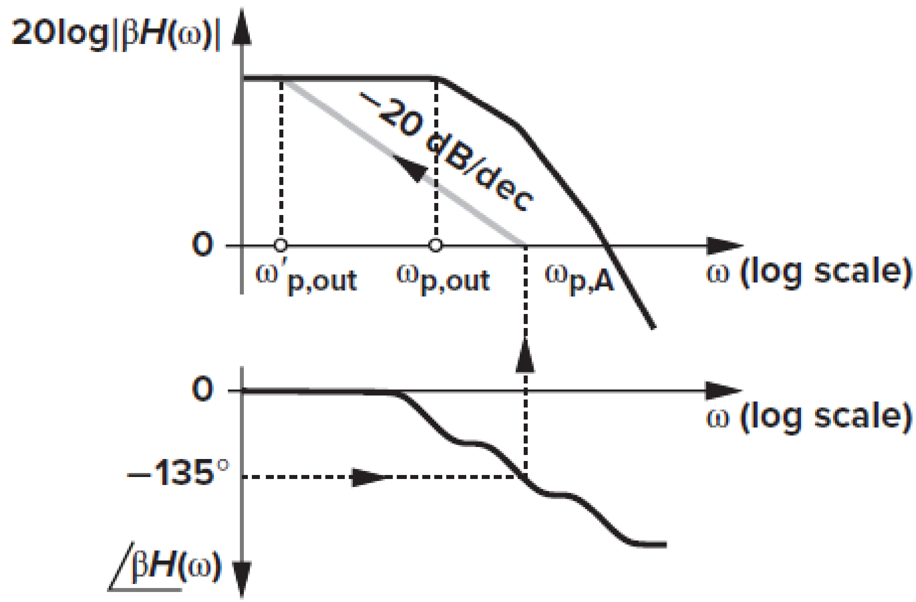


Figure 10.19 Translating the dominant pole toward the origin.

In order to determine how much the dominant pole must be shifted down as well as arrive at an important conclusion, let us assume that (1) the second nondominant pole ($\omega_{p,N}$) in Fig. 10.16 is much higher than the mirror pole so that the phase shift at $\omega = \omega_{p,A}$ is equal to -135°, and (2) a phase margin of 45° (which is usually inadequate) is necessary. To compensate the circuit, we begin from _ $\beta H(\omega) = -180^\circ + \text{PM} = -135^\circ$ and identify the corresponding gain crossover frequency, in this case, $\omega_{p,A}$ (Fig. 10.20). Since the dominant pole must drop the gain to unity at $\omega_{p,A}$ with a slope of 20 dB/dec, we draw a straight line from $\omega_{p,A}$ toward the origin with such a slope, thus obtaining the new magnitude of the dominant pole, ω_ p,out . Therefore, the load capacitance must be increased by a factor of$\omega_{p,out}/\omega$_p,out .



gure 10.20 Translating the dominant pole toward the origin for 45° phase margin.

4. **Compensation of Two-Stage Op Amps**

Consider the circuit shown in Fig. 10.25. We identify three poles: a pole at X (or Y), another at E (or F), and a third at A (or B). From our foregoing discussions, we know that the pole at X lies at relatively high frequencies. But how about the other two? Since node E exhibits a high small-signal resistance, even the capacitances of $M3$, $M5$, and $M9$ can create a pole relatively close to the origin. At node A, the small-signal resistance is lower, but CL may be large. Consequently, we say that the circuit contains *two* dominant poles.

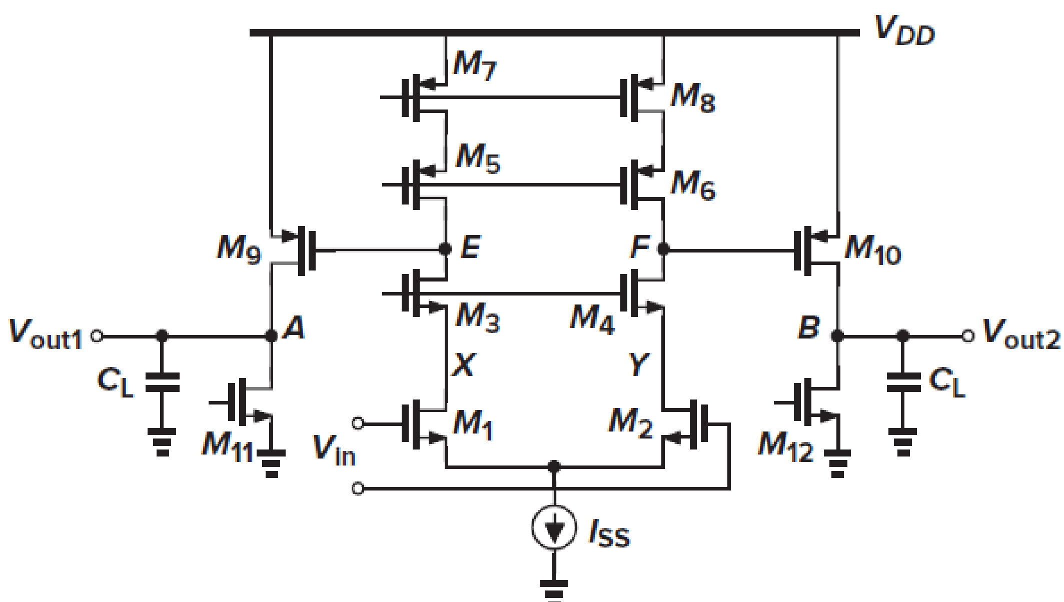


Figure 10.25 Two-stage op amp.

From these observations, we can construct the magnitude and phase plots shown in Fig. 10.26. Here, $\omega_{p,E}$ is assumed more dominant, but the relative positions of $\omega_{p,E}$ and $\omega_{p,A}$ depend on the design and the load capacitance. Note that, since the poles at E and A are relatively close to the origin, the phase

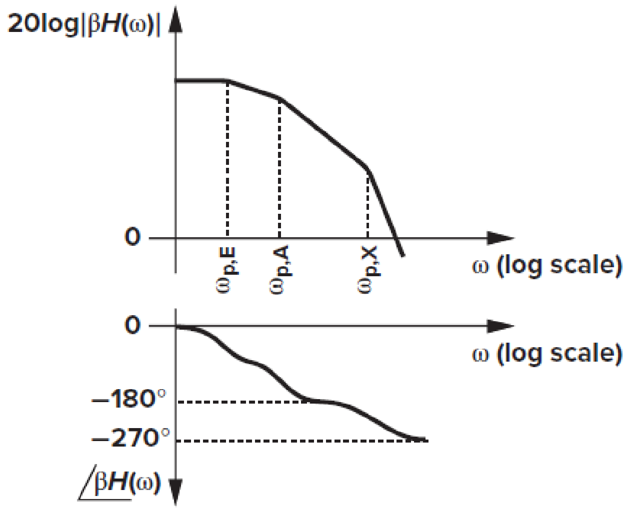


Figure 10.26 Bode plots of loop gain of two-stage op amp.

approaches −180◦ well below the third pole. In other words, the phase margin may be close to zero evenbefore the third pole contributes significant phase shift.

Let us now investigate the frequency compensation of two-stage op amps. In Fig. 10.26, one ofthe dominant poles must be moved toward the origin so as to place the gain crossover well below thephase crossover. However, recall from Sec. 10.4 that the unity-gain bandwidth after compensation cannot exceed the frequency of the second pole of the open-loop system for PM > 45◦. Thus, if in Fig. 10.26 the magnitude of *ωp,E* is to be reduced, the available bandwidth is limited to approximately *ωp,A*, a low value. Furthermore, the very small magnitude of the newdominant pole translates to a large compensation capacitor.

Fortunately, a more efficient method of compensation can be applied to the circuit of Fig. 10.25.To arrive at this method, we note that, as illustrated in Fig. 10.27(a), the first stage exhibits a highoutput impedance, *Rout*1, and the second stage provides a moderate gain, *Av*2, thereby creating a suitable environment for Miller multiplication of capacitors. Shown in Fig. 10.27(b), the idea is to create a large capacitance at node *E*, equal to *(*1+*Av*2*)CC,*moving the corresponding pole to *R*−1 *out*1[*CE*+*(*1+*Av*2*)CC*]−1, where *CE* denotes the capacitance at node *E* before *CC* is added. As a result, a low-frequency pole can be established with a moderate capacitor value, saving considerable chip area. This technique is called “Miller compensation.”

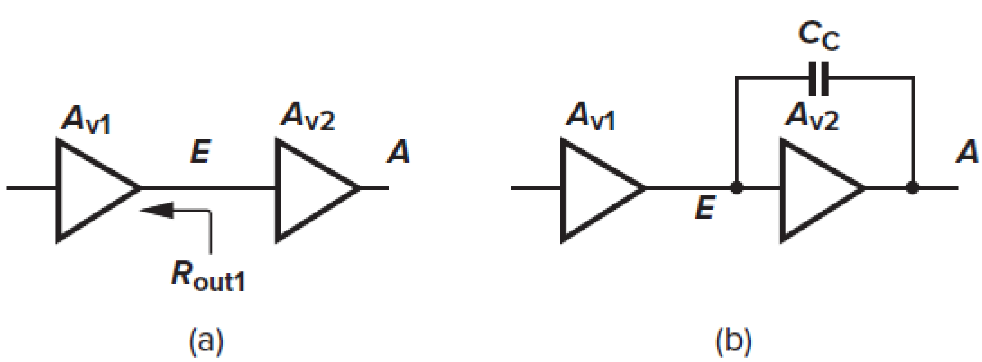


Figure 10.27 Miller compensation of a two-stage op amp.

In addition to lowering the required capacitor value, Miller compensation entails a very important property: it moves the *output* pole *away* from the origin. Illustrated in Fig. 10.28, this effect is called “pole splitting.” To understand the underlying principle, we simplify the output stage of Fig. 10.25 as in Fig. 10.29, where *RS* denotes the output resistance of the first stage and *RL* = *rO*9||*rO*11. From our