Design Linear Circuits Using OCTC Calculations

Open-circuit time constants analyze an analog design to quickly estimate its bandwidth.

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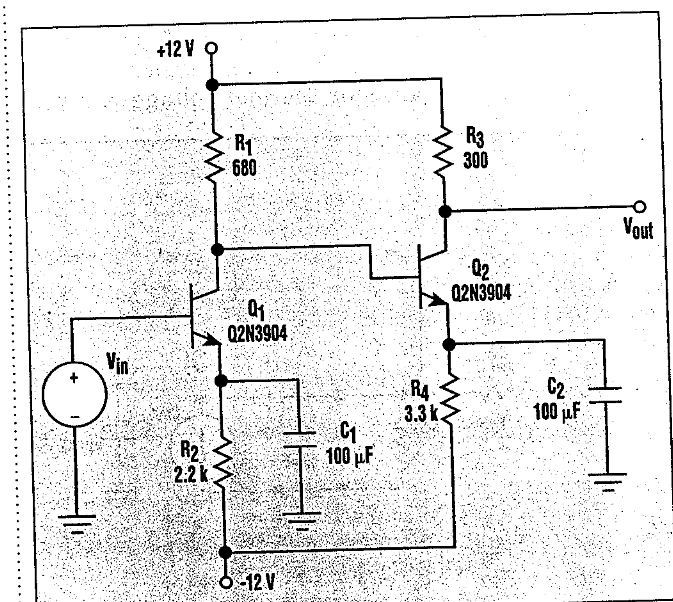
The analog-circuit analysis technique called "open-circuit time constants," developed in the 1960s by Professors R. Adler, P. E. Gray, and C. Searle at the Massachusetts Institute of Technology in Cambridge, makes two important contributions to analog circuit design. First, it allows for accurate hand calculations of a circuit's high-frequency characteristics. Second, the method provides a quick way to determine which portion of a circuit limits its bandwidth performance. The ultimate goal of open-circuit time constant analysis is to estimate the bandwidth of any circuit containing resistors, capacitors, and dependent sources.

Many times, analog-circuit designers rely exclusively on computer simulation to determine how their circuits run. Although simulation provides useful quantitative information about a circuit's operation, this approach results in little design insight. For instance, simulating a transistor circuit gives no guidance about how to design the circuit for maximum bandwidth. Because linear designers need that type of guidance, they should have approximation methods like OCTC in their engineering toolbox.

The open-circuit time constant methodology yields a systematic approach to meeting, for example, a bandwidth specification. A systematic approach is preferable to blindly tweaking resistor values during simulation to obtain exact results. That's because designers need to gain insight to assist them in future designs. Insight will teach them to understand the effect that changing an op amp or resistor will have on the circuit's bandwidth.



1 A simple amplifier built with two transistors can yield both high gain and high bandwidth.

Let's say you're designing a wideband amplifier for a communications link. The system specification tells you that the amplifier is to have a gain of at least 3000 with a high-frequency –3-dB point of 2 MHz. To meet specifications with one op amp, a gain-bandwidth product of 6 GHz would be required. However, an amplifier built with only a couple of inexpensive transistors can be designed that meets these specifications *(Fig. 1)*. But after simulation, it's found that the gain is 3750 and the –3-dB bandwidth is 1 MHz. Spice simulations show that the amplifier will not meet the bandwidth specification. What do you do?

The analysis problem rapidly becomes more difficult when the circuit has more transistors. For instance, consider the analysis of the inner workings of an operational amplifier. The LM741 op amp has 12 transistors in its signal path. Because each transistor can be modeled as having two independent capacitors, the transfer function of the amplifier has 24 poles. Finding the poles of this system involves finding the roots of a 24th-order polynomial equation, which is a difficult task. Without using computer simulation, it's impossible to predict the behavior of such a circuit without making some approximations. But even if you use computer simulation, it gives you no guidelines for design.

On the other hand, the method of open-circuit time constants applied to a linear circuit will allow you to identify the circuit elements responsible for bandwidth degradation. It's a powerful approximation method that offers significant design information which isn't available from computer simulation *(see "A more rigorous proof," p. 43)*. In addition, the method is applicable to both discrete and IC design. The procedure is as follows:

1. Generate the high-frequency incremental model for the circuit in question. In other words, replace all transistors by their small-signal models evaluated at the operating point. Replace all independent voltage sources by short circuits, and all independent current sources by open circuits. At high frequencies, all coupling and bypass capacitors behave as short circuits.

2. Find the resistance across the terminals of each capacitor that remains in the small-signal circuit, one by one, with all other capacitors acting as open circuits. These are the open-circuit resistances R_{oj}.

3. Calculate each individual time constant $\tau = R_{oj}C_j$.

SUMMARY OF RESULTS

	[illegible]	Try #2	[illegible]	Try #4
[illegible]	[illegible]	45 ns	[illegible]	4.7 ns
[illegible]	[illegible]	4.6 ns	[illegible]	0.5 ns
[illegible]	[illegible]	0.5 ns	[illegible]	0.5 ns
[illegible]	[illegible]	5 ns	[illegible]	2 ns
[illegible]	[illegible]	—	[illegible]	0.5 ns
[illegible]	[illegible]	—	[illegible]	4 ns
[illegible]	[illegible]	—	[illegible]	0.5 ns
[illegible]	[illegible]	—	[illegible]	3.2 ns
[illegible]	[illegible]	24 ns	[illegible]	0.2 ns
Sum of OCTC	[illegible]	79 ns	[illegible]	17 ns
Hand-calculated bandwidth	[illegible]	2 MHz	[illegible]	9.1 MHz
Spice-calculated bandwidth	[illegible]	2.5 MHz	[illegible]	19 MHz

4. Calculate the open-circuit-time-constant approximation to system bandwidth, which is the reciprocal of the sum of time constants, or:

$$\omega_h \approx \frac{1}{\sum_{1}^{j} \tau_{jo}} \qquad (1)$$

A design example will demonstrate the usefulness and simplicity of the procedure in transistor-circuit design. Consider the design of a transistor amplifier. The method of open-circuit time constants will guide you through the design changes necessary to meet the bandwidth specification. The minimum specifications for the transistor amplifier are:

Gain	$\lvert A_v \rvert \geq 100$
Source resistance	$R_s = 2\,k\Omega$
Load capacitance	$C_L = 10\,pF$
–3-dB bandwidth	$f_h \geq 10\,MHz$

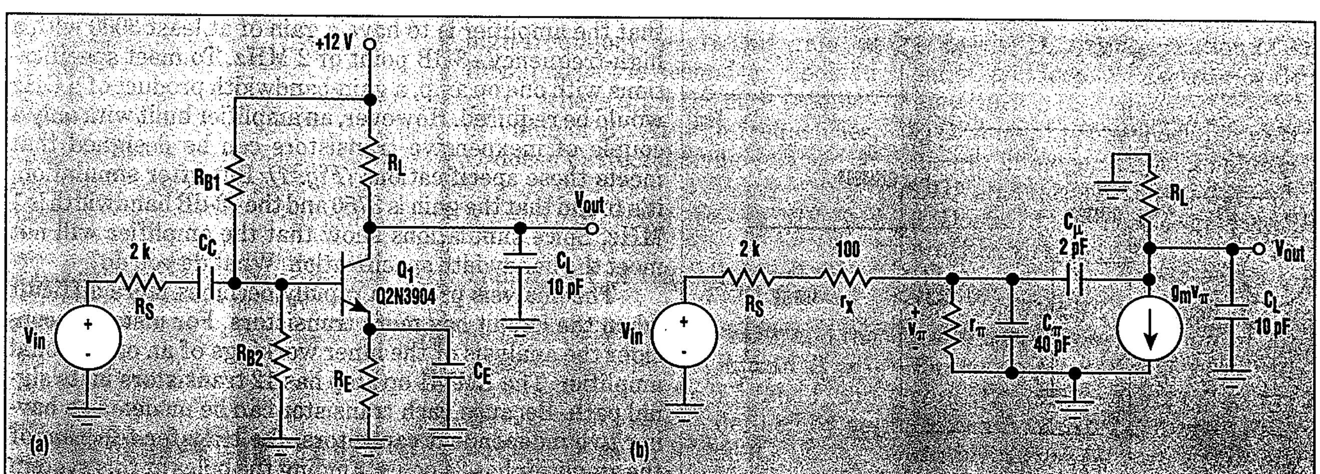
Assume that the transistor being used is the 2N3904, an inexpensive small-signal transistor with these typical specifications when it's biased at a collector current of 2 mA:

Current gain-bandwidth product	$f_T \approx 300\,MHz$
Small-signal current gain	$h_{fe} \approx 200$
Base spreading resistance	$r_x \approx 100\,\Omega$
Collector-base capacitance	$C_\mu \approx 2\,pF$ @ $V_{CB} = 10\,V$

TRY NUMBER 1

As a first attempt, try using a common-emitter amplifier in your circuit *(Fig. 2a)*. If you bias the transistor at a collector current of 2 mA, the biasing network R_{B1}, R_{B2}, and R_E set the collector current. C_E bypasses R_E at mid and high frequencies. C_C ac-couples the input signal to the base of Q_1.

To calculate gain and bandwidth, you need to know the parameters for the small-signal circuit *(Fig. 2b)*. For the bipolar transistor, the transistor thermal voltage is given



2 A common-emitter amplifier is a first attempt at meeting the bandwidth specifications (a). The parameters for the small-signal circuit help calculate the gain and bandwidth (b).

by the equation:

$$V_{TH} = \frac{kT}{q} \approx 25 \text{ mV at room temperature} \quad (2)$$

The transistor current cutoff frequency f_T is related to transistor capacitances and operating point by:

$$2\pi f_T \approx \frac{g_m}{C_\pi + C_\mu} \quad (3)$$

where g_m is the small-signal transconductance of the transistor. First, calculate the transistor small-signal parameters g_m and r_π at a quiescent collector current of 2 mA:

$$g_m = \frac{I_c}{V_{TH}} = \frac{0.002 \text{ A}}{0.025 \text{ V}} = 0.08\ \Omega^{-1} \quad (4)$$

$$r_\pi = \frac{h_{fe}}{g_m} = \frac{200}{0.08} = 2.5 \text{ k}\Omega \quad (5)$$

To calculate C_π, strictly speaking, you should first calculate the value of C_μ at the bias point V_{CB}. From the databook, assume that C_μ is approximately constant at 2 pF. This allows calculation of C_π:

$$C_\pi = \frac{g_m}{2\pi f_T} - C_\mu = \frac{0.08}{2\pi(300 \times 10^6)} - 2 \text{ pF} \approx 40 \text{ pF} \quad (6)$$

The small-signal equivalent circuit at high frequencies is given by Figure 2b. High frequency means that the coupling and bypass capacitors behave as short circuits.

To find midband gain, assume that R_{B1} and R_{B2} are much larger than R_S or r_π. You could lump the effects of R_{B1} and R_{B2} into an effective source resistance, but it's easier to ignore them altogether. This will give a small error in gain calculations, but you're interested in developing insight, not wading through lots of equations. Therefore:

$$v_o = -g_m v_\pi R_L \approx -g_m v_i \frac{r_\pi}{R_s + r_x + r_\pi} R_L \quad (7)$$

With transistor identity $g_m r_\pi = h_{fe}$:

$$A_v = \frac{v_o}{v_i} = -h_{fe} \frac{R_L}{R_s + r_x + r_\pi} \approx -0.043 R_L \quad (8)$$

A More Rigorous Proof

The OCTC method is a powerful approach to calculating the approximate bandwidth of any network of resistors, capacitors, and dependent sources. Assume a transfer function with the following constraints: The network has no zeros in the transfer function; there are only real-axis poles; and there are no poles at the origin. The transfer function is:

$$H(s) = \frac{A_o}{(\tau_1 s+1)(\tau_2 s+1)\cdots(\tau_n s+1)} \quad (B1)$$

where A_o is dc gain and each $(\tau_n s + 1)$ term defines one of the N poles in the transfer function.

You need to approximately find the "half-power frequency" ω_h (also called -3-dB or 0.707 point). As a starting point, multiply out the denominator of equation B1.

$$H(s) = \frac{A_o}{(\tau_1\tau_2\cdots\tau_n)s^n + \cdots + (\tau_1+\tau_2+\tau_n)s+1} \quad (B2)$$

Now you must make a key assumption. Assume that you're interested in frequencies "low enough" so that the s^2 and higher-order terms in equation B2 can be ignored (this is often a very good assumption, given the fact that most transistor amplifiers are operated at frequencies well below the transistor f_T). Then, you can approximate equation B2 by:

$$H(s) \approx \frac{A_o}{(\tau_1+\tau_2+\cdots+\tau_n)s+1} \quad (B3)$$

This assumption is the same as guessing that the first-order term dominates near the -3-dB frequency, resulting in a one-pole low-pass-filter approximation to the complicated transfer function. For this particular function, the half-power point is approximately given by:

$$\omega_h \approx \frac{1}{\tau_1+\tau_2+\cdots+\tau_n} \quad (B4)$$

The approximations leading to equation B4 are the basis for the OCTC method. R.B. Adler's proof shows that the sum of the time constants in the denominator of equation B4 is the sum of open-circuit time constants. With this proof, you can make quick estimates of bandwidth based on easily calculated circuit parameters.

There are several things to be aware of when using this method. First, the open-circuit-time-constants method delivers a result that's exact for a one-pole network. This makes sense because a one-pole network doesn't have s^2 and higher terms in the characteristic polynomial. It's not surprising, then, that the method is also quite accurate if there's one pole in the system much lower in frequency than all of the others. The method is least accurate if there are a large number of low-frequency dominant poles.

Second, the method's accuracy is diminished if zeroes or complex poles are in the transfer function. Third, the method always results in an estimate that's conservative. Warning to the user: There are many circuits in which the OCTC method gives a terribly pessimistic approximation. Exact answers aren't the reason to use open-circuit time constants.

Choosing $R_L = 2.4\ k\Omega$ gives a gain of approximately –103. There's some gain to spare to account for the fact that R_{B1} and R_{B2} were ignored in the gain calculation.

To find the approximate bandwidth, apply open-circuit time constants to C_π and C_μ. To find the time constant for C_π, denoted τ_{1o}, open-circuit C_μ and find the resistance R_{1o} across the C_π terminals *(Fig. 3a)*. Notice that the dependent current source $g_\mu v_\pi$ has no effect on R_{1o}, because C_μ is open-circuited. R_{1o} is given by the parallel combination of r_π and $(R_s + r_x)$:

$$R_{1o} = r_\pi \parallel (R_s + r_x) \approx 1.14\ k\Omega \qquad (9)$$

The time constant for C_π is given by:

$$\begin{aligned}\tau_{1o} &= R_{1o}C_\pi \\ &= (1.14\ k\Omega)(40\ pF) \approx 45\ ns\end{aligned} \qquad (10)$$

To find R_{2o} (the resistance facing C_μ), apply a test current source and measure the resulting voltage across the current source *(Fig. 3b)*. The resistance at the left side of C_μ is R_{1o}, which was calculated before. The open-circuit resistance R_{2o} is found by calculating the voltage generated across the current source:

$$R_{2o} = \frac{V_{test}}{I_{test}} \qquad (11)$$

By writing Kirchhoff's Voltage Law around the loop containing the test current source:

$$V_{test} = I_{test}R_{1o} - \left[-R_L(I_{test} + g_m v_a)\right] \qquad (12)$$

$$V_{test} = I_{test}R_{1o} + R_L(I_{test} + g_m I_{test} R_{1o}) \qquad (13)$$

The open-circuit resistance is:

$$\begin{aligned}R_{2o} &= R_L + (1 + g_m R_L)R_{1o} \\ &= 2.2 \times 10^5\ \Omega\end{aligned} \qquad (14)$$

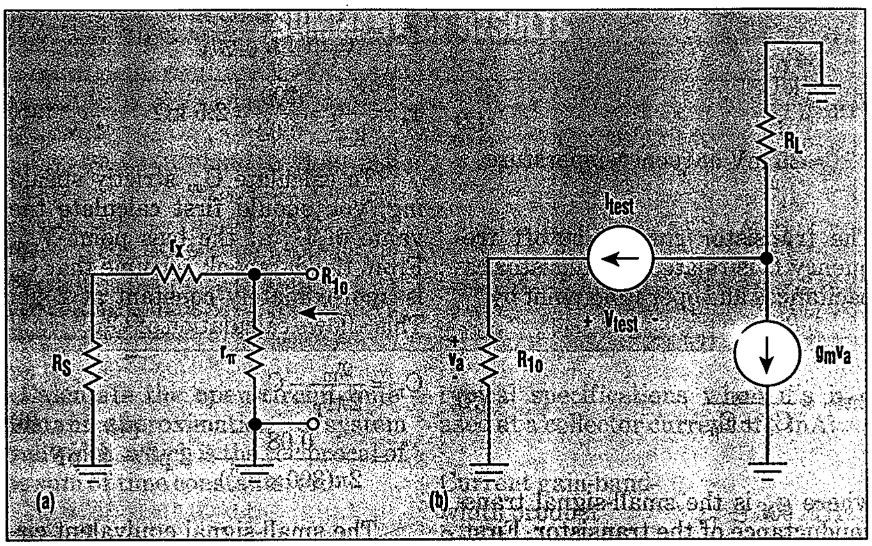
The time constant for C_μ is given by:

$$\begin{aligned}\tau_{2o} &= R_{2o}C_\mu \\ &= (2.2 \times 10^5\ \Omega)(2\ pF) = 442\ ns\end{aligned} \qquad (15)$$

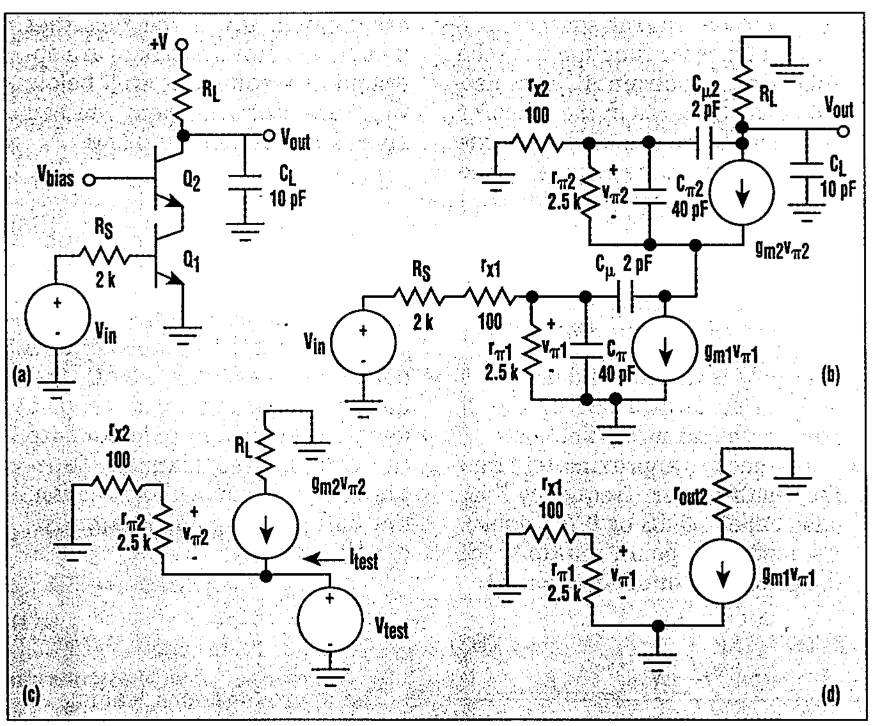
From equation 14, note that the time constant for C_μ has a $g_m R_L$ term in it. This term is the dominant factor in equation 14 and is proportional to the gain of the amplifier (Equation 7). This is the Miller effect, and is usually the dominant effect in high-gain common-emitter stages.

One other time constant is for the load capacitor C_L. The open-circuit resistance facing C_L is just the load resistor R_L, because the resistance looking into the collector of transistor Q_1 is very high:

$$\begin{aligned}\tau_{9o} &= R_L C_L \\ &= (2.4 \times 10^3)(10\ pF) = 24\ ns\end{aligned} \qquad (16)$$



3 To find the time constant for C_π, open-circuit C_μ and find the resistance across the C_π terminals (a). To find R_{2o}, which is the resistance facing C_μ, apply a test-current source and measure the resulting voltage across it (b).



4 A second attempt at meeting the specs employs a common-emitter amplifier with a cascode transistor Q_2 (a). Q_1's open-circuit time constants are found using the equivalent incremental circuit of the cascode amplifier (b). To calculate these constants, you must replace Q_2 by a resistance r_{out2}. To find r_{out2}, apply a test-voltage source to Q_2's small-signal equivalent circuit (c). Then, you can calculate τ_{2o} with the resulting circuit (d).

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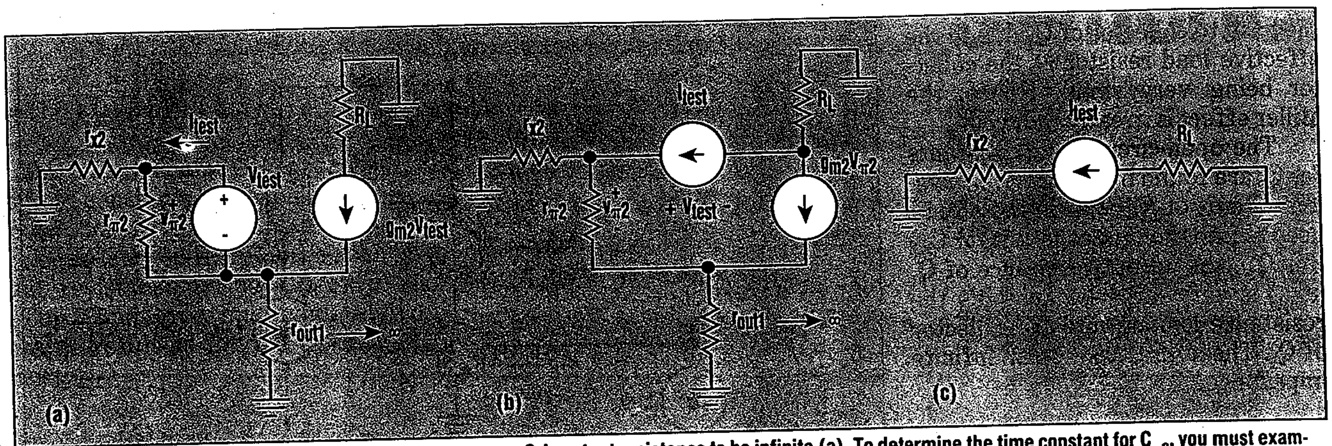


5 The circuit used to calculate Q_2's time constant assumes Q_1's output resistance to be infinite (a). To determine the time constant for $C_{\mu 2}$, you must examine the circuit and approximate that $v_{\mu 2} = 0$ (b). Consequently, the equivalent circuit makes it simple to calculate τ_{4o} (c).

The sum of the three open-circuit time constants for this amplifier is 511 ns, resulting in a bandwidth approximation of:

$$f_h \approx \frac{1}{2\pi(511\times10^{-9})} \approx 311\ \text{kHz} \qquad (17)$$

Spice simulation shows a bandwidth of approximately 315 kHz. Clearly, this amplifier topology won't meet the bandwidth specification. From the above analysis, you can see that the effects of the Miller capacitance must be reduced. The time constant associated with C_μ is much larger than those for C_π and load capacitance C_L. One way to reduce the Miller effect is by using a second transistor in the cascode configuration.

TRY NUMBER 2

In a second attempt to meet the specifications, you can add a cascode transistor Q_2 to the common-emitter amplifier *(Fig. 4a)*. Transistor Q_1 is a standard common-emitter amplifier (as in the previous try), except that its load resistor has been replaced by the cascode transistor Q_2. Q_2 is a common-base amplifier—signal current enters at the emitter and leaves at the collector.

For simplicity, the details of the biasing networks have been omitted. Because the small-signal current gain h_{fe} of the 2N3904 transistor is much greater than 1, a good approximation says that all of the small-signal collector current of Q_1 passes through Q_2 and goes through the load resistor R_L. Therefore, the gain of the cascode amplifier is approximately the same as the simple common-emitter case, and will not be calculated here.

To find the open-circuit time constants for Q_1, draw the equivalent incremental circuit of the cascode amplifier *(Fig. 4b)*. Although this circuit looks complicated at first, it's easy to break it down into smaller pieces.

To calculate the open-circuit time constants for Q_1, you can replace Q_2 by a resistor (r_{out2}) whose value is equal to the resistance looking into the emitter of Q_2. Once the value of this resistor is found, you can use the same equations as before to determine the C_μ time constant.

To find r_{out2}, apply a test voltage source to the small-signal equivalent circuit of Q_2 *(Fig. 4c)*. Q_2 is biased at approximately the same collector current as Q_1. Therefore, g_m, r_π, and C_π are the same as for Q_1. This results in:

$$r_{out2} = \frac{V_{test}}{I_{test}} = \frac{r_{x2} + r_{\pi 2}}{1 + h_{fe2}} \approx 13\ \Omega \qquad (18)$$

The time constants for Q_1 are the same as calculated before, with R_L replaced by the equivalent output resistance of Q_2 *(Fig. 4d)*. Note that R_{1o}, the resistance facing $C_{\pi 1}$, is the same as before:

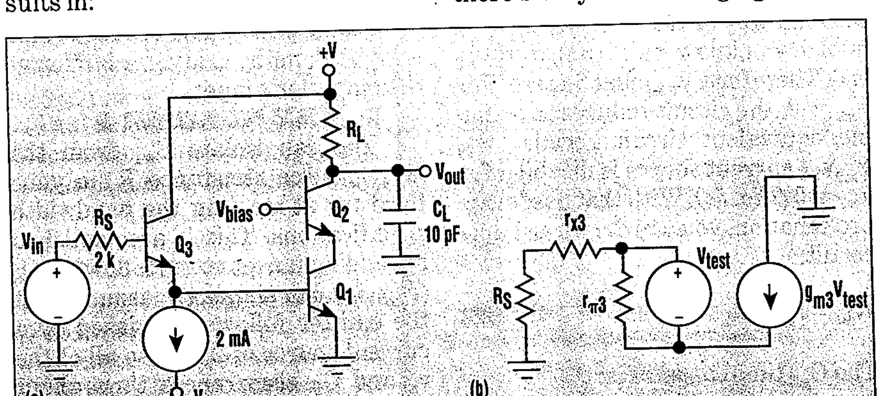
$$R_{1o} = r_\pi \parallel (R_s + r_x) \approx 1.14\ \text{k}\Omega \qquad (19)$$

As a result, τ_{1o} hasn't changed. You can find τ_{2o} by using equation 14 with R_L replaced by r_{out2}:

$$R_{2o} = R_{out2} + (1 + g_m r_{out2})R_{1o} \approx 2.3\ \text{k}\Omega \qquad (20)$$

$$\tau_{2o} = R_{2o}C_{\mu 2} = (2.3\ \text{k}\Omega)(2\ \text{pF}) \approx 4.6\ \text{ns} \qquad (21)$$

Cascoding reduces the time constant for $C_{\mu 1}$ from 442 to 4.6 ns. Intuitively, this makes sense because there's very little voltage gain from



6 Adding the Q_3 emitter follower isolates $C_{\pi 1}$ from the large source resistance in a third attempt at meeting the specifications (a). An equivalent circuit is used to find the open-circuit time constants for that emitter-follower transistor (b).

the base to collector of Q_1 due to the effective load resistor at the collector being very small. Hence, the Miller effect is greatly reduced.

The open-circuit time constants for Q_2 are found by using the equivalent circuit of Fig. 5a. In the circuit used to find the time constant for $C_{\pi 2}$ (for Q_2), the output resistance of Q_1 (r_{out1}) is very large because it's the resistance looking into the collector of Q_1. Therefore, assume an infinite impedance:

$$I_{test} = \frac{V_{test}}{r_{\pi 2}} + g_{m2}V_{test} \approx g_{m2}V_{test} \quad (22)$$

$$R_{3o} = \frac{V_{test}}{I_{test}} \approx \frac{1}{g_{m2}} \approx 12\ \Omega \quad (23)$$

The time constant for $C_{\pi 2}$ is:

$$\begin{aligned}\tau_{3o} &= R_{3o}C_{\pi 2} \\ &= (12\ \Omega)(40\ pF) \approx 0.5\ ns\end{aligned} \quad (24)$$

If you blindly start writing equations to find the time constant for $C_{\mu 2}$, soon you'll have a big mess on your hands. However, if you look at the circuit closely, there's a very useful approximation that can be made *(Fig. 5b)*. First, because r_{out1} is very large compared with the other resistances in the circuit, negligible current will be flowing through it. Therefore, all of the current from the $g_{m2}v_{\pi 2}$ generator flows through $r_{\pi 2}$. This constrains:

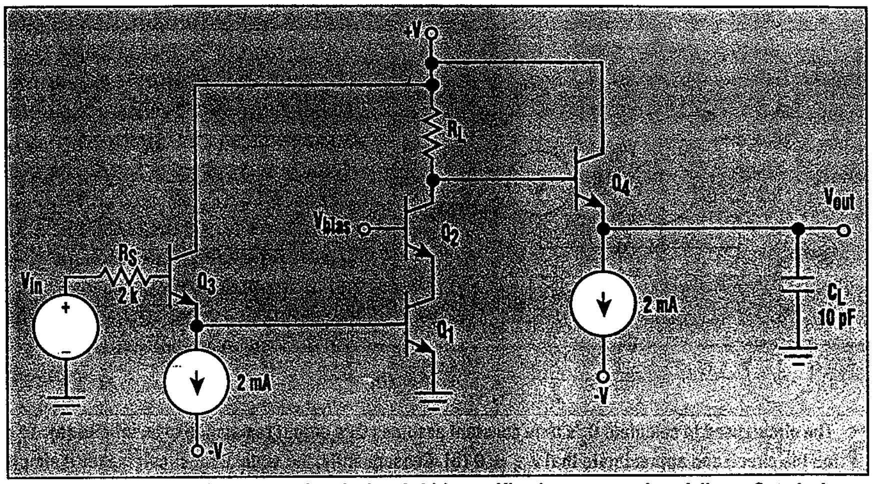
$$v_{\pi 2} = -g_{m2}r_{\pi 2}v_{\pi 2} \quad (25)$$

By recognizing the transistor relation $g_m r_\pi = h_{fe}$, the above relationship can only be true if either $h_{fe} = -1$ (which you know isn't true), or if $v_{\pi 2} = 0$. Therefore, $v_{\pi 2}$ must be zero. For $v_{\pi 2} = 0$, the circuit reduces to a simple equivalent circuit because the $g_{m2}v_{\pi 2}$ current source is disabled for $v_{\pi 2} = 0$ *(Fig. 5c)*. With that useful approximation, you can derive a simple result:

$$R_{4o} = r_{x2} + R_L = 2.5\ k\Omega \quad (26)$$

The $C_{\mu 2}$ time constant is:

$$\begin{aligned}\tau_{4o} &= R_{4o}C_{\mu 2} \\ &= (2.5\ k\Omega)(2\ pF) = 5\ ns\end{aligned} \quad (27)$$



7 The fourth and final try at meeting the bandwidth specifications uses emitter follower Q_4 to isolate the load capacitor from the large load resistor.

The sum of all five open-circuit time constants is 79 ns, corresponding to an approximate bandwidth of 2 MHz *(see the table)*. For non-believers, Spice results have also been included for comparison.

By adding the cascode transistor, your bandwidth is increased by a factor of six. Although you still haven't met the specification, other things can be done. The important result from the above calculation is that the dominant factor limiting the bandwidth of the amplifier is the time constant τ_{1o}, which is the C_π time constant for transistor Q_1. Reducing $C_{\pi 1}$, $r_{\pi 1}$, r_{x1}, or R_s will increase the bandwidth of the amplifier; reducing any of the other resistances in the circuit will only give small increases in bandwidth. Because $C_{\pi 1}$ and $r_{\pi 1}$ are fixed by the bias point and r_{x1} is process-dependent, the effective source impedance seen by $C_{\pi 1}$ should be reduced. This may be done by adding a buffer at the amplifier's input.

TRY NUMBER 3

You can isolate $C_{\pi 1}$ from the large source resistance R_s in your third try at meeting the bandwidth specifications. This is accomplished by adding an emitter follower, Q_3 *(Fig. 6a)*. For simplicity, assume that the emitter follower is biased by a current source of 2 mA, so all of the r_π and C_π occurrences throughout the circuit are equal.

The gain of the emitter follower is very close to 1. Its output resistance is given by:

$$r_{out3} = \frac{R_s + r_{x3} + r_{\pi 3}}{1 + h_{fe3}} \approx 23\ \Omega \quad (28)$$

The equivalent source resistance seen by $C_{\pi 1}$ is now much smaller. Adding the emitter follower also affects the amplifier's gain. Assuming that the emitter follower has unity gain, equation 8 yields:

$$\begin{aligned}A_v &\approx -h_{fe1}\frac{R_L}{r_{out3} + r_{x1} + r_{\pi 1}} \\ &\approx -0.076R_L\end{aligned} \quad (29)$$

This tells you that the same gain can be attained by decreasing R_L. Decreasing R_L will reduce the open-circuit time constants for both Q_2 and the load capacitor. Choosing a smaller load resistor, $R_L = 1500\ \Omega$, meets the gain specification (gain ≈ -114). The OCTC for $C_{\pi 1}$:

$$\begin{aligned}\tau_{1o} &= R_{1o}C_{\pi 1} \\ &= \{r_{\pi 1} \parallel (r_{out3} + r_{x1})\}C_{\pi 1} \\ &\approx (117\ \Omega)(40\ pF) \approx 4.7\ ns\end{aligned} \quad (30)$$

As expected, adding the buffer greatly reduces this time constant.

For $C_{\mu 1}$:

$$\begin{aligned}R_{2o} &= r_{out2} + (1 + g_{m1}r_{out2})R_{1o} \\ &\approx 250\ \Omega\end{aligned} \quad (31)$$

$$\tau_{2o} = (250\ \Omega)(2\ pF) = 0.5\ ns \quad (32)$$

For Q_2, τ_{3o} is unchanged. R_{4o} is decreased, because the load resistor has decreased:

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$$\tau_{4o} = (r_{x3} + R_L)C_{\mu 2} = (1.6\ k\Omega)(2\ pF) = 3.2\ ns \quad (33)$$

You must calculate the open-circuit time constants for the emitter-follower transistor Q_3. The equivalent circuit lets you find R_{5o}, the open circuit resistance facing $C_{\pi 3}$ *(Fig. 6b)*. This circuit gives approximately the same result as for the cascode transistor:

$$R_{5o} = \frac{1}{g_{m3}} \approx 12\ \Omega \quad (34)$$

$$\tau_{5o} = (12\ \Omega)(40\ pF) = 0.5\ ns \quad (35)$$

For $C_{\mu 3}$, you can again use the equivalent circuit of Figure 4b because there's a current source (which equals high impedance) in the emitter of Q_3:

$$R_{6o} = R_s + r_{x3} = 2.1\ k\Omega \quad (36)$$

$$\tau_{6o} = (2.6\ k\Omega)(2\ pF) = 4\ ns \quad (37)$$

The time constant associated with the load capacitor at the output has also decreased because of the smaller value of load resistance:

$$\tau_{9o} = R_L C_L = (1.5\ k\Omega)(10\ pF) = 15\ ns \quad (38)$$

The sum of the open-circuit time constants is now 28 ns. The approximate bandwidth, calculated from the sum of open-circuit time constants, is now 5.6 MHz.

The specification still isn't met, but there's hope. From the table, you can see the dominant time constant is τ_{9o}, due to the 10-pF load capacitor interacting with the 1.5-kΩ load resistor. Therefore, to increase the bandwidth, you must reduce the effects of the load capacitor coupling with the load resistor. This can be done by adding an output emitter-follower buffer.

TRY NUMBER 4

In your fourth try, add an emitter follower (Q_4) to isolate the load capacitor C_L from the large load resistor R_L *(Fig. 7)*. Adding the emitter follower won't change time constants τ_{1o} through τ_{6o}. Proving this is left as an exercise for the reader.

However, the other three time constants do change. The three time constants for the new transistor Q_4 are:

$$\tau_{7o} \approx \frac{C_{\pi 4}}{g_{m4}} \approx (12.5\ \Omega)(40\ pF) \approx 0.5\ ns \quad (39)$$

$$\tau_{8o} \approx (R_L + r_{x4})C_{\mu 4} = (1.6\ k\Omega)(2\ pF) \approx 3.2\ ns \quad (40)$$

$$\tau_{9o} \approx r_{out4}C_L = \frac{R_L + r_{x4} + r_{\pi 4}}{1 + h_{fe4}} \approx (20\ \Omega)(2\ pF) = 0.2\ ns \quad (41)$$

The approximate bandwidth, calculated from the sum of open-circuit time constants, is now 9.1 MHz. Because the method of open-circuit time constants gives an approximation that's always pessimistic, this fourth try is a good candidate for full computer simulation. Spice simulation shows that the desired circuit specifications are finally met with your fourth attempt.

Computer simulation provides a useful method for analyzing multistage transistor circuits. Unfortunately, simulation offers virtually no design insight to tell you what elements in your circuit are responsible for degrading bandwidth. On the other hand, open-circuit time constant analysis helps you determine the necessary tradeoffs for implementing a good design. ED

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