Chapter 2
Continuous-Time Analog Circuits

Analog performance has been improved by careful design, new process, or feedback. Feedback is the only systematic way to enhance analog performance such as linearity, signal range, bandwidth, and impedance. Series (voltage) feedback increases linear voltage range while shunt (current) feedback increases linear current range. Also series feedback raises impedance while shunt feedback lowers impedance. Low impedance is for broad-banding while high impedance is for buffering. Feedback can be applied at any local or global levels, in continuous-time or discrete-time modes, and in broadband or DC servo applications.

2.1 Negative and Positive Feedbacks

There are two feedbacks. Negative feedback is to make stable systems such as power supplies, amplifiers, filters, ΔΣ modulators, PLL, and adaptive equalizers while positive feedback is to make unstable systems such as latches and oscillators. Only stability tells the difference between the negative and positive feedbacks. That is, negative-feedback amplifiers should be stable while positive-feedback oscillators should be unstable.

The feedback system is sketched conceptually in Fig. 2.1. The ports marked as i and o are the points the input is injected into and the output is taken from, and the path gains a and f represent the forward and feedback gains, respectively. Stability and dynamic performance are not affected by the locations of the input and output ports, but entirely by the loop gain of af. Depending on the polarity of the feedback loop gain, it makes either the negative or positive feedback.

Most closed-loop analog circuits except for latches and oscillators operate in stable negative feedback modes under the standard assumption of the small-signal, linear, time-invariant (SLT) operating condition. They benefit greatly from high linearity given by feedback. On the other hand, narrowband RF circuits such as low-noise amplifier (LNA) and mixer operate in open loop, and nonlinearity

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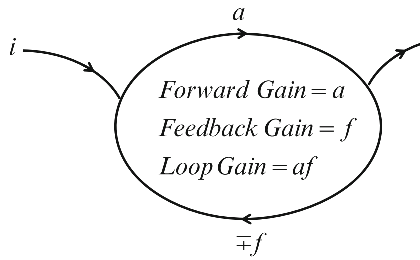


$$\frac{o}{i} = \frac{a}{1 \pm af}$$

Fig. 2.1 Feedback system model

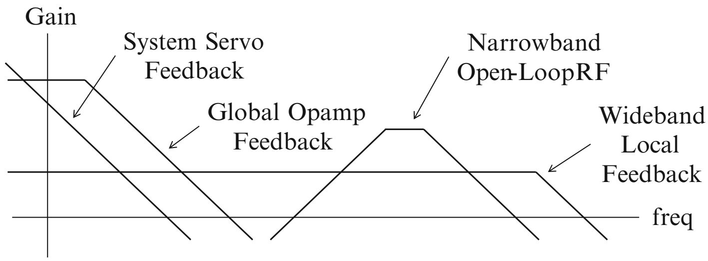


Fig. 2.2 Three kinds of analog circuits and feedbacks

becomes the most critical design constraint. RF design is basically how to achieve linearity while keeping noise low. However, in wideband baseband systems covering DC to GHz, local feedback can be applied to broadband amplifiers.

Figure 2.2 illustrates three operating frequency ranges of analog circuits. Feedback plays a critical role in obtaining the desired linearity except for narrowband RF circuits, which operate with inductor loads without feedback mainly to meet the low-power and low-noise requirements. Even at system levels, high-gain DC servo feedbacks can be applied so that performance can be enhanced by adapting to various circuit parameters.

Feedforward and feedback have quite different implications in circuits as shown in Fig. 2.3. The former is to feed the input forward and sum it at the output. The amplifier gain drops at high frequencies, but its output is held up by the signal fed forward instead. As a result, a zero is formed at the break point ω_z. However, since it is an open-loop implementation, there is no stability issue at all. On the other hand, the latter is to feed some of the amplifier output back and subtract it from the input. Since the subtracted difference is fed back into the amplifier, the input difference error is reduced by loop gain. That is, it makes a pole at the unity-loop gain frequency ω_k, where the loop gain becomes unity. If the extra phase delay of this amplified error approaches 180° at ω_k, the negative feedback becomes unstable positive feedback.

Feedforward is equivalent to the digital OR function as shown in Fig. 2.4. The analog OR function is to sum the responses of two parallel paths. In the wideband LNA, splitting noise into two parallel inverting and non-inverting paths and

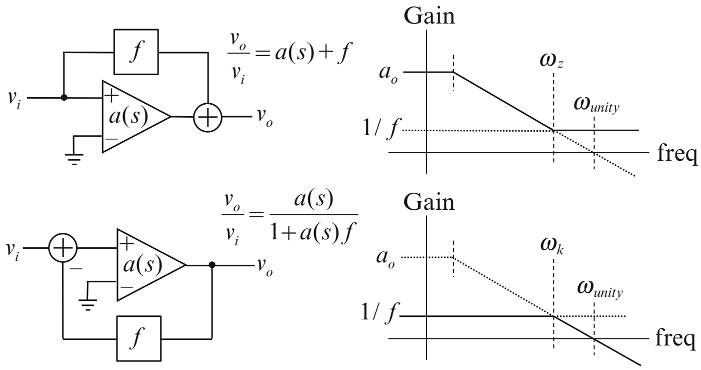


Fig. 2.3 Feedforward vs. feedback systems

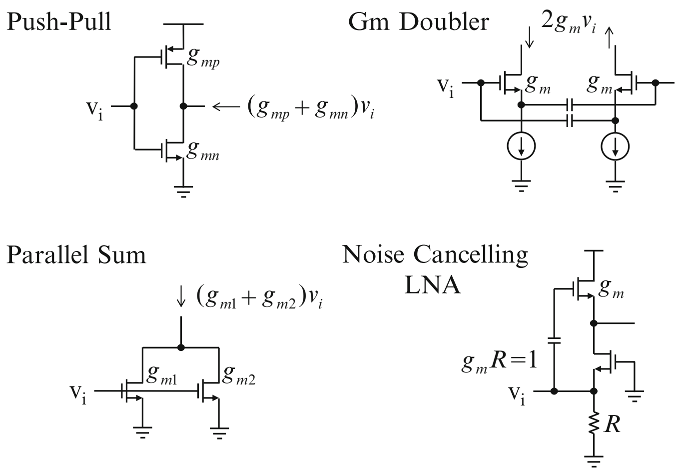
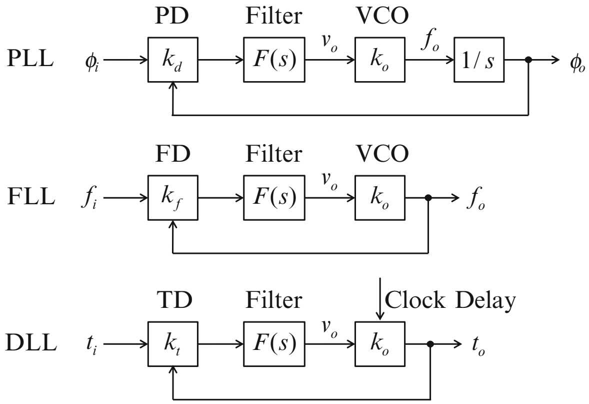


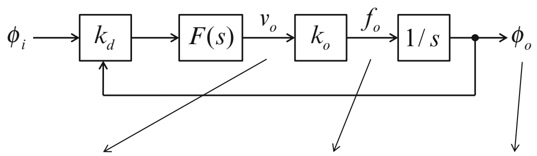
Fig. 2.4 Feedforward two-path examples

combining them also cancels the in-band noise—called noise cancellation. In addition, the sum of two parallel transistor currents effectively doubles the transistor g_{m} in the push–pull and parallel configurations. Although feedforward is useful to improve analog performance in some limited ways, it is only the negative feedback that can directly trade gain to enhance analog performance.

In the examples shown in Fig. 2.5, feedback loops encircles circuit elements not only in the voltage domain but also in the phase, frequency, and time domains. Depending on the error detector, the loop makes 3 unity-gain followers for phase, frequency, and time—called phase-locked loop (PLL), frequency-locked loop (FLL), and delay-locked loop (DLL), respectively. Among these, PLL is most widely used since the phase detector is the easiest among them to implement.

Fig. 2.5 Feedback examples of phase, frequency, and time followers





$$\frac{v_o}{\phi_i}=\frac{k_dF(s)}{1+k_dF(s)k_o/s},\quad \frac{f_o}{\phi_i}=\frac{k_dF(s)k_o}{1+k_dF(s)k_o/s},\quad \frac{\phi_o}{\phi_i}=\frac{k_dF(s)k_o/s}{1+k_dF(s)k_o/s}$$

$$\frac{f_o}{f_i}=\frac{\phi_o}{\phi_i}=\frac{k_dF(s)k_o}{s+k_dF(s)k_o}\;\Rightarrow\;\frac{f_o/k_o}{f_i}=\frac{v_o}{f_i}=\frac{k_dF(s)}{s+k_dF(s)k_o}$$

Fig. 2.6 Three forward gains with the same feedback loop gain

In the PLL, three outputs of voltage, frequency, and phase can be taken from the loop as shown in Fig. 2.6. The illustrated three transfer functions differ only in the forward gain, but they share the same loop gain. It is also shown that the voltage output is the demodulated FM output since the frequency is obtained by differentiating the phase, and the VCO converts voltage into frequency.

Most electronic systems use various local or global feedbacks. There are negative nonlinear feedback systems similar to PLL. An oversampling ΔΣ modulator suppresses quantization noise by loop gain. Manual trimming procedures carried out by engineers are also kinds of negative feedback schemes including human intelligence in the loop. However, the stability of all these loops is determined only by the loop gain and phase.

2.1.1 Phase Margin

In negative feedback, stability is determined by the phase margin at the frequency ω_k where the loop gain is unity or by the gain margin at the frequency ω_{180} where the excess loop phase is 180°. Both frequencies can be obtained from the following relations if the loop gain is set by $T(j\omega) = a(j\omega)f$.

$$\left|T(j\omega_k)\right| = 1 \quad \text{and} \quad \angle T(j\omega_{180}) = -180°. \tag{2.1}$$

These two frequencies have special meanings: $\omega_k < \omega_{180}$ for stability, and $\omega_k > \omega_{180}$ for instability as graphically explained in the Bode plots shown in Fig. 2.7.

Both gain and phase margins (GM and PM) are defined as extra rooms for more loop gain and extra loop phase until the oscillation condition is reached. Feedback systems become unstable if the loop gain is greater than unity at the frequency where the total loop phase delay becomes the multiples of 2π such as 0°, 360°, and 720°. So for negative feedback systems to be stable, the PM should be positive, and the GM should be greater than 1. Similarly, for positive feedback systems, the PM should be negative, and the GM is smaller than 1. That is, referring only PM is sufficient to assure the stability of the feedback system. GM and PM are defined as follows.

$$\text{GM} = \frac{1}{|T(j\omega_{180})|} \quad \text{and} \quad \text{PM} = 180° - \left|\angle T(j\omega_k)\right|. \tag{2.2}$$

Unless both GM and PM conditions are warranted, amplifiers get unstable and oscillate while oscillators get stable and amplify. That is, the boundary between the

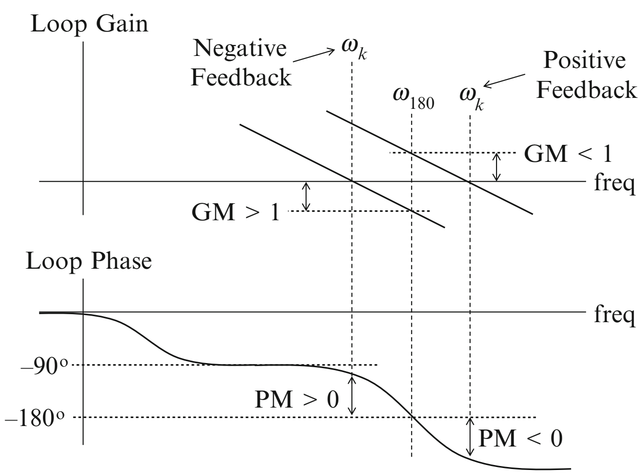


Fig. 2.7 Gain and phase margins for negative feedback

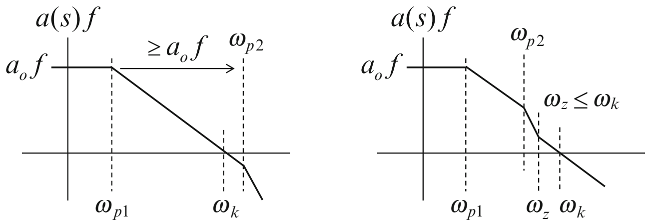


Fig. 2.8 Bode gain plots of two poles with and without a zero

negative and positive feedbacks is set by the loop gain at ω_{180}. If the loop gain is lower than unity, it is stable negative feedback. Otherwise, it becomes unstable positive feedback.

The stability condition can be stated in two ways as shown in the Bode gain plot of Fig. 2.8. The stability with PM greater than 45° is warranted if two poles are separated by more than the DC loop gain of $a_o f$. If two poles are not separated enough, a zero ω_z should be placed at lower frequencies than the unity loop-gain frequency ω_k. Then the PM becomes

$$\text{PM} = 90^\circ + \tan^{-1}\frac{\omega_k}{\omega_z} - \tan^{-1}\frac{\omega_k}{\omega_{p2}}. \tag{2.3}$$

2.1.2 Stability of Negative Feedback

As noted, stability has been analyzed in many different ways. A few of them are: (1) Poles should be on the open left half plane. (2) The complex plot of the loop gain shouldn't encircle the (−1,0) point in negative feedback, and encircle the (1,0) point in positive feedback. (3) The zero-input response should die out as time goes by. (4) There should be GM and PM in the loop gain. Among them, the PM is most handy, and widely referred for stability. To relate the pole location and the PM, Root Locus which projects the trajectory of poles in the complex plane as a function of the feedback loop gain can be considered. The relation between the PM and the pole location for Chebyshev and Butterworth poles is illustrated for two-pole cases in Fig. 2.9.

If the unity loop-gain frequency ω_k is about the same as the second pole ω_{p2}, the PM is about 45°, and the complex conjugate poles are at 60° from the real axis in the Chebyshev case. On the other hand, if $\omega_k = \omega_{p2}/1.4$, then it makes the maximally flat Butterworth poles at 45° from the real axis. The closer to the imaginary axis the pole is, the higher the Q goes. High-Q complex conjugate pole pair causes peaking in the frequency response, and also makes overshoot and ringing in the transient response. The Root Locus shows the movement of poles as the loop gain is increased. The common two-pole feedback examples are shown in Fig. 2.10.

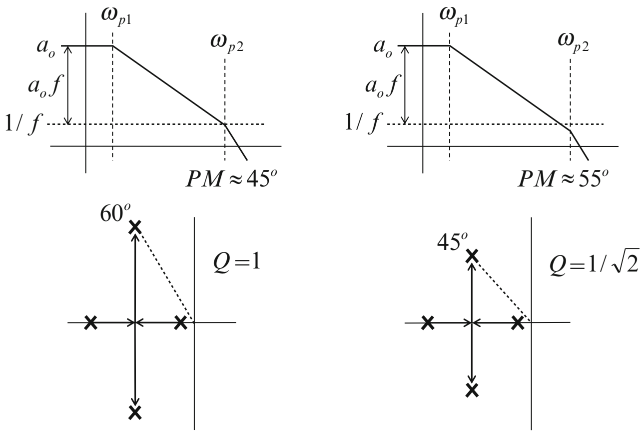


Fig. 2.9 Bode plots vs. poles in the complex plane

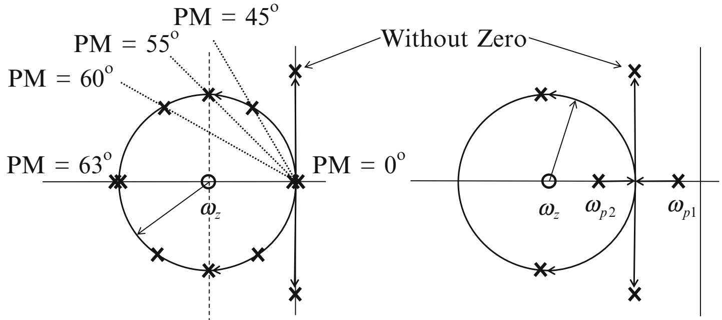


Fig. 2.10 Root Locus of two-pole feedback examples

Two open-loop poles enclosed in the feedback loop are two DC poles of integrators and two negative real poles such as in opamps, respectively. In both cases, two poles split into the vertical directions as the loop gain increases. A zero is needed to pull the poles into the open left half plane. Otherwise, as the loop gain further increases, the third pole easily pushes them into the right half plane and causes instability. The two DC poles split immediately while two negative real open-loop poles of ω_{p1} and ω_{p2} come closer before they split to be complex conjugate poles.

Let's consider two DC poles with a unity-gain frequency of ω_k. Due to the high DC gain, integrators are used as error amplifiers in most feedback systems such as PLL and $\Delta\Sigma$ modulator. If the unity-gain frequency is ω_k, the two-pole loop gain is $(\omega_k/s)^2$, and the closed loop transfer function has two imaginary axis poles at $+j\omega_k$ and $-j\omega_\kappa$. To move these poles into the open left half plane, a zero is placed at ω_z, where the loop gain is ω_k/ω_z. Then the open-loop gain becomes

$$\text{Loop Gain} = \frac{\omega_k}{s}\left(\frac{\omega_k}{s} + \frac{\omega_k}{\omega_z}\right). \tag{2.4}$$

Therefore, the closed-loop gain is given by

$$H(s) = \frac{\frac{\omega_k}{s}\left(\frac{\omega_k}{s} + \frac{\omega_k}{\omega_z}\right)}{1 + \frac{\omega_k}{s}\left(\frac{\omega_k}{s} + \frac{\omega_k}{\omega_z}\right).} \tag{2.5}$$

Here the gain factor is the ratio of ω_k/ω_z, which implies that the zero is placed at lower frequency than the unity-gain frequency by this ratio. By solving for the roots of the denominator polynomial, two poles can be found to be at

$$\left\{-\frac{\omega_k}{2\omega_z} \pm j\sqrt{1 - \left(\frac{\omega_k}{2\omega_z}\right)^2}\right\} \times \omega_k. \tag{2.6}$$

Two poles given by (2.6) are plotted in Fig. 2.10 as a function of ω_k/ω_z, and marked when its values are 0, 1, 1.414, 1.732, and 2. The PM can be also defined as follows.

$$\text{PM} = \tan^{-1}\left(\frac{\omega_k}{\omega_z}\right). \tag{2.7}$$

From (2.6), it can be shown that the Root Locus is also a circle centered at $-\omega_z$ with a radius of ω_z.

$$\sqrt{\left(-\frac{\omega_k^2}{2\omega_z} + \omega_z\right)^2 + \left\{1 - \left(\frac{\omega_k}{2\omega_z}\right)^2\right\} \times \omega_k^2} = \omega_z. \tag{2.8}$$

The Root Locus of two negative real poles with one zero as found in opamps is similar. The closed-loop gain with a DC gain of a_o and a loop gain of T_o is

$$H(s) = \frac{\frac{a_o\omega_{p1}\omega_{p2}}{\omega_z} \times \frac{(s+\omega_z)}{(s+\omega_{p1})(s+\omega_{p2})}}{1 + \frac{T_o\omega_{p1}\omega_{p2}}{\omega_z} \times \frac{(s+\omega_z)}{(s+\omega_{p1})(s+\omega_{p2})}}. \tag{2.9}$$

By solving for the roots of the denominator polynomial, two real poles are moved to

$$-\left(\frac{\omega_{p1} + \omega_{p1}}{2} + \frac{T_o\omega_{p1}\omega_{p2}}{2\omega_z}\right) \pm j\sqrt{(1+T_o)\omega_{p1}\omega_{p2} - \left(\frac{\omega_{p1} + \omega_{p1}}{2} + \frac{T_o\omega_{p1}\omega_{p2}}{2\omega_z}\right)^2}. \tag{2.10}$$

Poles	X-Axis Angle	Phase Margin	Q	Band-Width
2 Real Poles	0°	63°	0.5	0.64
Bessel (Linear Phase)	30°	60°	0.58	0.8
Butterworth (Maximally Flat)	45°	55°	0.71	1
Chebyshev (1dB Ripple)	60°	45°	1	1.3

Fig. 2.11 Locations of two complex-conjugate poles

The Root Locus makes a circle around ω_z with a radius of the geometric mean of the distances to ω_{p1} and ω_{p2} from ω_z as follows.

$$\sqrt{\left(-\frac{\omega_{\mathrm{p1}}+\omega_{\mathrm{p1}}}{2}-\frac{T_{\mathrm{o}}\omega_{\mathrm{p1}}\omega_{\mathrm{p2}}}{2\omega_z}+\omega_z\right)^2+(1+T_{\mathrm{o}})\omega_{\mathrm{p1}}\omega_{\mathrm{p2}}-\left(\frac{\omega_{\mathrm{p1}}+\omega_{\mathrm{p1}}}{2}+\frac{T_{\mathrm{o}}\omega_{\mathrm{p1}}\omega_{\mathrm{p2}}}{2\omega_z}\right)^2}$$
$$=\sqrt{(\omega_z-\omega_{\mathrm{p1}})(\omega_z-\omega_{\mathrm{p2}})}. \tag{2.11}$$

Without ω_z, two poles split vertically at the middle frequency of $-(\omega_{\mathrm{p1}}+\omega_{\mathrm{p2}})/2$.

Figure 2.11 lists approximate the PM, Q, and -3 dB bandwidth for commonly used feedback amplifiers with different poles located with angles from the real axis: Two real poles, Bessel, Butterworth, and 1 dB-ripple Chebyshev. As two poles get closer to the imaginary axis, the Q gets higher, and both frequency and transient responses peak with larger overshoot. The maximally flat Butterworth response with poles at 45° gives about a PM of 55° while the higher-Q 1 dB ripple Chebyshev response with poles at 60° gives about a PM of 45°. Therefore, Bessel poles are required to design feedback amplifiers with a PM of over 60°.

Any negative feedback systems should be stabilized. Feedback amplifiers based on opamps are stabilized by separating two poles widely by more than the loop gain. If they are not separated widely or there are more non-dominant poles, inserting a zero is the way to get extra PM. However, for switched-capacitor applications that require accurate settling, inserting zero to cancel the phase delay of the second pole which is lower than the unity loop gain frequency should be avoided.

2.1.3 Instability of Positive Feedback

Positive feedback itself doesn't warrant instability though its purpose is to make systems like latches and oscillators unstable. Therefore, the oscillation condition for positive feedback systems to start with is

$$T(j\omega_{180}) = a(j\omega_{180})f > 1. \tag{2.12}$$

Unless the condition of (2.12) is met, even positive feedback stay stable, and oscillation will never grow. If met, the noise spectrum around ω_{180} will grow since the poles are on the right half plane. However, once the oscillation hits the voltage ceiling, the oscillation magnitude stops growing further, and the magnitude starts to be clipped and limited.

Therefore, the instability should be built into the design of a tuned oscillator as shown in Fig. 2.12. The limiting goes on and stops when the fundamental filtered by the tuned amplifier inside the loop meets the steady-state oscillation condition of

$$T(j\omega_{180}) = a(j\omega_{180})f = 1. \tag{2.13}$$

Due to the finite Q of the tuned amplifier, sidebands on both sides of ω_{180} grow too until limited, which makes the phase noise spectrum around the oscillation frequency.

Two examples of positive feedback are shown in Fig. 2.13. The bistable latch is unstable at DC if it meets the oscillation condition of (2.12). It settles back to one of the bistable states quickly with the g_m/C time constant. If the parasitic capacitance of the resistive load is tuned out with an inductor, it makes an unstable oscillator at a resonant frequency if the same condition of (2.12) is met. Once variable capacitor is added, the standard VCO biased with a tail current is obtained as shown on the right side.

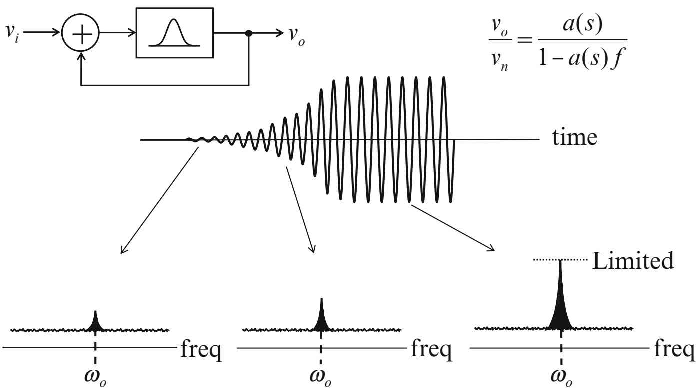


Fig. 2.12 Positive feedback example

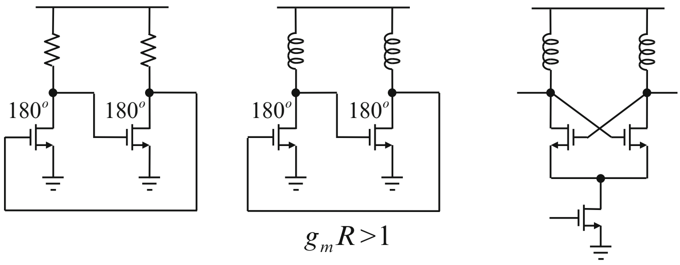
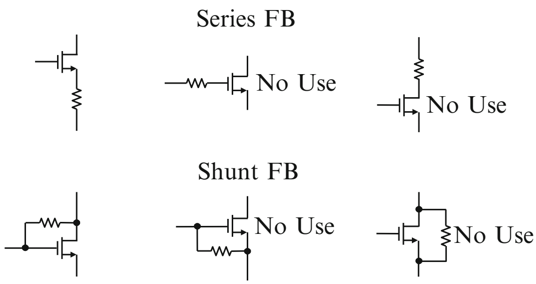


Fig. 2.13 Latch and VCO

Fig. 2.14 Series and shunt feedback examples using a resistor



2.2 Local Series and Shunt Feedbacks

Feedback comes in two different forms: Series and shunt. The former is the voltage feedback while the latter is the current feedback. Therefore, the former widens linear voltage range and raises impedance while the latter widens linear current range and lowers impedance. They are dual in concept. Simple feedback circuits can be made using one transistor and one passive component such as resistor, capacitor, and inductor.

All six combinations possible with a resistor are shown in Fig. 2.14. Among them, only source degeneration and shunt feedback are useful for local feedbacks. Other four configurations are not necessary. Similarly, only three useful feedbacks are possible with an inductor and a capacitor as shown in Fig. 2.15.

In the inductor source degeneration, the driving-point input resistance becomes real when the inductor series-resonates with the gate-source capacitance, which is used as a matching load to antenna in the LNA design. Examples of the shunt feedback using an inductor can be found in the Colpitts and Pierce oscillator designs, and the capacitive shunt feedback is an integrator often used to make a Miller capacitance.

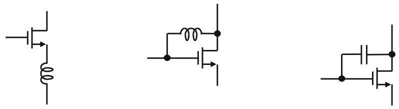


Fig. 2.15 Other useful shunt and series feedback examples

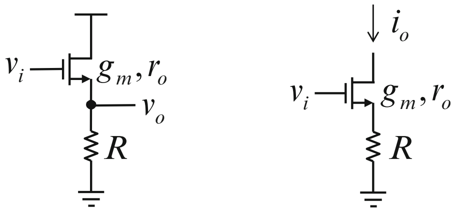


Fig. 2.16 Two examples of series feedback

2.2.1 Series Feedback

The only useful series feedback using a resistor is the source degeneration. In BJT circuits, transistors are rarely used without emitter degeneration since the input resistance looking into the base is low limited by the current gain. On the other hand, the source degeneration for MOS transistors is not necessary, but it still makes a useful circuit configuration for the following two cases: Source follower and source degeneration as shown in Fig. 2.16.

Assuming two small-signal parameters are g_m and r_o, the low-frequency small-signal closed-loop gain v_o/v_i, the forward gain a_o, the feedback gain f, and the loop gain $a_o f$ can be derived for the source follower shown on the left side as follows.

$$
\begin{aligned}
\frac{v_o}{v_i} &= \frac{g_m(r_o\|R)}{1+(g_m-g_{mb})(r_o\|R)} \approx \frac{g_m}{g_m-g_{mb}},\\
a_o &= g_m(r_o\|R), \quad f = \frac{g_m-g_{mb}}{g_m},\\
a_o f &= (g_m-g_{mb})(r_o\|R),
\end{aligned}
\tag{2.14}
$$

respectively. Unlike the BJT emitter follower, the small-signal gain of the MOS source follower is lower than unity due to the body g_{mb}, which is about 10–20 % of g_m.

The trans-conductance of the MOS transistor with the source degeneration works similarly, but the output is current drawn from the high-impedance drain side. The total trans-conductance i_o/v_i, the forward gain a_o, the feedback gain f, and the loop gain $a_o f$ can be also derived as follows.

$$\frac{i_o}{v_i} = \frac{g_m}{1 + (g_m - g_{mb})(r_o \| R)},$$
$$a_o = g_m, \quad f = \frac{g_m - g_{mb}}{g_m}(r_o \| R), \tag{2.15}$$
$$a_o f = (g_m - g_{mb})(r_o \| R),$$

respectively. Note that g_m decreases by the amount of the loop gain in series feedbacks, but its linearity improves by the same amount instead. Both cases are the same series feedback with the same loop gain. The loop gain is irrelevant of input and output ports.

In most bulk N-well processes, all NMOS bodies are tied to one substrate. In analog circuits which are not switched, the body effect of the PMOS transistor can be alleviated if the substrate is tied to its source. However, in digital circuits, even PMOS substrates are tied to the high supply voltage. If the source is floating with the body tied to the supply, the body effect raises the effective g_m to be $(g_m - g_{mb}) = (1.1–1.2)g_m$. Furthermore, since the output resistance r_o of the MOS transistor is not as high as that of bipolar transistors, the gain of the NMOS source follower is even lowered to the 0.8–0.9 level, and rarely approaches unity.

2.2.2 *Source Follower*

The source follower is a unity-gain voltage buffer. The basic function of the buffer is to transform impedance from high to low. That is, it is a light load to the input, but its low output impedance is to drive a heavy load. Its input impedance is high and mostly capacitive. Therefore, at low frequencies, the high-impedance input is open, but as frequencies go higher, the input capacitance looking into the gate is given by the Miller effect as follows.

$$C_i = (1 - A)C_{gs} + C_{gd} \approx C_{gd}, \tag{2.16}$$

where A is the source follower gain. That is, if $A = 0.9$, only about 10 % of C_{gs} loads the input. On the other hand, its output driving-point resistance is low.

$$R_o = \frac{1}{g_m - g_{mb}} \| r_o \| R \approx \frac{1}{g_m - g_{mb}}. \tag{2.17}$$

This characteristic of high input and low output impedances stays valid up to the almost device unity-gain frequency.

At high frequencies, the feedforward zero of the MOS transistor is always at g_m/C_{gs}. Note that the gate-source feedforward zero makes a negative real zero while the gate-drain feedforward zero creates a positive real zero due to the polarity inversion of the signal path. After the zero frequency, signal just bypasses the

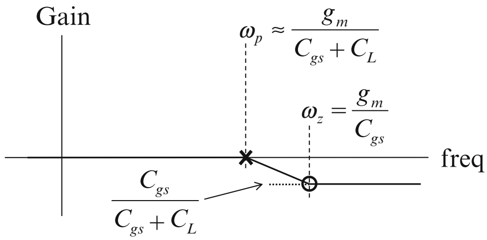


Fig. 2.17 Frequency response of the source follower

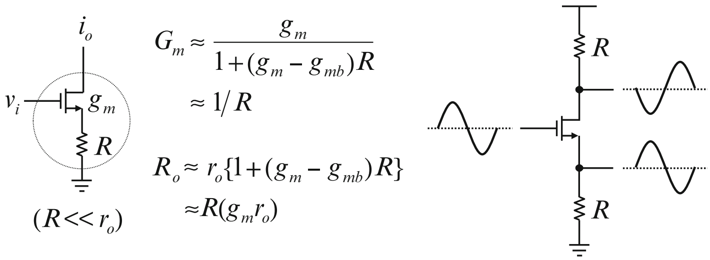


Fig. 2.18 Phase splitter using source degeneration

transistor, goes through C_{gs}, and drives the loading capacitance C_L directly, which implies that the high-frequency attenuation converges to the capacitor divider ratio of $C_{gs}/(C_{gs}+C_L)$. Therefore, the pole and the zero are separated by this gain factor, and the pole frequency is lower as sketched in Fig. 2.17.

Therefore, the frequency response can be derived as

$$\frac{v_o}{v_i}(s) = \frac{g_m R\left(1+sC_{gs}/g_m\right)}{1+(g_m-g_{mb})R+sR\left(C_{gs}+C_L\right)}. \tag{2.18}$$

If r_o is ignored and s is set to 0, it is the same equation as (2.14). The feedforward zero g_m/C_{gs} of the source follower affects its high-frequency performance as it provides a leaky forward path for signal. The feedforward signal leak is very troublesome when using it as a unity-gain buffer such as in Sallen-Key type low-pass filters that require unity-gain buffers. The output noise over wide bandwidth is also a problem when it is aliased into the signal band when the filter output is sampled.

Since the same current flows through the MOS transistor, the small-signal source and drain voltages are 180° out of phase from each other. Therefore, an RF phase splitter can be made as shown in Fig. 2.18. The source-degenerated MOS transistor encircled by the dotted line exhibits the trans-conductance of approximately $1/R$ while its driving-point output resistance is improved by the common gain factor of $g_m r_o$ as approximated.

2.2.3 Inductor Source Degeneration

The g_m of MOS transistor is usually low, but has a wider linear range than that of BJT. If degenerated with R for series feedback, the trans-conductance further decreases by the amount of the loop gain $\{1+(g_m-g_{mb})R\}$. Therefore, the input linear range of the MOS transistor is widened, and its input capacitance also gets smaller by the same factor. That is, the transistor characteristic approaches that of an ideal transistor, which has high input and output impedances plus linearized trans-conductance.

Except in broadband networking systems such as Giga-bit Ethernet and fiber, most RF circuits operate in open-loop conditions without feedback. Due to low g_m, the source degeneration is not common in low-frequency circuits such as in opamps, but it is often used in RF open-loop circuits if high linearity is required such as in the LNA and mixer. In RF circuits, inductors tune out parasitic capacitances.

Figure 2.19 shows the standard LNA with the input impedance matched to the source impedance. The source degeneration given by the inductor L makes an inductive trans-conductance device.

$$\frac{i_o}{v_i}(s)=\frac{g_m}{1+(g_m-g_{mb})\times sL+s^2LC_{gs}}=\frac{g_m}{(g_m-g_{mb})}\times\frac{1}{sL}, \tag{2.19}$$

at the resonant frequency where $s^2=-1/LC_{gs}$. It is unusual to use it as a source follower with inductor degeneration as a buffer, but the source follower gain can be obtained as

$$\frac{v_o}{v_i}(s)=\frac{g_m\times sL\left(1+sC_{gs}/g_m\right)}{1+(g_m-g_{mb})\times sL+s^2LC_{gs}}, \tag{2.20}$$

which is the same as (2.18) with the same feedforward zero at g_m/C_{gs}. When the series L and C_{gs} resonate, a real resistance remains. Let's get the voltage drop (v_i-v_o) across the gate and source. From (2.20), we obtain

$$\frac{(v_i-v_o)}{v_i}(s)=\frac{1}{1+(g_m-g_{mb})\times sL+s^2LC_{gs}}. \tag{2.21}$$

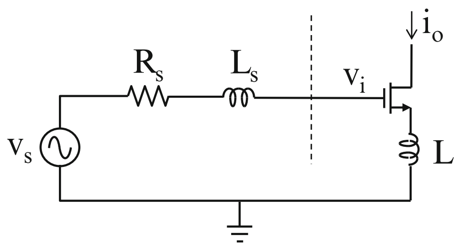


Fig. 2.19 Impedance-matched LNA with inductor source degeneration

Then from (2.21), the input impedance looking into the gate is given by

$$\frac{v_i}{i_i}(s) = \frac{v_i}{sC_{gs}(v_i - v_o)} = \frac{1 + (g_m - g_{mb}) \times sL + s^2 LC_{gs}}{sC_{gs}} = \frac{(g_m - g_{mb})L}{C_{gs}}, \quad (2.22)$$

again at the resonant frequency. This is the real resistance which can terminate the input with impedance matched to R_s. For LNA, noise figure (NF) improves with larger g_m. Therefore, the inductor L can be minimized, and an extra inductance L_s is inserted to make a resonance while keeping the impedance matched.

$$\omega_o = \frac{1}{\sqrt{(L + L_s)C_{gs}}}, \quad R_s = \frac{(g_m - g_{mb})L}{C_{gs}}. \quad (2.23)$$

Therefore, the LNA design is straightforward. $(L + L_s)$ tunes out C_{gs}, and the residual resistance $R_L + R_{Ls} + R_g + (g_m - g_{mb})L/C_{gs}$ can be matched to R_s, which is typically 50 Ω, where physical inductor and transistor gate resistances are included. Since the dominant $(g_m - g_{mb})L/C_{gs}$ doesn't contribute to noise directly, and NF can go below 3 dB. If the MOS noise is referred to the input, the NF can be approximated as follows.

$$\mathrm{NF} = 1 + \frac{2}{3} \times \frac{(\omega_o C_{gs})^2 R_s}{g_m} = 1 + \frac{2}{3} \times \left(\frac{\omega_o}{\omega_T}\right)^2 \times g_m R_s, \quad (2.24)$$

where ω_T is the device unity-gain frequency defined as g_m/C_{gs}. Large g_m/C_{gs} obtained by device scaling lowers the NF.

The gate resistance R_g can be made small by careful layout, and the series inductor resistances R_L and R_{Ls} are small. Therefore, after C_{gs} is tuned out, and $(g_m - g_{mb})L/C_{gs}$ is matched to R_s, the effective total g_m becomes

$$G_m = \frac{i_o}{v_s} = \frac{i_i \times \frac{g_m}{sC_{gs}}}{2v_i} = \frac{i_i \times \frac{g_m}{sC_{gs}}}{2v_i} = \frac{\omega_T}{\omega_o} \times \frac{1}{2R_s}, \quad (2.25)$$

which is independent of the device g_m. That is, technology scaling with smaller input capacitance will increase G_m. Therefore, the LNA design is almost set once technology is given: (1) Set the trans-conductance g_m for noise. (2) Set the overdrive voltage $(V_{gs} - V_{th})$ for the intercept point and linearity. Then, the device size and bias current are set. (3) Estimate the gate resistance. (4) Select L for matching. (5) Select L_s for input resonance. (6) Check the total resistance by estimating series resistances of L and L_s, and iterate the procedure if needed. More power is consumed with non-ideal factors such as input pad parasitic and Miller capacitances, time-variant channel charge, and hot carrier effects. Usually better NF is observed with the input resistance set lower than R_s.

2.2.4 Resistance Reflection in Series Feedback

Series voltage feedback raises impedance by the amount of the loop gain. The series resistance R can be inserted in the source and drain branches as shown in Fig. 2.20. If the body effect given by g_{mb} is ignored for simplicity, the resistance looking into the drain and source sides increases or decreases by the same amount of $g_m r_o$, which is the maximum gain obtainable from one transistor amplifier. This value of 20–40 dB varies depending on the process, channel length, and bias condition. The resistances looking into the drain and the source are r_o and $1/g_m$, respectively, without source degeneration, but they change to $R(g_m r_o)$ and $R/(g_m r_o)$, respectively. Therefore, from the drain side, the source-side resistance looks larger, but from the source side, the drain-side resistance looks smaller by the same factor [1].

The driving-point resistances of dual and triple cascode circuits are generalized in Fig. 2.21 assuming again that all devices have the same g_m and r_o. The resistance value R can be an output resistance r_o of another transistor. The highest resistance level possible in MOS circuits is limited by leakage. The output resistance offered by the triple cascode is getting closer to the highest impedance node limited by the junction leakage.

The cascode examples to get higher gain in opamps are shown in Fig. 2.22. Cascoding raises the output resistance by $g_m r_o$, thereby enhancing the gain by the same factor. The high output resistance when seen from the cascoded node is reduced by the same factor. That is, the driving-point resistance of the cascoded

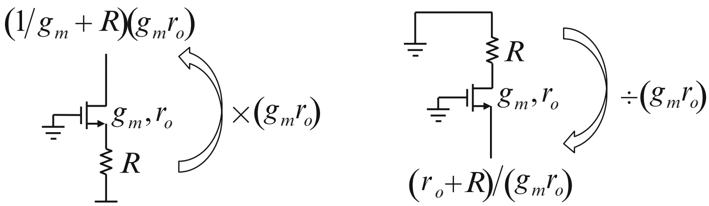
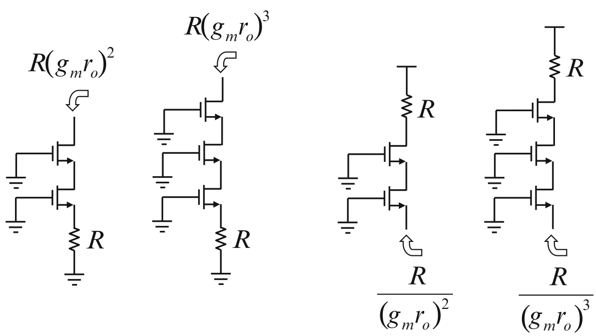


Fig. 2.20 Two useful resistance reflection rules

Fig. 2.21 Resistance reflection rules generalized



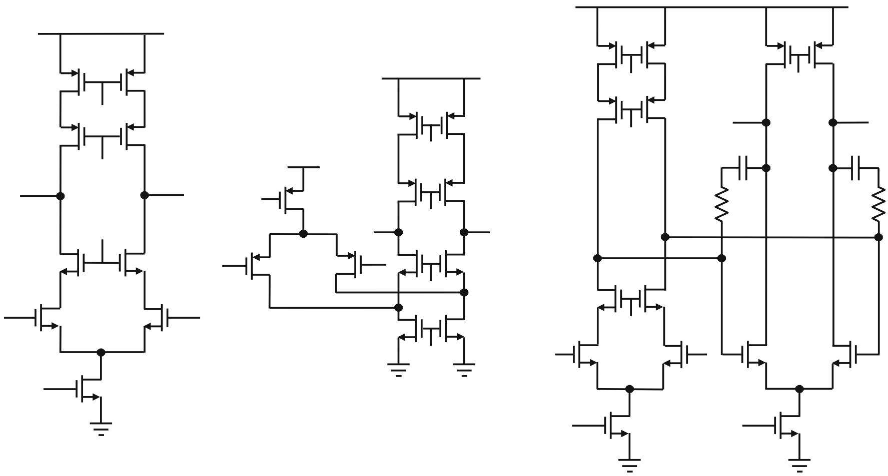


Fig. 2.22 Single-stage cascode, folded-cascode, and two-stage opamps

node approaches $1/g_m$ at high frequencies since the output node is loaded by the large capacitance to make a dominant pole. Therefore, cascoding only adds a very high frequency non-dominant pole at g_m/C_p with a parasitic capacitance C_p.

Standard opamp designs have been well established. Although the impedance level is raised for high gain by cascoding, it is difficult to cascode with low supply voltages since it requires additional DC voltage drop across the cascode device. At low-voltage uses, two-stage opamps have been preferred to single-stage opamps. In switched-capacitor applications, the input stage can be cascoded for high gain while the second stage gives high swing. In high-swing buffer applications that require high input-common mode voltages, either rail-to-rail or class AB input stages are used. However, in scaled technologies, supply voltages are still tight even for double cascoding.

To get higher gain without using multiple cascoding, a gain boosting technique based on shunt feedback can be used as sketched in Fig. 2.23. One problem that comes with the local shunt feedback is that the unity loop-gain frequency of the local feedback loop becomes a zero in the main gain path. The doublet effect on settling can be alleviated by moving the zero to higher frequencies than the unity-gain frequency of the main loop.

2.2.5 Shunt Feedback

When compared to the series voltage feedback, some of the output current is fed back to the input in the current shunt feedback, thereby reducing both the input and output driving-point impedances. The only useful shunt feedback with a resistor is the trans-resistance configuration with a resistor connected between the gate and the drain. It is compared to the source-degeneration series feedback in Fig. 2.24.

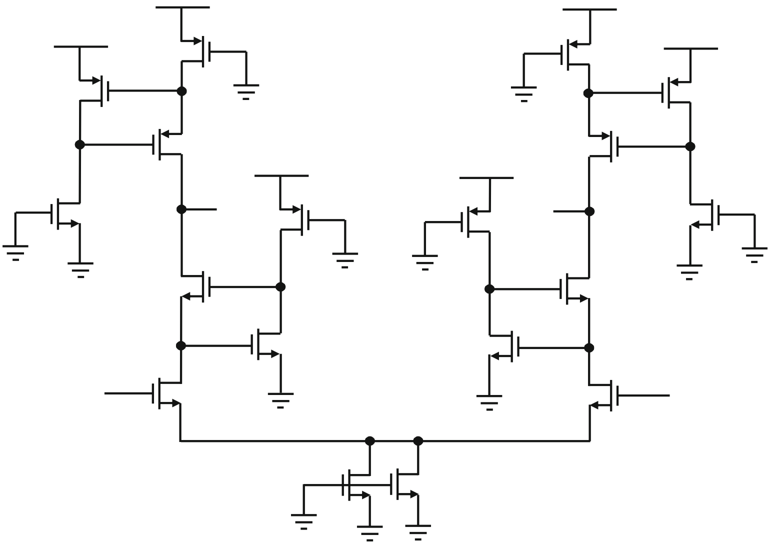
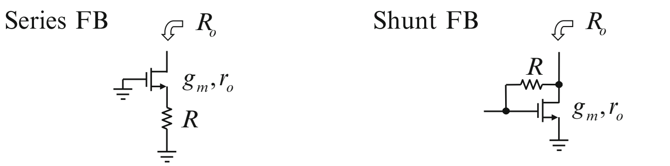


Fig. 2.23 Gain boosting example by shunt feedback



$$R_o = R + r_o\{1 + (g_m - g_{mb})R\} \approx R(g_m r_o)$$

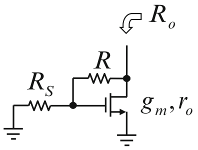
$$R_o = \frac{1}{g_m} \| r_o = \frac{r_o}{1 + g_m r_o} \approx \frac{1}{g_m}$$

Fig. 2.24 Comparison between series and shunt feedbacks

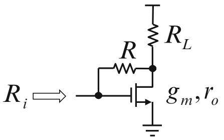
The output resistance of the shunt feedback is the parallel combination of the diode resistance and the transistor output resistance since the gate-side resistance of the MOS transistor is infinite. That is, the drain and the gate are shorted at low frequencies.

Figure 2.25 summarizes the resistance reflection rules for shunt feedbacks. The driving-point resistances looking into the output and input ports are the same diode resistance $1/g_m$ plus shunt resistance R divided by the loop gains of $g_m R_S$ and $g_m R_L$, respectively. This low resistance offered by the shunt feedback helps to broadband amplifiers. That is why the standard trans-resistance amplifier has been used for benchmarking new high-speed technologies. It has also been used to amplify low light-sensitive currents from photo diodes since it provides low impedance load to photo detector current.

This symmetry of the input and output resistance reflection rule gives a hint that the driving-point resistances looking into the input and output ports can be matched.



$$R_o = (R+R_S) \| \frac{R+R_S}{g_m R_S} \| r_o \approx \frac{R+R_S}{1+g_m R_S}$$
$$\approx \frac{1}{g_m} + \frac{R}{1+g_m R_S}$$



$$R_i = \frac{R+R_L \| r_o}{1+g_m(R_L \| r_o)} \approx \frac{R+R_L}{1+g_m R_L}$$
$$\approx \frac{1}{g_m} + \frac{R}{1+g_m R_L}$$

Fig. 2.25 Resistance reflection rules for shunt feedbacks

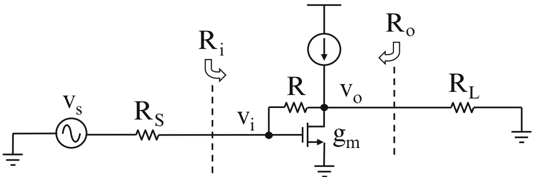


Fig. 2.26 Impedance-matched wideband amplifier

That is, wideband amplifiers with both matched input and output resistances can be implemented using a trans-resistance amplifier as shown in Fig. 2.26.

For input and output driving-point resistances to be matched, the following condition should be met.

$$R_S = R_i = \frac{R + R_L}{1 + g_m R_L} = \frac{R + R_S}{1 + g_m R_S} = R_o = R_L, \tag{2.26}$$

which gives the simple relation of

$$R = g_m R_S R_L. \tag{2.27}$$

Due to this symmetric matching characteristic, any local shunt-feedback stages can be cascaded for higher gain. Shunt feedback also improves linearity while lowering the resistance level for broadbanding. The small-signal voltage gain of the shunt-feedback stage can be derived as follows.

$$\frac{v_o}{v_i} = -\frac{g_m R R_L - R_L}{R + R_L} \approx -\frac{g_m R R_L}{R + R_L} = -g_m (R \| R_L). \tag{2.28}$$

Similarly, the gain including R_S is obtained.

$$\frac{v_o}{v_s} = -\frac{g_m R R_L - R_L}{R + R_S + R_L + g_m R_S R_L} \approx -\frac{g_m R R_L}{R + g_m R_S R_L}. \tag{2.29}$$

Alternatively, (2.29) can be approximated in two steps of attenuation and amplification as follows.

$$\frac{v_o}{v_s} = \frac{v_i}{v_s} \times \frac{v_o}{v_i} = -\frac{\dfrac{R+R_L}{1+g_mR_L}}{R_S + \dfrac{R+R_L}{1+g_mR_L}} \times \frac{g_mRR_L - R_L}{R+R_L} \approx -\frac{g_mRR_L}{R+g_mR_SR_L}. \tag{2.30}$$

At high frequencies, the input and output driving-point resistances can be matched to the standard 50 Ω, which facilitates its use as an amplifier with both the input and output ports loaded by transmission lines as shown in Fig. 2.27.

If $R_i = R_o = 50\ \Omega$, we obtain the following from (2.27) and (2.28).

$$\begin{aligned} &R = g_mR_SR_L = 2500g_m, \\ &\frac{v_o}{v_s} \approx -\frac{g_mRR_L}{R+g_mR_SR_L} = -\frac{g_mR_L}{2} = -25g_m = -\frac{R}{100}. \end{aligned} \tag{2.31}$$

Fog higher gain, both g_m and R should be set higher as follows.

$$\begin{aligned} &g_m = 1/10\ \ \Omega, \quad R = 250\ \ \Omega, \quad \text{Gain} \approx -2.5. \\ &g_m = 1/5\ \ \Omega, \quad R = 500\ \ \Omega, \quad \text{Gain} \approx -5. \\ &g_m = 1/2.5\ \ \Omega, \quad R = 1\ \ \text{k}\Omega, \quad \text{Gain} \approx -10. \end{aligned} \tag{2.32}$$

Impedance matched amplifiers are mostly to drive transmission lines or antenna ports such as in monolithic microwave integrated circuits or RF transceivers. For integrated on-chip uses, it is not necessary to match impedance for amplifier input or output ports since there are no transmission lines. Therefore, either current or voltage source is required to drive high or low impedance load, respectively, as shown in Fig. 2.28.

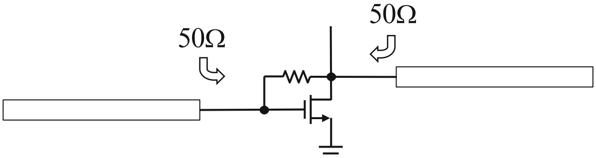


Fig. 2.27 Amplifier with matched input and output ports

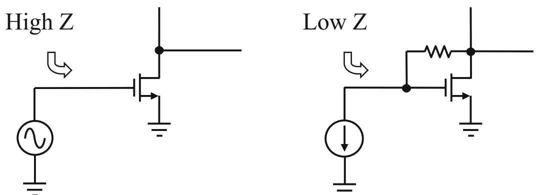


Fig. 2.28 Sources for high and impedance nodes

The stability of local series or shunt feedback is not an issue in general since only one dominant pole is involved. However, in shunt feedback, as the shunt resistance value increases, input and output nodes are separated, and make two poles in the loop. It sets the upper bound to the maximum bandwidth achievable using shunt-feedback amplifiers such as trans-resistance amplifiers. In particular, the capacitance at the input node is very critical since it lowers the pole frequency in the feedback path. The impedance at the input node is affected differently by the shunt feedback. The resistive shunt feedback lowers the input resistance while the capacitive shunt feedback increases the input capacitance due to the Miller effect as shown in the two cases of Fig. 2.29.

Note that unlike the voltage-driven opamp case, the effective gain of the shunt-feedback transistor amplifier decreases by the loading of the shunt feedback resistance. The shunt resistance from the input side looks smaller by the shunt-stage gain while the shunt capacitor looks larger by the same amount. The former helps to broadband the frequency response while the latter does the opposite as shown in Fig. 2.30.

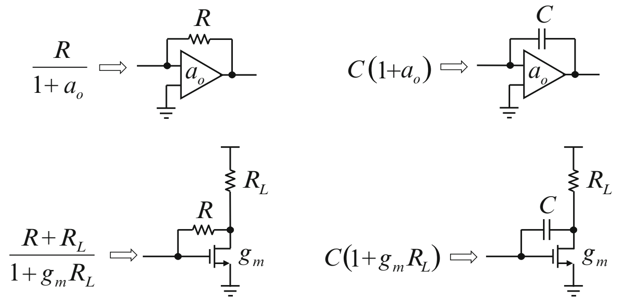


Fig. 2.29 Shunt feedback vs. Miller effect

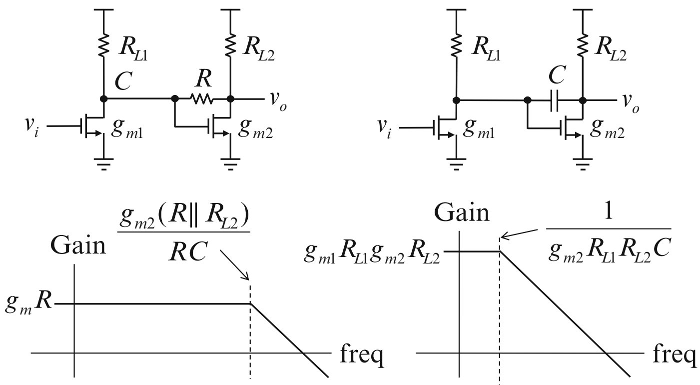


Fig. 2.30 Two-stage gain plots with shunt feedbacks

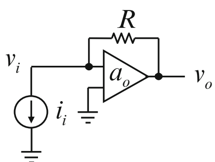
Broad-banding by shunt feedback is possible as the resistance level drops. However, its upper limit is set by the non-dominant second pole at the output node if this pole is pushed out too high. On the other hand, the Miller effect is used to frequency-compensate two-stage opamps by moving the dominant pole lower and separating two poles widely, which is called narrow-banding.

2.3 Trans-Resistance Amplifier

If the shunt-feedback amplifier is driven by the current source, it makes another useful local feedback circuit like the local series-feedback source degeneration. The input and output resistances are lowered by shunt feedback while they are raised by the series feedback. Its input resistance, voltage gain, and trans-resistance can be obtained as summarized in Fig. 2.31.

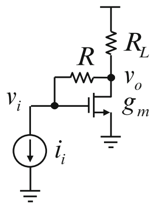
That is, the input current makes the voltage drop at the input of the trans-resistance stage, which is amplified by the voltage gain stage. Since the gain is defined as the ratio of the output voltage to the input current which has the resistance unit, it has been called trans-resistance amplifier.

Figure 2.32 illustrates the logic behind the preamplifier issue in optical receivers. Photo diodes generates low-level currents of a nA ~ μA order depending on the intensity of the incident light. To convert it into voltage, a resistor is needed to develop a voltage across it. If there is a parasitic C_D of the detector, the impedance drops after the pole frequency at $1/RC_D$. If an amplifier drives the resistance in the shunt-feedback form, the bandwidth can be widened by the loop gain $(1 + a_o)$. The parasitic capacitance C_D at the detector input node is the most important parameter to consider in the trans-resistance amplifier design. For higher output current, the size of photo detector should be made large. Then large diode gives large parasitic capacitance C_D.



$$\frac{v_i}{i_i} = \frac{R}{1+a_o}, \quad \frac{v_o}{v_i} = a_o$$

$$\frac{v_o}{i_i} = \frac{v_i}{i_i} \times \frac{v_o}{v_i} = \frac{a_o R}{1+a_o} \approx R - \frac{R}{a_o}$$



$$\frac{v_i}{i_i} = \frac{R+R_L}{1+g_m R_L}, \quad \frac{v_o}{v_i} = \frac{g_m R R_L - R_L}{R+R_L}$$

$$\frac{v_o}{i_i} = \frac{v_i}{i_i} \times \frac{v_o}{v_i} = \frac{g_m R R_L - R_L}{1+g_m R_L} \approx R - \frac{1}{g_m}$$

Fig. 2.31 Trans-resistance amplifier

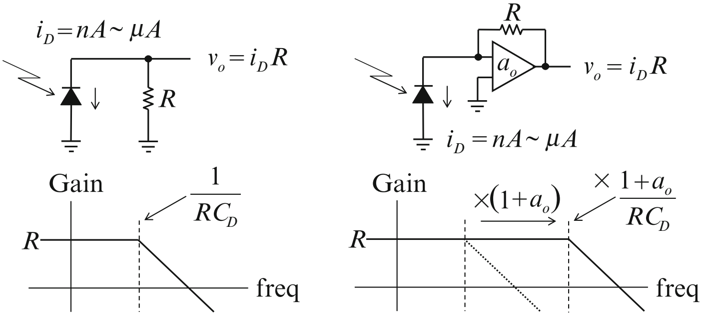
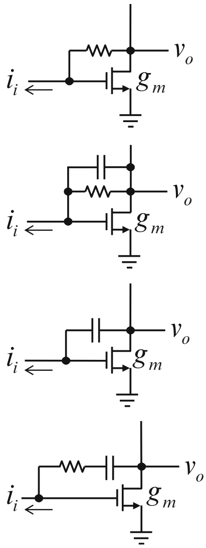


Fig. 2.32 Trans-resistance amplifier for optical receivers



$$\frac{v_o}{i_i} \approx R \times \left(1 - \frac{1}{g_m R}\right)$$

$$\frac{v_o}{i_i} \approx R \times \left(1 - \frac{1}{g_m R} - \frac{sC}{g_m}\right) \times \frac{1}{(1+sRC)}$$

$$\frac{v_o}{i_i} \approx \frac{1}{sC} \times \left(1 - \frac{sC}{g_m}\right)$$

$$\frac{v_o}{i_i} \approx \frac{1}{sC} \times \left(1 + sRC - \frac{sC}{g_m}\right)$$

Fig. 2.33 Trans-impedances of four local shunt feedbacks

There are four possible cases of local shunt feedbacks with R and C as shown in Fig. 2.33. The first one is the standard trans-resistance amplifier. The third one is a Miller integrator. Since the signal is fed forward through the feedback capacitor into the inverting output, it makes a right-half plane zero. The fourth one is a straightforward voltage sum, and the series resistance with a capacitor makes a left-half-plane zero. By setting the RC value to be C/g_{m} or higher the right-half plane zero can be canceled or moved to the left-half plane. The second one is a current sum of two paths, which gives a pole and a right half-plane zero. It is the most demanding task for analog designers to derive the frequency response of this two-pole amplifier. Driving the shunt-feedback using a high-impedance current source complicates the hand analysis as two poles interact as shown in Fig. 2.34.

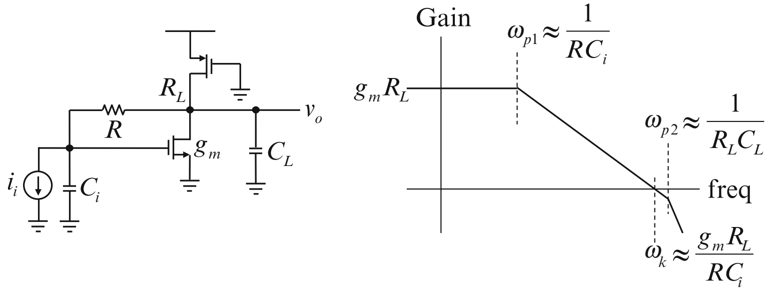


Fig. 2.34 Trans-resistance amplifier and open-loop gain

There are parasitic capacitances at the input and the output nodes marked as C_i and C_L ignoring the feedforward capacitance. The open-loop gain can be considered to analyze stability as follows.

$$\begin{aligned} a(s)f &= \frac{g_m}{\dfrac{1}{R_L} + sC_L + \dfrac{sC_i}{1+sRC_i}} \times \frac{1}{1+sRC_i} \\ &= \frac{g_m R_L}{1 + s\{RC_i + R_L(C_i + C_L)\} + s^2 R R_L C_i C_L}. \end{aligned} \tag{2.33}$$

This quadratic equation from the denominator cannot be factored algebraically. If $R \gg R_L$ is true, the $R_L C_i$ term can be ignored, and two factored poles become negative real. That is, they can be separated by more than the DC loop gain $g_m R_L$ for stability as explained in Fig. 2.34. Otherwise, two poles become complex conjugate poles on a circle with a radius equivalent to the geometric mean.

$$\omega_o = \sqrt{\omega_{p1}\omega_{p2}} \approx \sqrt{\frac{1}{RC_i} \times \frac{1}{R_L C_L}}. \tag{2.34}$$

To stay stable, the feedback loop pole frequency of $1/RC_i$ should be much lower than the output pole frequency of $1/R_L C_L$, which is often limited by the speed of the process technology.

Using (2.33), the closed-loop transfer function can be also derived as follows.

$$\frac{v_o}{i_i}(s) = \frac{g_m R R_L - R_L}{1 + g_m R_L + s\{RC_i + R_L(C_i + C_L)\} + s^2 R R_L C_i C_L}. \tag{2.35}$$

Now the closed-loop poles become even higher-Q poles as they move vertically further into the complex plane. They are on a circle with a radius of

$$\omega_o \approx \sqrt{(1 + g_m R_L) \times \frac{1}{RC_i} \times \frac{1}{R_L C_L}}. \tag{2.36}$$

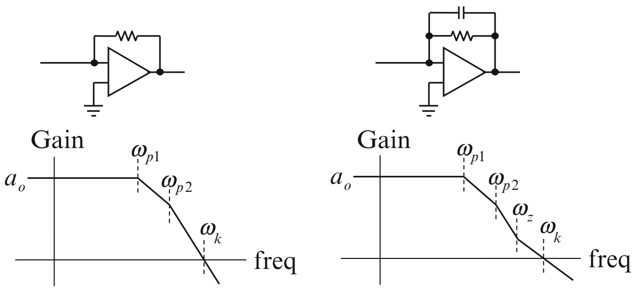


Fig. 2.35 Shunt feedbacks without and with a zero

The stability condition of two-pole networks can be approximately defined as $\omega_k < \omega_{p2}$ in open loop for the PM to be greater than 45°. That is,

$$R_L C_L < \frac{RC_i}{g_m R_L}, \tag{2.37}$$

which is difficult to meet in most wideband amplifier designs. The desirable solution is to add a zero below the unity loop-gain frequency. In general, the extra loop delay in the feedback path should be cancelled with a real zero inserted before the unity loop-gain frequency as explained in Fig. 2.35.

If the shunt feedback resistor is bypassed by a capacitor, it makes a zero in the open-loop gain at $1/RC$. Then the open-loop and closed-loop gains of (2.33) and (2.35) are modified as follows including C.

$$\begin{aligned} a(s)f &= \frac{g_m}{\dfrac{1}{R_L} + sC_L + \dfrac{sC_i(1+sRC)}{1+sR(C+C_i)}} \times \frac{1+sRC}{1+sR(C+C_i)} \\ &= \frac{g_m R_L(1+sRC)}{1+s\{R(C+C_i)+R_L(C_i+C_L)\}+s^2RR_L(CC_i+CC_L+C_iC_L)}. \end{aligned} \tag{2.38}$$

$$\frac{v_o}{i_i}(s) = \frac{g_m R_L R\left(1-\dfrac{1}{g_m R}-\dfrac{sC}{g_m}\right)}{1+g_m R_L + s[R\{C_i+(1+g_m R_L)C\}+R_L(C_i+C_L)]+s^2RR_L(CC_i+CC_L+C_iC_L)}.$$

These equations further complicate the assessment of stability with greater complexity, but one thing to note is that two poles are moving farther from the imaginary axis as the first-order term increases as the loop gain increases. It implies the Q of the poles gets lower, but the right half-plane zero is created in the close-loop gain due to feedforward through the capacitor C.

If $R \gg R_L$ is true again as before, the $R_L C_i$ term in the denominator can be ignored, and two factored poles in the open-loop gain become negative real as expected.

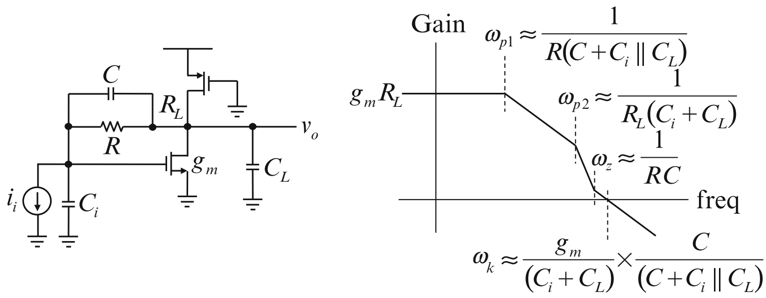


Fig. 2.36 Trans-resistance amplifier with shunt compensation and open-loop gain

$$\omega_{\mathrm{p1}} \approx \frac{1}{R(C + C_{\mathrm{i}} \| C_{\mathrm{L}})} \quad \text{and} \quad \omega_{\mathrm{p2}} \approx \frac{1}{R_{\mathrm{L}}(C_{\mathrm{i}} + C_{\mathrm{L}})}, \tag{2.39}$$

where $(C_{\mathrm{i}} \| C_{\mathrm{L}})$ denotes the value of $C_{\mathrm{i}}C_{\mathrm{L}}/(C_{\mathrm{i}} + C_{\mathrm{L}})$ for the series connection of two capacitors. The open-loop gain is shown in Fig. 2.36.

Again for the PM to be greater than 45° for this case, $\omega_z < \omega_k$ so that the zero frequency can be lower than the unity loop-gain frequency. That is,

$$RC > \frac{C + C_{\mathrm{i}} \| C_{\mathrm{L}}}{C} \times \frac{C_{\mathrm{i}} + C_{\mathrm{L}}}{g_{\mathrm{m}}}, \tag{2.40}$$

which can be easily met. Otherwise, two open-loop poles become complex conjugate high-Q poles on a circle with a radius of

$$\omega_{\mathrm{o}} = \sqrt{\omega_{\mathrm{p1}}\omega_{\mathrm{p2}}} \approx \sqrt{\frac{1}{R(C + C_{\mathrm{i}} \| C_{\mathrm{L}})} \times \frac{1}{R_{\mathrm{L}}(C_{\mathrm{i}} + C_{\mathrm{L}})}}. \tag{2.41}$$

Similarly, the closed-loop poles would move to a new circle but with a lower Q since the zero pulls the poles away from the imaginary axis.

$$\omega_{\mathrm{o}} \approx \sqrt{(1 + g_{\mathrm{m}}R_{\mathrm{L}}) \times \frac{1}{R(C + C_{\mathrm{i}} \| C_{\mathrm{L}})} \times \frac{1}{R_{\mathrm{L}}(C_{\mathrm{i}} + C_{\mathrm{L}})}}. \tag{2.42}$$

This is the same result as obtained by the famous pole-splitting Miller effect of a two-pole system. If the Miller capacitance C is very large and there is no shunt feedback, the dominant pole is generated by the Miller capacitance at the input, and the non-dominant pole is created by the sum of input and output capacitances driven by the trans-conductance. From (2.38), two poles can be derived as follows using the dominant pole approximation and the geometric mean.

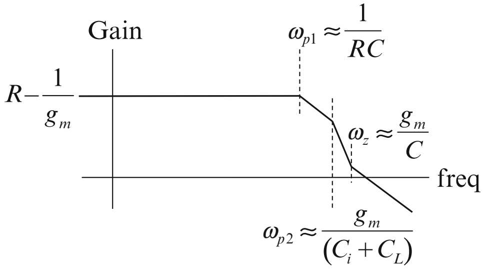


Fig. 2.37 Trans-resistance amplifier gain with two widely separated poles

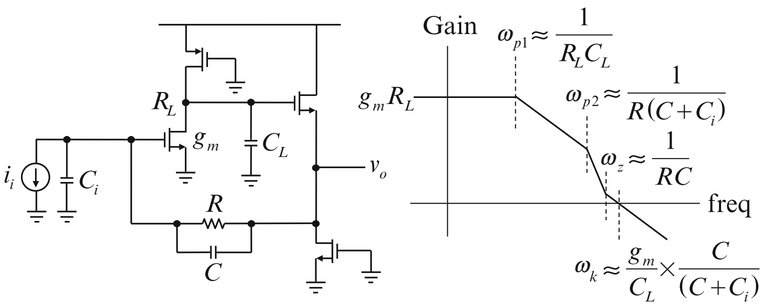


Fig. 2.38 Buffered trans-resistance amplifier and open-loop gain

$$\begin{aligned}
\omega_{\mathrm{p1}} &\approx \frac{1 + g_{\mathrm{m}} R_{\mathrm{L}}}{R\{C_{\mathrm{i}} + (1 + g_{\mathrm{m}} R_{\mathrm{L}})C\}} \approx \frac{1}{RC}, \\
\omega_{\mathrm{p2}} &\approx \frac{C_{\mathrm{i}} + (1 + g_{\mathrm{m}} R_{\mathrm{L}})C}{R_{\mathrm{L}}(CC_{\mathrm{i}} + CC_{\mathrm{L}} + C_{\mathrm{i}}C_{\mathrm{L}})} \approx \frac{g_{\mathrm{m}}}{C_{\mathrm{i}} + C_{\mathrm{L}}}.
\end{aligned} \tag{2.43}$$

The only difference in this shunt feedback is that the Miller pole is now at $-1/RC$ due to the shunt-feedback resistor. The frequency response of the trans-resistance amplifier with two widely separated real open-loop poles is sketched in Fig. 2.37.

The trans-resistance amplifier with feedforward compensation offers a desirable very high-frequency dominant pole, and the input and output capacitances are driven by the diode resistance of $1/g_{\mathrm{m}}$. It is because the feedforward path goes through the shunt capacitance. One way to eliminate the right half-plane zero and make the non-dominant second pole independent of the input capacitance is to use a feedback buffer amplifier. The low impedance of the buffer amplifier output can stop the signal feedforward, but the feedback path is not affected. Therefore, there is no right half-plane zero created, and the trans-conductance doesn't need to drive the input capacitance.

The buffered trans-resistance amplifier is shown along with its open-loop gain in Fig. 2.38. In most multi-stage wideband amplifier designs, it is common to use a source-follower buffer for Miller capacitance feedback. Assuming that the source follower has an ideal unity gain and the pole at the source follower output is high enough to ignore, the open- and closed-loop gains can be obtained as follows.

$$a(s)f = \frac{g_m R_L (1 + sRC)}{(1 + sR_L C_L)\{1 + sR(C + C_i)\}}.$$
$$\frac{v_o}{i_i}(s) = \frac{g_m R_L R}{1 + g_m R_L + s[R\{C_i + (1 + g_m R_L)C\} + R_L C_L] + s^2 R_L R (C + C_i) C_L}. \tag{2.44}$$

Note that there is no right half-plane zero, and both the open-loop gain and the closed-loop gain are greatly simplified.

If $R \gg R_L$ is true, two factored poles are

$$\omega_{p1} \approx \frac{1}{R_L C_L} \quad \text{and} \quad \omega_{p2} \approx \frac{1}{R(C + C_i)}, \tag{2.45}$$

respectively, as shown in Fig. 2.38. For PM to be greater than 45°,

$$g_m R > \frac{C_L}{C} \times \frac{(C + C_i)}{C}, \tag{2.46}$$

which can be easily met. The closed-loop poles would move to a new circle but with a lower Q since the zero pulls the poles away from the imaginary axis.

$$\omega_o \approx \sqrt{(1 + g_m R_L) \times \frac{1}{R(C + C_i)} \times \frac{1}{R_L C_L}}. \tag{2.47}$$

If the Miller capacitance C is very large, two familiar widely separated poles can be approximated as follows.

$$\begin{aligned} \omega_{p1} &\approx \frac{1 + g_m R_L}{R\{C_i + (1 + g_m R_L)C\}} \approx \frac{1}{RC}, \\ \omega_{p2} &\approx \frac{C_i + (1 + g_m R_L)C}{R_L (C + C_i) C_L} \approx \frac{g_m}{C_L}. \end{aligned} \tag{2.48}$$

The frequency response of this case is shown in Fig. 2.39.

The shunt feedback implements wideband amplifiers with both low input and output impedances, but makes an extra pole, thereby requiring frequency compensation for stability. It is possible to further broadband the shunt feedback amplifier with dual or triple gain stages as shown in Fig. 2.40.

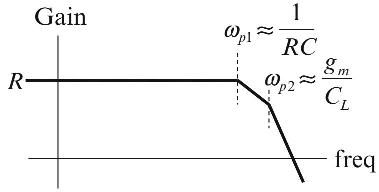


Fig. 2.39 Buffered trans-resistance amplifier gain with two widely separated poles

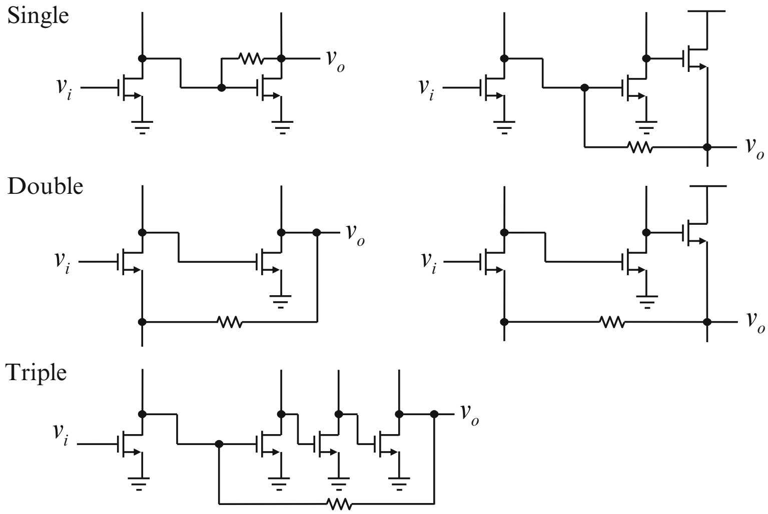


Fig 2.40 Trans-resistance amplifiers with single, double, and triple gain stages

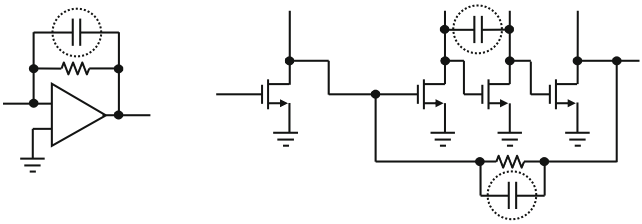


Fig. 2.41 Feedforward frequency compensations

However, it is a formidable task to stabilize three- or four-pole response though the buffered feedback helps to reduce the feedforward effect. There is no simple way, but the common wisdom is to insert as many feedforward zeros that bypass gain stages like using the Miller capacitance. Since zeros should be added after poles, the gain attainable by extra poles is limited and incremental. Stabilizing the loop with multiple integrators in $\Delta\Sigma$ modulators is a good example of the feedforward compensation.

Examples of the feedforward compensation are shown in Fig. 2.41. It is to lower the path impedance between the two nodes, and to let the signal directly pass through the capacitor at frequencies higher than $1/RC$ or g_m/C, which creates a zero effect by definition. If it bypasses the inverting signal, the zero moves to the right half plane. There are three ways to cancel the right half-plane zero. The source follower feedback cuts the feedforward path, but its pole in the feedback loop creates left-half plane zero. The G_m boosting moves the zero to higher frequencies, but extra pole inside the local feedback loop for G_m boosting complicates the overall settling. Lastly, the right half-plane zero is canceled and moved to the left half plane by just adding a resistance in series with the capacitor.

2.4 G_m Boosting and Noise Cancellation

A need arises to make an effective trans-conductance larger than it actually is for buffering and low noise. Active shunt feedback either lowers the resistance level, or boosts the conductance level as shown in Fig. 2.42.

When looking into the input side, the shunt feedback resistance looks smaller by the amplifier gain. By active shunt feedback, the input conductance is made very small. If the output is taken from the shunt transistor, this common-source (CS) and common-drain (CD) stages can be used to boost the output resistance of the series feedback as shown in Fig. 2.43.

It also shows a G_m-boosted source follower with CS and common-gate (CG) feedback. The former raises the output resistance by the loop gain as sketched

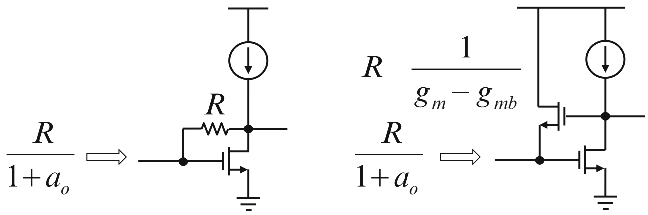


Fig. 2.42 Active shunt-feedback by G_m boosting

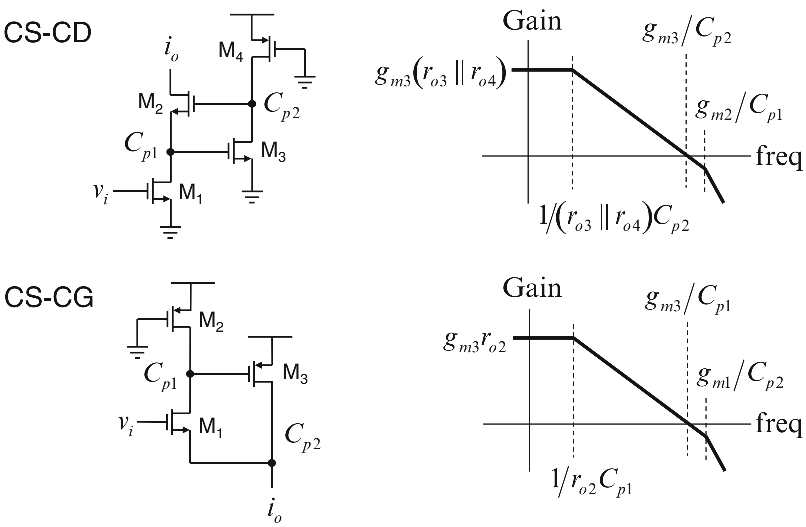


Fig. 2.43 Two examples of active G_m boosting

while the latter lowers the output resistance by the loop gain. Boosted transconductance helps to lower impedance and widen bandwidth. That is, their transconductances are further increased to be

$$(g_{m2} - g_{mb2}) \times g_{m3}(r_{o3} \| r_{o4}) \quad \text{and} \quad g_{m1} \times g_{m3} r_{o2}, \tag{2.49}$$

respectively. Due to these shunt feedback gains, the gain-boosted stage and the super-G_m source follower are made practical overcoming the handicap of low transconductance values of MOS transistors.

Examples of two super-G_m source followers are shown in Fig. 2.44. Due to the body effect, their gains approach the same $g_{m1}/(g_{m1} - g_{mb1})$ as the regular source follower, but their output resistances are made much lower enhancing the load drive capacity greatly.

Noise sources in feedback circuits are shown in Fig. 2.45. Feedback only enhances analog performance limited by deterministic parameters, but noise is a random power with a variance with no magnitude information. Therefore, all noise powers in feedback networks are added without being lowered by the negative feedback. Noise is further enhanced in high-Q circuits like resonators.

The strategies to achieve low noise in open-loop LNA are mostly of two kinds. One is to make the effective G_m higher than real G_m, which contributes to actual noise, and the other is noise cancellation. Narrow-banding is another way, but system requirements set the bandwidth. Oversampling lowers the in-band noise, but pays the speed penalty.

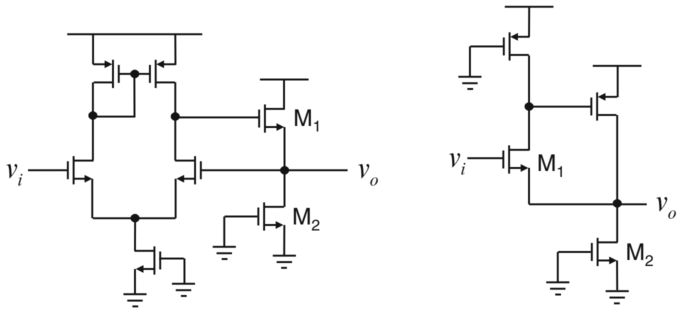


Fig. 2.44 Two G_m-boosted source follower examples

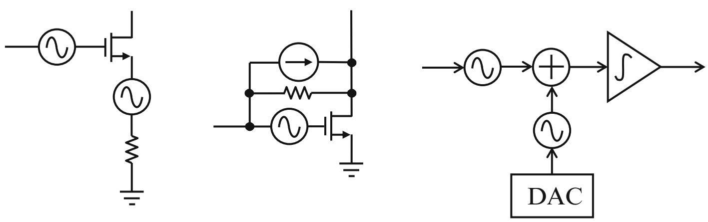


Fig. 2.45 Two G_m-boosted source follower examples

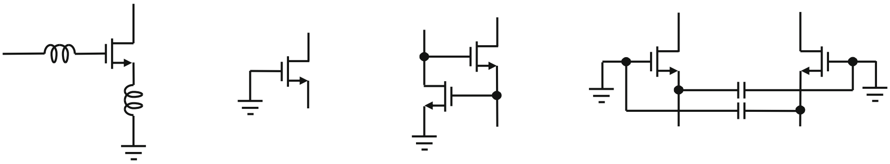


Fig. 2.46 Low-noise amplifier schemes

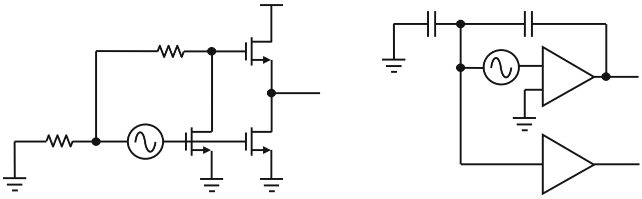


Fig. 2.47 Concept of noise cancellation by feedforward

Widely used low-noise techniques that increase the effective G_m are shown in Fig. 2.46. The passive inductor degeneration is very effective to achieve low noise. The CG amplifier has a factor of 2/3, but the empirical filling factor doesn't justify its effectiveness. The G_m boosting enhances the effective G_m, but the feedback amplifier contributes some noise. Lastly, multiplying the signal by feeding forward through capacitors doubles the input swing, but capacitors get larger and extra power is demanded. If G_m enhancement is by adding gain stages, they also add noise and power. There are no obvious solutions to LNA designs, but new process improves the noise performance incrementally.

Alternatively, feedforward can cancel the in-band noise as sketched in Fig. 2.47. It is a two-path system for noise [2, 3]. Although the noise polarity is not known, one noise source can be amplified through two identical gain paths with inverting and non-inverting gains, and summed later. The end result of this summing is the cancellation of the in-band noise of one source. If two-path gains are matched, everything is cancelled, but noise and offset of the cancelling path remain. That is, out-of-band noise and the noise of the additional amplifier are not cancelled. When designing such noise-cancelling amplifiers, the difficulty also lies in achieving large signal linearity.

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