

### 2.6.1 Resistor Series Peaking

It was shown in [43, 44] that a resistor can be added between the gates of the input and output transistors of the basic current amplifier, as shown in Fig.2.14, to introduce a zero to the system. The current transfer function is given by

$$H(s) = \left( \frac{g_{m2}}{g_{m1}} \right) \frac{sRC_{gs1} + 1}{s^2 \frac{RC_{gs1}C_{gs2}}{g_{m1}} + s \frac{C_{gs1} + C_{gs2}}{g_{m1}} + 1}. \quad (2.30)$$

The two poles are at

$$p_{1,2} = \frac{C_{gs1} + C_{gs2}}{2RC_{gs1}C_{gs2}} \left[ -1 \pm \sqrt{1 - \frac{4RC_{gs1}C_{gs2}g_{m1}}{(C_{gs1} + C_{gs2})^2}} \right], \quad (2.31)$$

and the zero is at  $z = -\frac{1}{RC_{gs1}}$ . Depending upon the value of the peaking resistor  $R$ , the locations of both the zero and poles of the system differ and the amplifier exhibits distinct characteristics. (1) *Distinct real poles* - When  $R = \frac{1}{g_{m1}}$ , the amplifier has two distinct negative real poles located at  $p_1 = -\frac{g_{m1}}{C_{gs2}}$  and  $p_2 = -\frac{g_{m1}}{C_{gs1}}$ . Observe that  $p_2$  is identical to the zero given by  $z = -\frac{g_{m1}}{C_{gs1}}$  and cancels out the zero. As a result, the transfer function is simplified to

$$H(s) = \left( \frac{g_{m2}}{g_{m1}} \right) \frac{1}{s \frac{C_{gs2}}{g_{m1}} + 1}. \quad (2.32)$$

As compared with the bandwidth of the basic current amplifier, the resistor series peaking with two distinct real poles boosts the bandwidth to  $\omega_b = \frac{g_{m1}}{C_{gs2}}$ . Note that if  $C_{gs2} \gg C_{gs1}$ , the bandwidth improvement from resistive series peaking with two distinct real poles is rather small. (2) *Identical real poles* - When  $R = \frac{1}{4g_{m1}} \frac{(C_{gs1} + C_{gs2})^2}{C_{gs1}C_{gs2}} \approx \frac{1}{4g_{m1}} \frac{C_{gs2}}{C_{gs1}}$ , the circuit has two identical negative real poles  $p_{1,2} = -\frac{2g_{m1}}{C_{gs1} + C_{gs2}}$ . The bandwidth becomes  $\omega_b = 2\sqrt{\sqrt{2} - 1} \frac{g_{m1}}{C_{gs2}}$ . (3) *Complex conjugate poles* - A further increase of  $R$  will lead to a pair of complex conjugate poles. An overshoot in the frequency-domain response and ringing in the time-domain response exist. A critical point is when  $R = \frac{1}{2g_{m1}} \frac{C_{gs2}}{C_{gs1}}$ . The amplifier has two complex conjugate poles that are separated by  $\frac{\pi}{2}$  and has a maximally flat response, called Butterworth response, with the bandwidth given by  $\omega_b = \sqrt{2} \frac{g_{m1}}{C_{gs2}}$  [45]. In this case, the time-domain response of the

amplifier to a unit-step input has an overshoot of 4.32% approximately [46]. Fig. 2.15 shows the dependence of the bandwidth of the current amplifier with the resistor series peaking on the resistance of the peaking resistor.

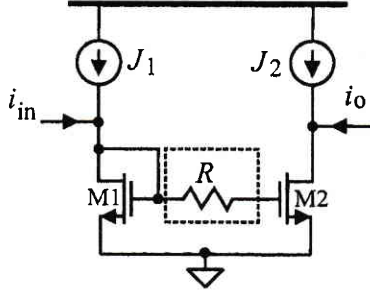


Figure 2.14. Basic current amplifier with resistor series peaking.

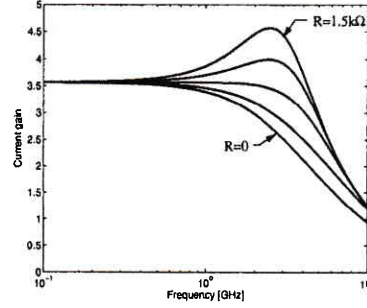


Figure 2.15. Simulated frequency response of current amplifier with resistor series peaking.  $R$  is varied from 0 to 1.5 k $\Omega$  with step 0.375 k $\Omega$ . The amplifier is implemented in TSMC-0.18 $\mu$ m CMOS technology.

### 2.6.2 Inductor Series Peaking

The thermal noise of the series peaking resistor increases the total noise of the amplifier. For low-noise applications, such as the front-end of Gb/s transceivers and optical pre-amplifiers, noiseless elements, such as inductors, are preferred over noisy resistors for bandwidth enhancement. It has been demonstrated that inductor shunt peaking can increase the bandwidth of voltage-mode circuits by as much as 70% [47, 48]. Inductor shunt-peaking technique, however, is not particularly applicable to current-mode circuits due to the existence of biasing current sources between the devices forming the dominant poles and the supply voltage. The fact that the dominant pole of the basic current amplifier is located at the gates of  $M_1$  and  $M_2$  suggests that an inductor can be placed between the gates of  $M_1$  and  $M_2$ , as shown in Fig. 2.16, to boost bandwidth. By assuming  $C_{gs2} \gg C_{gs1}$ , we obtain the current transfer function

$$\frac{I_o(s)}{I_{in}(s)} = \left( \frac{g_{m2}}{g_{m1}} \right) \frac{1}{s^2 LC_{gs2} + s \frac{C_{gs2}}{g_{m1}} + 1}. \quad (2.33)$$

The two poles are located at

$$p_{1,2} = \frac{1}{2Lg_{m1}} \left( -1 \pm \sqrt{1 - \frac{4Lg_{m1}^2}{C_{gs2}}} \right). \quad (2.34)$$

Under the condition  $L = \frac{C_{gs2}}{4g_{m1}^2}$ , the amplifier has two identical real poles  $p_{1,2} = -\frac{2g_{m1}}{C_{gs2}}$ . Its time response is critically damped with no ringing.

The bandwidth in this case is given by  $\omega_b = 2\sqrt{\sqrt{2} - 1} \left( \frac{g_{m1}}{C_{gs2}} \right)$ . When  $L$  is increased to  $L = \frac{C_{gs2}}{2g_{m1}^2}$ , the amplifier has two complex conjugate poles that are separated by  $\frac{\pi}{2}$ . In this case, the response of the amplifier has a maximally flat (Butterworth) passband with the bandwidth given by  $\omega_b = \sqrt{2} \left( \frac{g_{m1}}{C_{gs2}} \right)$  [45]. Fig.2.17 shows the frequency response of the current amplifier with inductor series peaking. The attenuation rate of the response in the stop band, which is approximately -40 dB/decade, is higher as compared with that of the amplifier with resistive series peaking. This is because the former increases the bandwidth by adding a pole whereas the latter enhances the bandwidth by introducing a zero. The peaking inductor does not affect the dc characteristics of the amplifier, a similar property as that of the resistor series peaking.

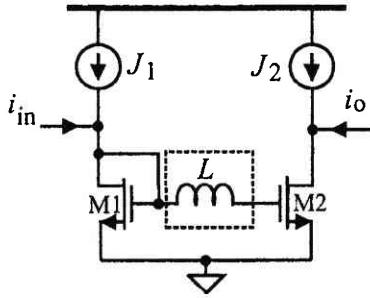


Figure 2.16. Inductor series peaking in current-mode circuits.

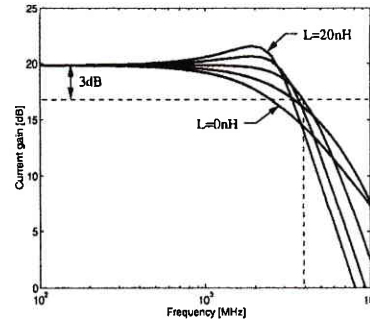


Figure 2.17. Simulated frequency response of current amplifier with inductor series peaking. The value of the series peaking inductor is varied from 0 to 20nH with step 5nH. The amplifier is implemented in TSMC-0.18 $\mu$ m CMOS technology.

On-chip inductors are usually implemented in either planar or stacked spiral configurations, as shown in Fig.2.18. Spiral inductors have the characteristics of a low quality factor, a low inductance, and extremely area-consuming [47–49]. The ohmic loss of spiral inductors, mainly due to the skin-effect induced loss at high frequencies, is usually depicted

using a series resistor  $R_s$ . Its capacitive loss, arising from the large capacitance between the spirals and the substrate, is represented by two shunt capacitors  $C_{ox}$  at the terminals of the inductor. The capacitive coupling between the upper and lower spirals at the under-paths and fringe capacitance between neighboring spirals is characterized by  $C_s$ , as shown in the simple lumped model of on-chip spiral inductors in Fig.2.19.

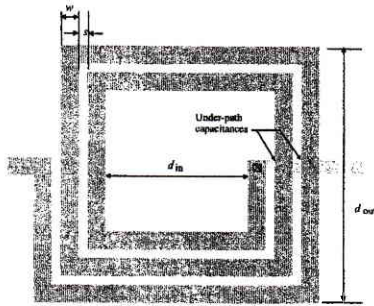


Figure 2.18. Typical layout of square-shaped spiral inductors.

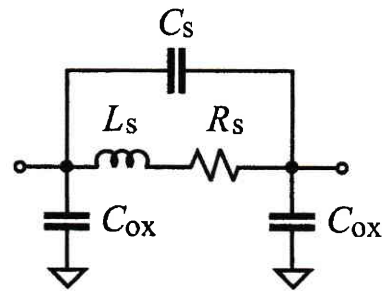


Figure 2.19. Simple lumped model of on-chip spiral inductors.

The series resistance of the inductor behaves as a series peaking resistor and improves the bandwidth, as shown in Fig.2.20. The spiral-substrate shunt capacitance, however, is directly added to the total capacitance of the gates of  $M_{1\sim2}$  of the current amplifier, lowering the bandwidth, as evident in Fig.2.21.

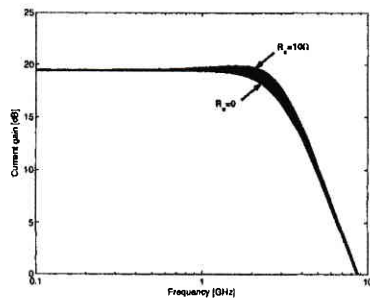


Figure 2.20. Simulated effect of the series resistance of series peaking inductors on the frequency response of the current amplifier implemented in TSMC-0.18 $\mu$ m CMOS technology.

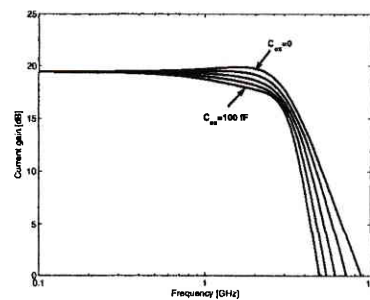


Figure 2.21. Simulated effect of the parasitic capacitance of series peaking inductors on the frequency response of the current amplifier implemented in TSMC-0.18 $\mu$ m CMOS technology.

### 2.6.3 Current Feedback

It is well known that negative current-current feedback increases the output impedance and lowers the input impedance [50, 51]. To sense the output current without affecting both the dc biasing condition and the supply voltage, the current feedback mechanism shown in Fig.2.22 can be used. The transfer function of the amplifier is given by

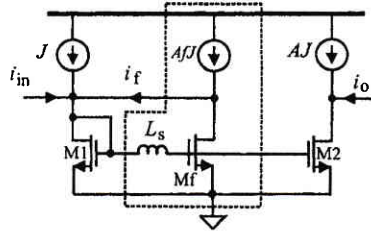


Figure 2.22. Current amplifier with both current-current feedback and inductor series peaking.

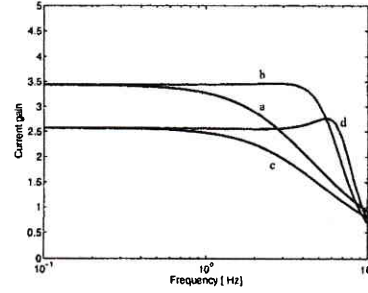


Figure 2.23. Simulated frequency response of current amplifiers with both current-current feedback and inductor series peaking. (a) Basic current amplifier; (b) Inductor series peaking only; (c) Current-current feedback; (d) Inductor series peaking and current-current feedback. The amplifiers are implemented in TSMC-0.18 $\mu$ m CMOS technology.

$$H(s) \approx \left( \frac{g_{m2}}{g_{m1}} \right) \frac{\frac{1}{LC_{gs2}}}{s^2 + s \frac{1}{g_{m1}L} + \frac{g_{m1} + g_{mf}}{g_{m1}LC_{gs2}}}, \quad (2.35)$$

with two poles at

$$p_{1,2} = \frac{1}{2g_{m1}L} \left[ -1 \pm \sqrt{1 - \frac{4(g_{m1} + g_{mf})g_{m1}L}{C_{gs2}}} \right]. \quad (2.36)$$

Complex conjugate poles that are  $\frac{\pi}{2}$  apart occurs when  $L = \frac{C_{gs2}}{2g_{m1}^2(1+Af)}$ , where  $A = \frac{(W/L)_2}{(W/L)_1}$  and  $f = \frac{(W/L)_f}{(W/L)_2}$ . In this case, the amplifier has a maximally flat response with the bandwidth  $\omega_b = \sqrt{2}(1+Af)\frac{g_{m1}}{C_{gs2}}$ . The value of the series peaking inductor that gives a maximally flat response is reduced from  $L = \frac{C_{gs2}}{2g_{m1}^2}$  without the current-current feed-