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Academic Year 2023– 2024 / Even Semester
CEC334 Analog IC design
UNIT II High Frequency and noise characteristics of Operational Amplifier

1. Noise in single stage amplifier

The noise performance of single-stage amplifiers at low frequencies. Before considering specific topologies, we describe a lemma that simplifies noise calculations.

Lemma: The circuits shown in Fig. 7.39(a) and (b) are equivalent at low frequencies if

$$\overline{V_n^2} = \overline{I_n^2}/g_m^2$$

and the circuits are driven by a finite impedance.

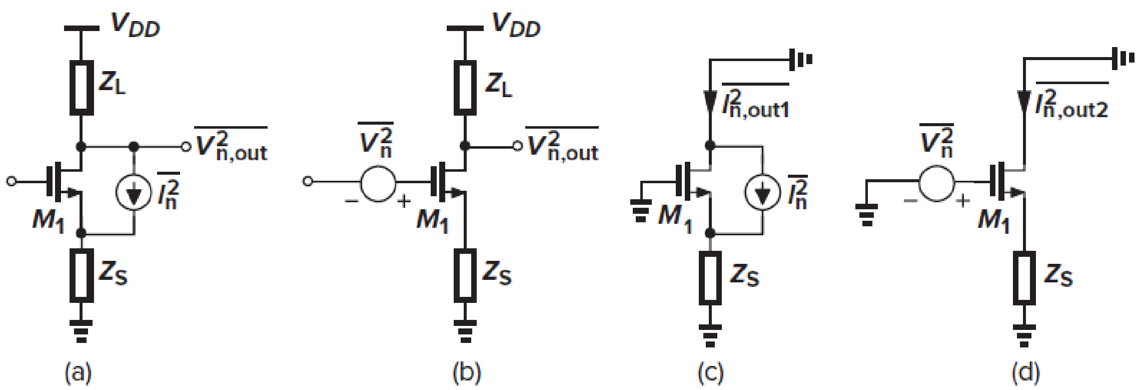

Figure 7.39 Equivalent CS stages.

Proof Since the circuits have equal output impedances, we simply examine the output short-circuit currents [Figs. 7.39(c) and (d)]. It can be proved (Problem 7.4) that the output noise current of the circuit in Fig. 7.39(c) is given by

$$I_{n,out1} = \frac{I_n}{Z_S(g_m + g_{mb} + 1/r_O) + 1} \tag{7.73}$$

and that of Fig. 7.39(d) is

$$I_{n,out2} = \frac{g_m V_n}{Z_S(g_m + g_{mb} + 1/r_O) + 1} \tag{7.74}$$

Equating (7.73) and (7.74), we have $V_n = I_n/g_m$. We call V_n the "gate-referred" noise of M_1.

This lemma suggests that the noise source can be transformed from a drain-source current to a gate series voltage for arbitrary Z_S.

2. Noise in differential amplifier

noise in basic amplifiers, we can now study the noise behavior of differential pairs. Shown in Fig. 7.55(a), a differential pair can be viewed as a two-port circuit. It is therefore possible to model the overall noise as depicted in Fig. 7.55(b). For low-frequency operation, $I^2_{n,in}$ is negligible.

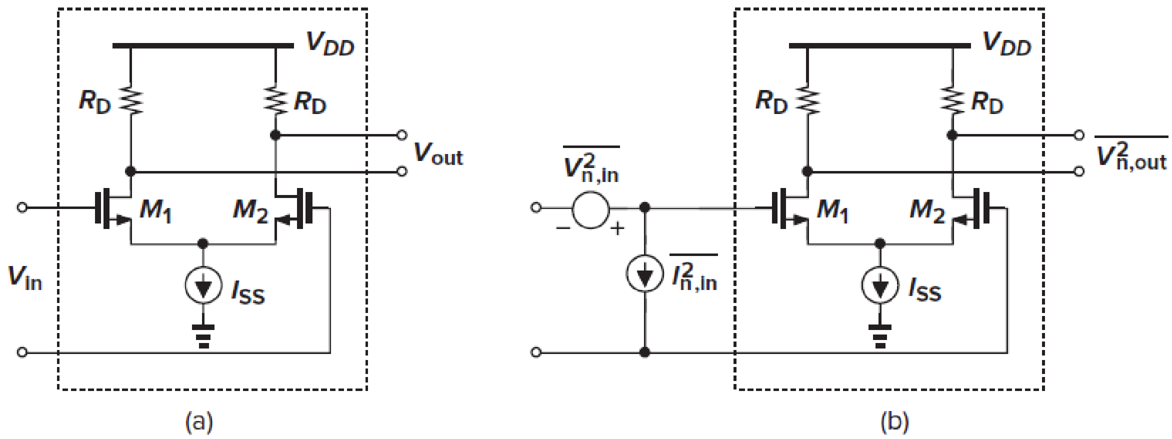

Figure 7.55 (a) Differential pair; (b) circuit including input-referred noise sources.

To calculate the thermal component of $\overline{V^2_{n,in}}$, we first obtain the total output noise with the inputs shorted together [Fig. 7.56(a)], noting that superposition of power quantities is possible because the noise

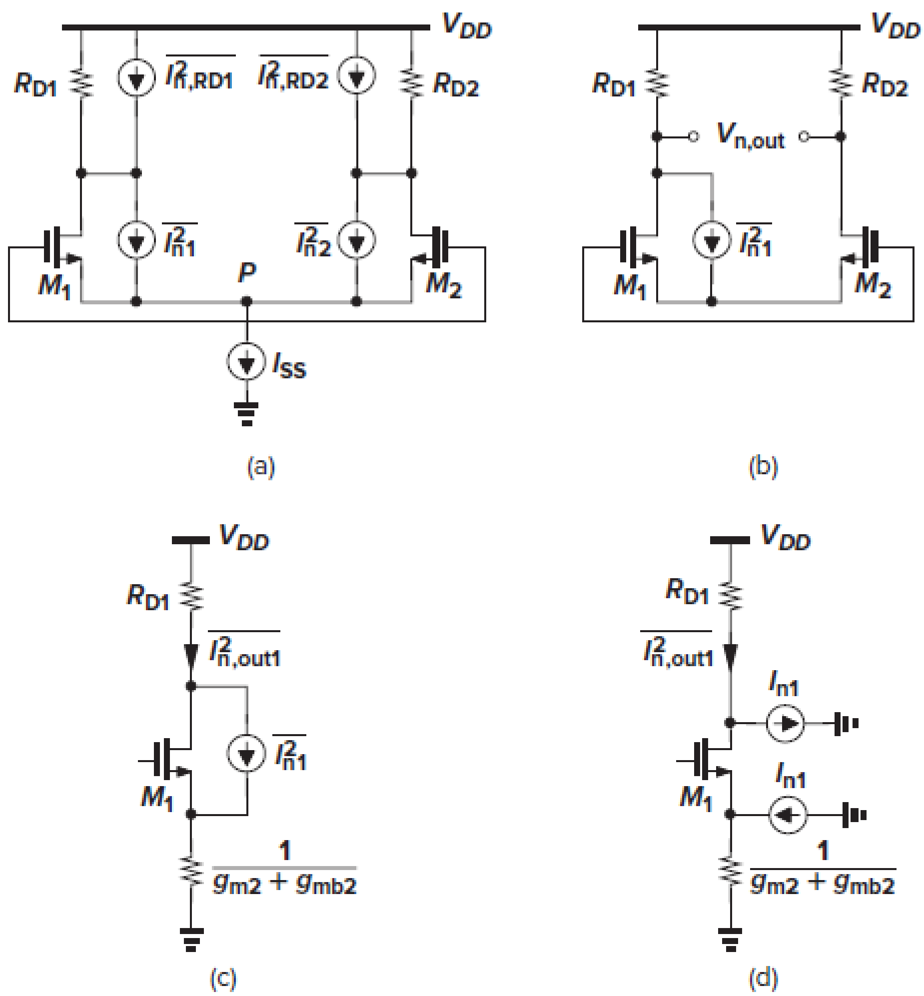

Figure 7.56 Calculation of input-referred noise of a differential pair.

sources in the circuit are uncorrelated. Since I_{n1} and I_{n2} are uncorrelated, node P cannot be considered a virtual ground, making it difficult to use the half-circuit concept. Thus, we simply derive the effect of each source individually. Depicted in Fig. 7.56(b), the contribution of I_{n1} is obtained by first reducing the circuit to that in Fig. 7.56(c). With the aid of this figure and neglecting channel-length modulation, the reader can prove that half of I_{n1} flows through R_{D1} and the other half through M_2 and R_{D2}. [As shown in Fig. 7.56(d), this can also be proved by decomposing I_{n1} into two (correlated) current sources and calculating their effect at the output.] Thus, the differential output noise due to M_1 is equal to

$$V_{n,out}|_{M1} = \frac{I_{n1}}{2}R_{D1} + \frac{I_{n1}}{2}R_{D2} \tag{7.119}$$

Note that the two noise voltages are directly added because they both arise from I_{n1} and are therefore correlated. It follows that, if $R_{D1} = R_{D2} = R_D$,

$$\overline{V_{n,out}^2}\Big|_{M1} = \overline{I_{n1}^2}R_D^2 \tag{7.120}$$

Similarly,

$$\overline{V_{n,out}^2}\Big|_{M2} = \overline{I_{n2}^2}R_D^2 \tag{7.121}$$

yielding

$$\overline{V_{n,out}^2}\Big|_{M1,M2} = (\overline{I_{n1}^2} + \overline{I_{n2}^2})R_D^2 \tag{7.122}$$

Taking into account the noise of R_{D1} and R_{D2}, we have for the total output noise

$$\overline{V_{n,out}^2} = \left(\overline{I_{n1}^2} + \overline{I_{n2}^2}\right)R_D^2 + 2(4kTR_D) \tag{7.123}$$

$$= 8kT\left(\gamma g_m R_D^2 + R_D\right) \tag{7.124}$$

Dividing the result by the square of the differential gain, $g_m^2 R_D^2$, we obtain

$$\overline{V_{n,in}^2} = 8kT\left(\frac{\gamma}{g_m} + \frac{1}{g_m^2 R_D}\right) \tag{7.125}$$

This is simply twice the input noise voltage squared of a common-source stage.
The input-referred noise voltage can also be calculated by exploiting the lemma illustrated in Fig. 7.39. As shown in Fig. 7.57, the noise of M_1 and M_2 is modeled as a voltage source in series with their gates, and the noise of R_{D1} and R_{D2} is divided by $g_{2m}R_{2D}$, thereby resulting in (7.125). The reader is encouraged to repeat these calculations if the tail current source is replaced with a short circuit.

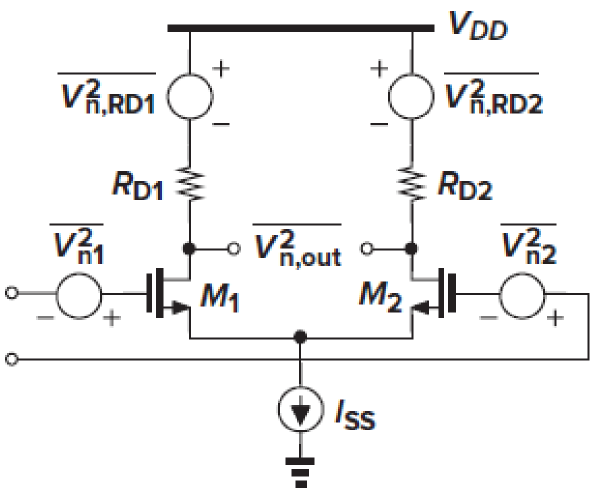

Figure 7.57 Alternative method of calculating the input-referred noise.

The noise modeling of Fig. 7.57 can readily account for $1/f$ noise of the transistors as well. Placing the voltage sources given by $K/(C_{ox}WL)$ in series with each gate, we can rewrite (7.125) as

$$\overline{V_{n,in,tot}^2} = 8kT\left(\frac{\gamma}{g_m} + \frac{1}{g_m^2 R_D}\right) + \frac{2K}{C_{ox}WL}\frac{1}{f} \tag{7.126}$$

These derivations suggest that the input-referred noise voltage squared of a fully-differential circuit is equal to twice that of its half-circuit equivalent (because the latter employs half as many devices in the signal path). The following example reinforces this point.