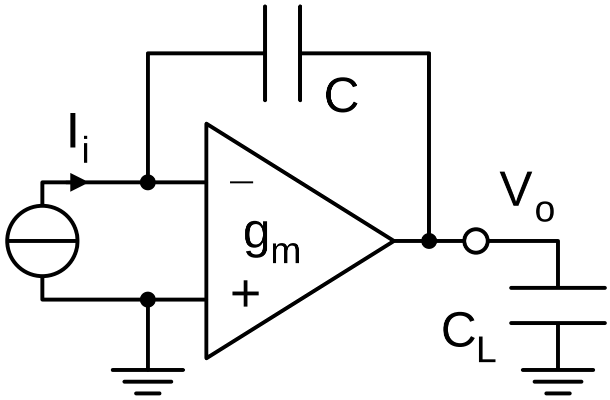
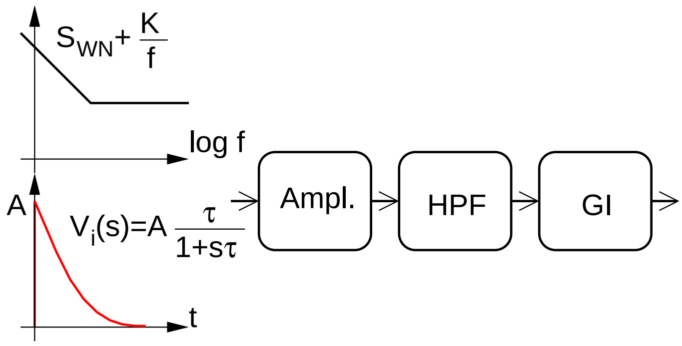
Electronics – March 3 2016   Prof. A. Spinelli





Problem 1

The OA in the left figure is a *transconductance* amplifier, providing an output <u>*current*</u> proportional to the differential input voltage, $I_o = g_m(V^+ - V^-)$, with $g_m = 10^{-2}$ S. The amplifier has an output resistance $R_o = 10$ kΩ.

1. Compute the <u>ideal</u> closed-loop gain of the stage (V_o/I_i) and find the capacitor values granting unit gain magnitude at 10 MHz.
2. Compute the loop gain of the stage and find the capacitor values granting a zero-dB frequency of 100 MHz.
3. Compute the total output noise PSD due to the input noise sources of the amplifier, $\sqrt{S_V} = 5$ nV/$\sqrt{\text{Hz}}$ and $\sqrt{S_I} = 10$ pA/$\sqrt{\text{Hz}}$.
4. Modify the circuit to achieve an integrator for a *voltage* source input and <u>discuss the proposed solution</u>. Neglect C_L for simplicity.

Problem 2

An exponential signal (right figure) having $A \approx 100$ μV and $\tau = 0.5$ μs sits on top of a background noise having <u>unilateral</u> $S_V = S_{WN} + K/f = (10^{-15} + 3\times10^{-10}/f)$ V^2/Hz. The signal is fed into a preamplifier having single-pole bandwidth of 10 MHz.

1. Design an HPF that allows to neglect the effect of the flicker noise component (hint: set $\overline{V^2_{o,WN}} = 10\overline{V^2_{o,FN}}$).
2. Compute the HPF output signal as a function of time.
3. Design a GI following the HPF, to enhance S/N.
4. A baseline restorer is used immediately after the amplifier, replacing HPF and GI. Discuss the effectiveness of the solution.

Question

Describe the noise parameters of Operational Amplifiers and their dependence on technology.

For a correct evaluation you are asked to write your answers in a readable way; thank you Do a good job!

Results will be posted by March 7th   **Mark registration: Friday, March 11th**

Solution

Problem 1

1.1

In the ideal case, the gain of the operational transconductance amplifier (OTA in the following) is infinite, meaning that its inverting input becomes a virtaul ground, and the input current I_i flows into the feedback capacitor, setting

$$V_o = -\frac{I_i}{sC}.$$

The stage is then an inverting integrator. To have gain magnitude of 0 dB at $f_0 = 10$ MHz we must have

$$\frac{1}{2\pi f_0 C} = 1\ \mathrm{S} \Rightarrow C = \frac{1}{2\pi f_0} = 16\ \mathrm{nF},$$

while the value of C_L is irrelevant. We note in passing that in this circuit the *direct gain* in non-zero. In fact, if we remove the OTA output current source, we still get a non-zero transfer, as I_i flows directly through C and C_L giving

$$G_{dir} = Z_{C_L} \parallel R_o = \frac{R_o}{1 + sC_L R_o},$$

so that the true expression for the closed-loop gain is

$$G = \frac{G_{id}}{1 - 1/G_{loop}} + \frac{G_{dir}}{1 - G_{loop}} = \frac{G_{OL} + G_{dir}}{1 - G_{loop}}.$$

The additional term becomes relevant at high frequencies, but we will neglect it in the following for simplicity.

1.2

To compute the loop gain we need to open the loop and insert a test signal, having shut off the input current source. We note that the OTA output is a *current source*, so that the impedance must be carefully reconstructed if a voltage source is to used as a test signal, as we usually do. This is by no means difficult, but is left as a control exercise: we instead avoid this by breaking the loop at the amplifier input (Fig. 1, left) and observe that no current is flowing into the capacitor C, yielding

$$G_{loop} = -g_m (R_o \parallel Z_L) = -\frac{g_m R_o}{1 + sC_L R_o},$$

which is a transfer function with a single pole, causing no stability problem and crossing the zero-dB axis at a frequency $f_c = g_m/(2\pi C_L)$. The condition on $GBWP$ becomes then

$$C_L = \frac{g_m}{2\pi f_c} = 16\ \mathrm{pF},$$

meaning that the circuit is able to drive only very small capacitive loads over the required bandwidth. Note that now is the value of the capacitor C that does not enter the calculation.

1.3

The input noise curent is transferred as the signal, while the input voltage noise experiences a unity-gain transfer. This translates into an ouput noise PSD

$$S_{V_o} = S_v + S_I \frac{1}{(2\pi f C)^2}$$

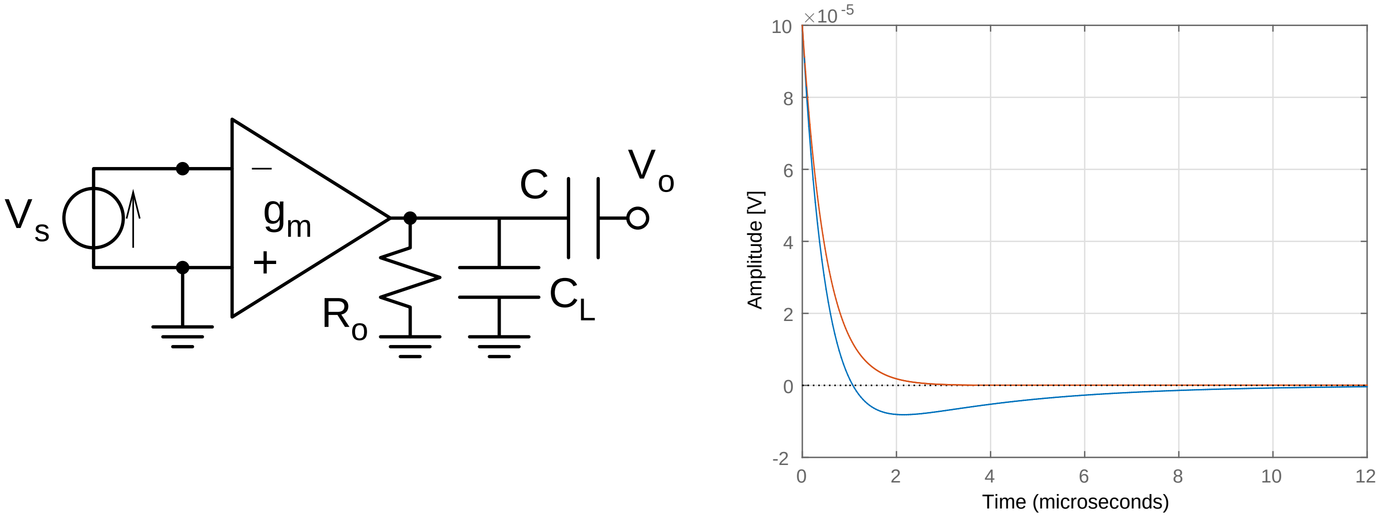


Figure 1: Left: Schematic for the loop gain calculation. Right: time dependence of input (red) and output (blue) voltage to the HPF filter.

1.4

We use the standard configuration for an integrator, placing a resistor R in series to the inverting input. The ideal transfer is obviously $-1/sCR$, but we need to pay attention to the loop gain. Neglecting C_L for simplicity, it becomes

$$G_{loop} = -g_m \frac{R_o}{Z_C + R + R_o} R = -\frac{sCg_m RR_o}{1 + sC(R + R_o)}.$$

This, again, shows no stability problem, but if we look at the flat part beyond the pole, we see a loop gain value given by $-g_m(R \parallel R_o)$, meaning that small values of R will degrade the loop gain, whose maximum value is $-g_m R_o = -100$, achieved for $R \gg R_o$.

Problem 2

2.1

The output noise due to the white component is simply

$$\overline{V^2_{o,WN}} = S_{WN}\frac{\pi}{2}(BW - f_{HP}) \approx S_{WN}\frac{\pi}{2}BW,$$

where we assumed $BW \gg f_{HP}$. The flicker noise contribution after the introduction of the HPF is

$$\overline{V^2_{o,FN}} = K \ln\left(\frac{BW}{f_{HP}}\right),$$

so that our requirement means:

$$\overline{V^2_{o,FN}} = \frac{\overline{V^2_{o,WN}}}{10} \Rightarrow f_{HP} = BWe^{-\overline{V^2_{o,WN}}/10K} \approx 53 \text{ kHz},$$

meaning that the HPF time constant is $\tau_{HP} \approx 3$ μs. Note that f_{HP} is close to the noise corner frequency.

2.2

By noting that the Laplace transform of the signal $V_i(t)$ is simply

$$V_i(s) = A\frac{\tau}{1 + s\tau},$$

we see immediately that its bandwidth is $1/2\pi\tau \approx 320$ kHz, meaning that it is not affected by the amplifier. To compute the effect of the HPF, we stick to the frequency domain: the output signal becomes

$$V_o(s) = V_i(s)\frac{s\tau_{HP}}{1+s\tau_{HP}} = A\frac{\tau}{1+s\tau}\frac{s\tau_{HP}}{1+s\tau_{HP}},$$

which we need to antitransform. This is easily done by decomposition:

$$V_o(s) = A\frac{\tau\tau_{HP}}{\tau_{HP}-\tau}\left(\frac{1}{1+s\tau} - \frac{1}{1+s\tau_{HP}}\right) \Rightarrow V_o(t) = A\left(\frac{\tau_{HP}}{\tau_{HP}-\tau}e^{-t/\tau} - \frac{\tau}{\tau_{HP}-\tau}e^{-t/\tau_{HP}}\right)u(t),$$

$u(t)$ being the usual unit step function. The output (and input) voltage are shown in Fig. 1 (right).

Of course, it is perfectly fine to work out the qualitative behavior just by inspecting the circuit: HPF has a zero in the origin, i.e., no DC transfer, i.e., zero mean value of $V_o(t)$. The input exponential has then to be followed by a negative tail dictated by the slower time constant.
Yet another different approach would be to recognize that real poles in the transfer function translate into exponential signals, so that we could write:

$$V_o(t) = (C_1 e^{-t/\tau} + C_2 e^{-t/\tau_{HP}})u(t),$$

where the constants can be found requiring $V_o(0) = A$ and $\int V_o(t)dt = C_1\tau + C_2\tau_{HP} = 0$.

2.3

The input signal to the GI is what we have just computed. At the GI input we then have

$$\frac{S}{N} = \frac{A}{\sqrt{\overline{V_{o,WN}^2}}} = 0.8,$$

not too bad, but let's see if we can do better. We set the integration time equal to the zero-crossing time of $V_o(t)$, $t_0 = \frac{\tau\tau_{HP}}{\tau_{HP}-\tau}\log\left(\frac{\tau_{HP}}{\tau}\right) \approx 1$ μs, obtaining

$$\frac{S}{N} = \frac{\int_0^{t_0} V_o(t)dt}{\sqrt{S_{WN}t_0/2}} = A\sqrt{\frac{2}{S_{WN}t_0}}\frac{\tau\tau_{HP}}{\tau_{HP}-\tau}\left(e^{-t_0/\tau_{HP}} - e^{-t_0/\tau}\right) = 1.5.$$

A slightly better result can be achieved approximating the early stage of the exponential decay by a linear decay (see Fig. 1, right) and recalling that the best integration time for this case is equal to 2/3 of the pulse time; integrating up to $2t_0/3$ we get $S/N = 1.7$.
To be more precise, however, we should point out that the GI has largely reduced the bandwidth, hence the white noise, so that now the flicker noise contribution could be no longer negligible. As a matter of fact, the normalized (DC gain $= 1/t_0$) output noises are $\sqrt{S_{WN}/2t_0} \approx 22$ μV and $\sqrt{K\log(\tau_{HP}/2t_0)} \approx 11$ μV, leading to a total noise of about 24 μV rms; actual values of S/N are somewhat degraded.

2.4

The BLR is effective in cutting correlated noise, but cannot help in reducing the white noise, meaning that the best S/N that we can achieve is 0.8, the value computed in #2.3. The proposed arrangement is then not a particularly effective solution.