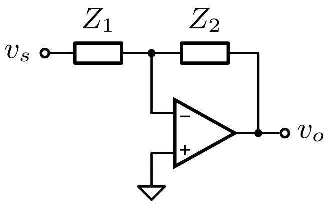
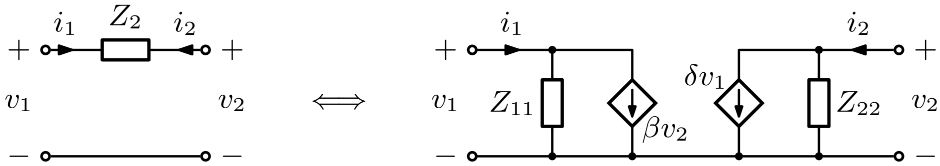
Inverting configuration. The inverting configuration of an operational amplifier employs *voltage sensing* (shunt) at the output and *current mixing* (shunt) at the summing node of the equivalent schematic, even if the ideal operational amplifier sinks no current from its input terminals. This amplifier circuit is therefore a *transresistance* amplifier, from feedback theory's point of view.



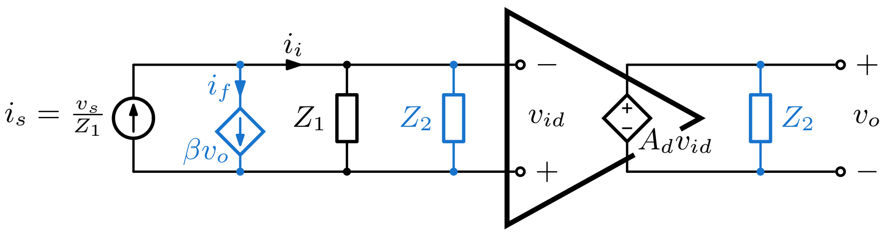
As a consequence, the β-network can be represented as a shunt-shunt (Norton-Norton) equivalent two-port network, where generators are controlled by the quantity complementary to the generator type at the other port. In this case, both ports generate a current; therefore, both controlled generators are driven by the voltage at the other port:



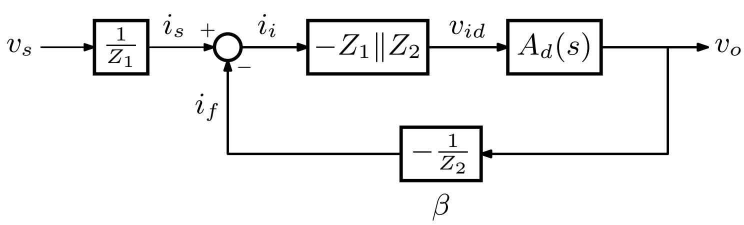
where $\beta, \delta, Z_{11}, Z_{22}$ are determined from the equations of the two-port network:

$$\begin{cases} i_1 = \dfrac{v_1}{Z_{11}} + \beta v_2 \\ i_2 = \dfrac{v_2}{Z_{22}} + \delta v_1 \end{cases} \Longrightarrow \begin{cases} Z_{11} = \left.\dfrac{v_1}{i_1}\right|_{v_2=0} = Z_2 \\ \beta = \left.\dfrac{i_1}{v_2}\right|_{v_1=0} = -\dfrac{1}{Z_2} \\ Z_{22} = \left.\dfrac{v_2}{i_2}\right|_{v_1=0} = Z_2 \\ \delta = \left.\dfrac{i_2}{v_1}\right|_{v_2=0} = -\dfrac{1}{Z_2}. \end{cases}$$

Neglecting the effect of the controlled generator at port 2 for the unidirectionality hypothesis (i. e. $\delta \approx 0$), the equivalent schematic is the following, where the β-network is highlighted in blue:



which is associated to the following block schematic:



The open-loop gain $A_{OL}(s)$ and the loop gain $T(s)$ of the amplifier are therefore

$$A_{OL}(s) \triangleq \left.\frac{v_o}{i_s}\right|_{\beta=0} = -(Z_1 \parallel Z_2)A_d(s)$$

and

$$T(s) \triangleq \beta A_{OL}(s) = \frac{Z_1 \parallel Z_2}{Z_2} A_d(s) = \frac{Z_1}{Z_1 + Z_2} A_d(s).$$

The loop gain, being a property of the circuit and not of the particular feedback representation, is identical to the one calculated with loop inspection.

The feedback amplifier voltage gain $A_F(s)$ is then

$$A_F(s) \triangleq \frac{A_{OL}}{1 + \beta A_{OL}} = -\frac{(Z_1 \parallel Z_2)A_d(s)}{1 + \frac{Z_1 \parallel Z_2}{Z_2} A_d(s)} = -\frac{(Z_1 \parallel Z_2)A_d(s)}{1 + \frac{Z_1}{Z_1+Z_2} A_d(s)}$$

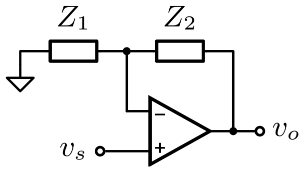
which, again, is a resistance gain; to convert it to a voltage gain we have to substitute back the expression of $i_s = \frac{v_s}{Z_1}$ of the equivalent Norton source generator:

$$A_v(s) \triangleq \frac{v_o}{v_s} = \frac{v_o}{i_s} \cdot \frac{i_s}{v_s} = A_F(s) \cdot \frac{1}{Z_1} = -\frac{Z_2}{Z_1 + Z_2} \cdot \frac{A_d(s)}{1 + \frac{Z_1}{Z_1+Z_2} A_d(s)}.$$

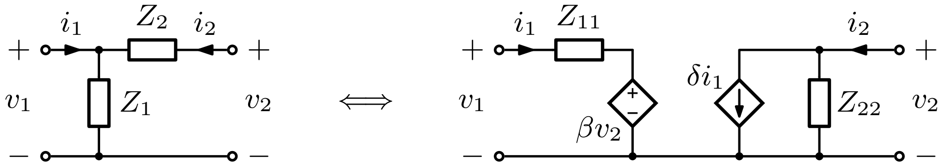
When the loop gain is very large, the 1 in the denominator is negligible and the voltage gain takes the «ideal» form:

$$A_v(s) \xrightarrow{\beta A_d(s) \gg 1} -\frac{Z_2}{Z_1 + Z_2} \cdot \frac{A_d(s)}{\cancel{1+} \frac{Z_1}{Z_1+Z_2} A_d(s)} = -\frac{Z_2}{Z_1}.$$ ■

Non-inverting configuration. The inverting configuration of an operational amplifier employs *voltage sensing* (shunt) at the output and *voltage mixing* (series) at the input mesh. This amplifier circuit is therefore a *voltage* amplifier, from feedback theory's point of view.



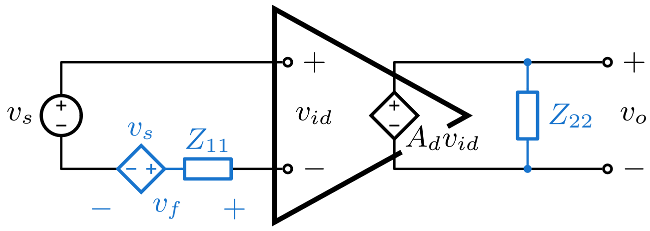
As a consequence, the β-network can be represented as a series-shunt (Thévenin-Norton) equivalent two-port network, where generators are controlled by the quantity complementary to the fed one at the other port. In this case, port 1 is fed a voltage while port 2 is fed a current; therefore, the left generator is driven by v_2 and the right generator is driven by i_1:



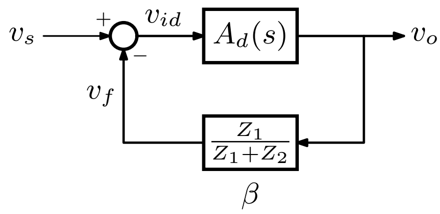
where $\beta, \delta, Z_{11}, Z_{22}$ are determined from the equations of the two-port network:

$$\begin{cases} v_1 = Z_{11} i_1 + \beta v_2 \\ i_2 = \dfrac{v_2}{Z_{22}} + \delta i_1 \end{cases} \Longrightarrow \begin{cases} Z_{11} = \left.\dfrac{v_1}{i_1}\right|_{v_2=0} = Z_1 \parallel Z_2 \\ \beta = \left.\dfrac{v_1}{v_2}\right|_{i_1=0} = \dfrac{Z_1}{Z_1 + Z_2} \\ Z_{22} = \left.\dfrac{v_2}{i_2}\right|_{i_1=0} = Z_1 + Z_2 \\ \delta = \left.\dfrac{i_2}{i_1}\right|_{v_2=0} = -\dfrac{Z_1}{Z_1 + Z_2}. \end{cases}$$

Neglecting the effect of the δv_{id} controlled generator for the unidirectionality hypothesis, the equivalent schematic is the following, where the β-network is highlighted in blue:



which is associated to the following block schematic, where each interconnection corresponds to a physical quantity in the schematic. The effect of Z_{11} is not relevant, because the ideal operational amplifier has an infinite input resistance, so the current through Z_{11} is zero.



The resistive open-loop gain is $A_{OL}(s) = A_d(s)$, while the loop gain $T(s)$ of the amplifier is

$$T(s) \triangleq \beta A_{OL}(s) = \frac{Z_1}{Z_1 + Z_2} A_d(s) = \frac{Z_1}{Z_1 + Z_2} A_d(s).$$

The loop gain, being a property of the circuit and not of the particular feedback representation, is identical to the one calculated with loop inspection. Moreover, the loop gain is the same for both the inverting and non-inverting configurations, because when switching off v_s the circuits are identical.

The feedback amplifier voltage gain $A_F(s)$ is then

$$A_F(s) \triangleq \frac{A_{OL}}{1 + \beta A_{OL}} = \frac{A_d(s)}{1 + \frac{Z_1}{Z_1+Z_2} A_d(s)}$$

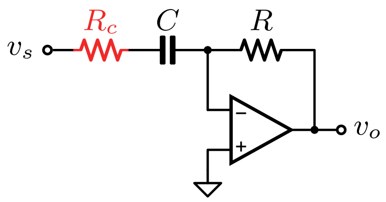
which, in this case, coincides with the required voltage gain $A_v(s)$.

When the loop gain is very large, the 1 in the denominator is negligible and the voltage gain takes the «ideal» form:

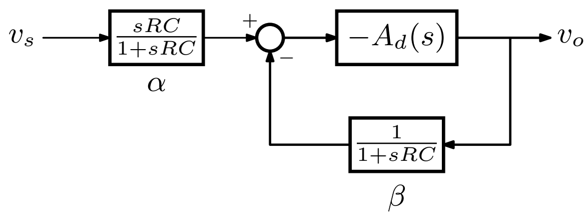
$$A_v(s) \xrightarrow{\beta A_d(s) \gg 1} \frac{A_d(s)}{\cancel{1+} \frac{Z_1}{Z_1+Z_2} A_d(s)} = \frac{Z_1 + Z_2}{Z_1} = 1 + \frac{Z_2}{Z_1}. \quad \blacksquare$$

Derivator compensation. The derivator can produce an unbounded output, in the sense that a sufficiently high-frequency signal generates an arbitrarily large output signal. For example, an ideal step input produces an impulsive output if the operational amplifier has an infinite gain.

Instability is highlighted by the amplifier's loop gain, which may cross 0 dB with a -40 dB/dec slope; therefore, in the worst case, the phase margin is close to $m_\phi \approx 0°$. A zero must be added to the β-network to boost phase margin ($\Rightarrow$ limit bandwidth).



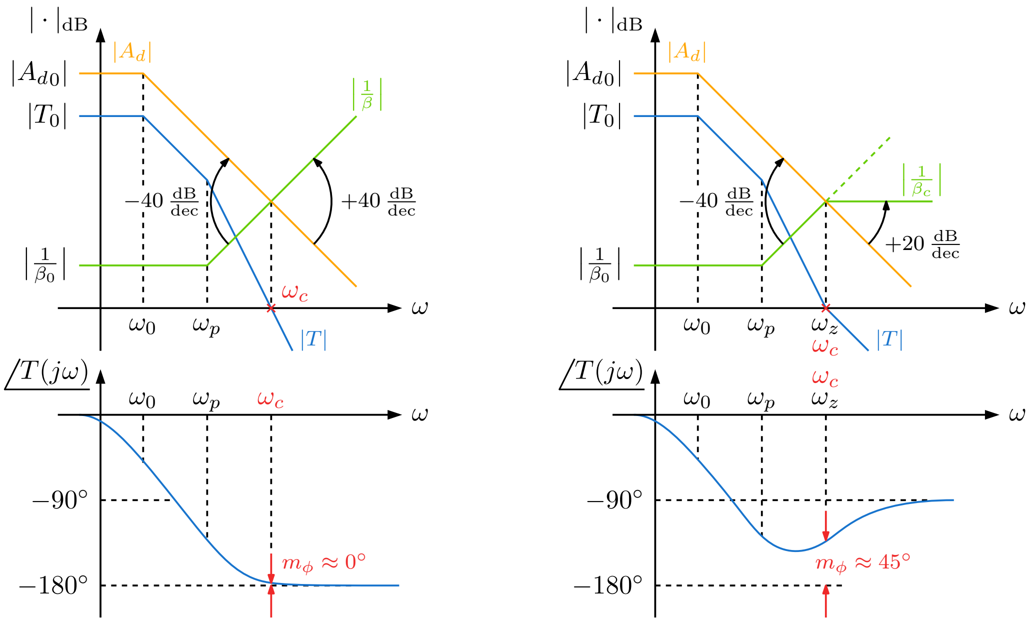
Let's analyse the derivator without compensation. By loop inspection we obtain the following block diagram, which is *not coherent* with feedback theory and so interconnections do not represent physical signals in the circuit. However, the stability analysis is fully equivalent.



To design the compensator network, we can use a graphical method based on *closing* and *opening* ratios, which are the differences of slope between $A_d(j\omega)$ and $\frac{1}{\beta(j\omega)}$ around the crossing $\omega = \omega_c$. Now, letting

$$\omega_p = \frac{1}{RC} \qquad \text{and} \qquad A_d(s) = \frac{A_{d0}}{1 + \frac{s}{\omega_0}},$$

the resulting diagrams are the following:



On the right diagram, compensation is obtained by placing the zero of the β-network exactly in ω_c: this way, the opening ratio becomes $+20$ dB and phase margin raises to around 45°.

If other values of m_ϕ are requested, the graphical method is less practical to use. ■

Example 1. Consider the high-pass inverting configuration (derivator), with $R = 100\,\mathrm{k\Omega}$, $C = 1\,\mathrm{nF}$ and $R_c = 1\,\mathrm{k\Omega}$. The operational amplifier has a first order open-loop transfer function

$$A_d(s) = \frac{A_{d0}}{1 + \frac{s}{\omega_0}}.$$

where $A_{d0} = 10^5$ and $\omega_0 = 10^2\,\mathrm{rad/s}$.

1. Ideal voltage gain and phase margin.

   The ideal voltage gain $A_v^{\mathrm{id}}(s)$ is obtained for an infinite loop gain. Being an inverting amplifier,

   $$A_v^{\mathrm{id}}(s) = -\frac{R}{R_c + \frac{1}{sC}} = -\frac{sRC}{1 + sR_cC}$$

   From loop inspection, the β-network has transfer function

   $$\beta(s) = \frac{R_c + \frac{1}{sC}}{R + R_c + \frac{1}{sC}} = \frac{1 + sR_cC}{1 + s(R + R_c)C}$$

   and therefore the loop gain is

   $$T(s) = \beta(s)A_d(s) = 10^5 \frac{1 + \frac{s}{\omega_z}}{\left(1 + \frac{s}{\omega_0}\right)\left(1 + \frac{s}{\omega_p}\right)}$$

   where $\omega_0 = 10^2\,\mathrm{rad/s}$, $\omega_z = \frac{1}{R_cC} = 10^6\,\mathrm{rad/s}$ and $\omega_p = \frac{1}{(R+R_c)C} \simeq 9901\,\mathrm{rad/s}$.

   The crossing frequency is found hypothesizing that $T(j\omega)$ crosses $0\,\mathrm{dB}$ with a slope of $-40\,\mathrm{dB}$ ($\Leftrightarrow |T(j\omega_p)| > 1$); if results are incoherent this hypothesis must be revised. Reasoning on the gain-bandwidth products:

   $$\begin{cases} T_0\,\omega_0 = |T(j\omega_p)|\,\omega_p \implies |T(j\omega_p)| = T_0\frac{\omega_0}{\omega_p} = 10^5 \cdot \frac{10^2}{9901} \simeq 1010 > 1 \quad \checkmark \\ |T(j\omega_p)|\,\omega_p^2 = \cancel{|T(j\omega_c)|}\,\omega_c^2 \end{cases}$$

   $$\implies \omega_c = \omega_p\sqrt{|T(j\omega_p)|} = \sqrt{T_0 \cdot \omega_0 \cdot \omega_p} = \sqrt{10^5 \cdot 10^2 \cdot 9901}\,\mathrm{rad/s} \simeq 314.7\,\mathrm{krad/s},$$

   where $|T(j\omega_c)| = 1$ by definition and $T_0 = \beta(j0)A_{d0} = A_{d0}$, because $\beta(j0) = 1$ in this case.

   The phase margin is then

   $$m_\phi = 180° + \overbrace{\arctan\frac{\omega_c}{\omega_z}}^{17.5°} - \overbrace{\arctan\frac{\omega_c}{\omega_0}}^{89.98°} - \overbrace{\arctan\frac{\omega_c}{\omega_p}}^{88.2°} \simeq 19.32°.$$

2. Compute R_c such that $m_\phi^* \gtrsim 45°$.

   To compensate the amplifier, we can act on R_c; since $R \gg R_c$ the position of ω_p does not change too much, while ω_z can be placed at will without affecting the low frequency behaviour.

   Since the required phase margin must be increased by 45°, and poles and the zero are sufficiently split apart, we can use the graphical method, placing ω_z in ω_c:

   $$\omega_z = \omega_c = \frac{1}{R_cC} \implies R_c = \frac{1}{\omega_cC} = \frac{1}{314.7\,\mathrm{krad/s} \cdot 1\,\mathrm{nF}} \simeq 3178\,\Omega.$$

3. Determine the voltage gain at $\omega_H = 10^7\,\text{rad/s}$.

   The precise voltage gain at ω_H is

$$A_v(j\omega_H) = -\alpha(j\omega_H) \cdot \frac{A_d(j\omega_H)}{1+\beta(j\omega_H)A_d(j\omega_H)},$$

   where α is the attenuation factor at the input

$$\alpha = \frac{sRC}{1+s(R+R_c)C}.$$

   Since $\omega_H \gg \omega_0, \omega_p$, for $R_c = 1\,\text{k}\Omega$

$$\begin{aligned}
|A_v(j\omega_H)| &= \frac{\omega_H RC}{\sqrt{1+\frac{\omega_H^2}{\omega_p^2}}} \cdot \frac{\frac{A_{d0}}{\sqrt{1+\frac{\omega_H^2}{\omega_0^2}}}}{1+A_{d0}\frac{\sqrt{1+\frac{\omega_H^2}{\omega_z^2}}}{\sqrt{1+\frac{\omega_H^2}{\omega_0^2}}\sqrt{1+\frac{\omega_H^2}{\omega_p^2}}}} = \\
&\approx \frac{\omega_H RC}{\omega_H/\omega_p} \cdot \frac{\frac{A_{d0}}{\omega_H/\omega_0}}{1+A_{d0}\frac{\sqrt{1+\frac{\omega_H^2}{\omega_z^2}}}{(\omega_H/\omega_0)(\omega_H/\omega_p)}} = \\
&= RC \cdot \frac{A_{d0}}{\frac{\omega_H}{\omega_0\omega_p} + A_{d0}\sqrt{\frac{1}{\omega_H^2}+\frac{1}{\omega_z^2}}} = \\
&= 100\,\text{k}\Omega \cdot 1\,\text{nF} \cdot \frac{10^5}{\frac{10^7}{10^2 \cdot 9.9\,\text{krad/s}} + 10^5\sqrt{\frac{1}{10^{14}\,\text{rad}^2/\text{s}^2}+\frac{1}{10^{12}\,\text{rad}^2/\text{s}^2}}} = \\
&\simeq 0.98.
\end{aligned}$$

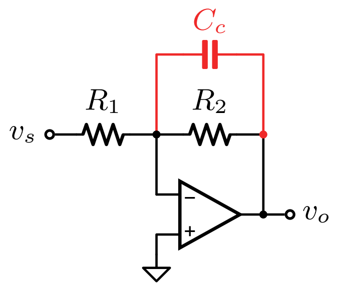
   A more approximate result can be obtained simply by considering that, at $\omega_H \gg \omega_c$, the loop gain is very small, and in particular $|T(j\omega_H)| \ll 1$; so

$$\frac{A_d}{1+\beta A_d} \xrightarrow{|\beta A_d| \ll 1} A_d.$$

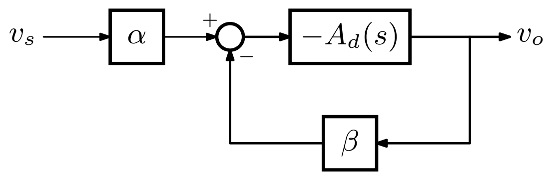
   Therefore,

$$|A_v(j\omega_H)| \approx \frac{R}{R+R_c} \cdot \frac{A_{d0}}{\sqrt{1+\frac{\omega_H^2}{\omega_0^2}}} \approx \frac{R}{R+R_c} \cdot \frac{\omega_0 A_{d0}}{\omega_H} = 0.99.$$

Dominant pole compensation (zero-pole). In the case of non-compensated operational amplifiers (i. e. which exhibit two poles relatively close to each other), a possible approach is to use the feedback network to boost phase margin.



To do calculations, it is more practical to use loop inspection:

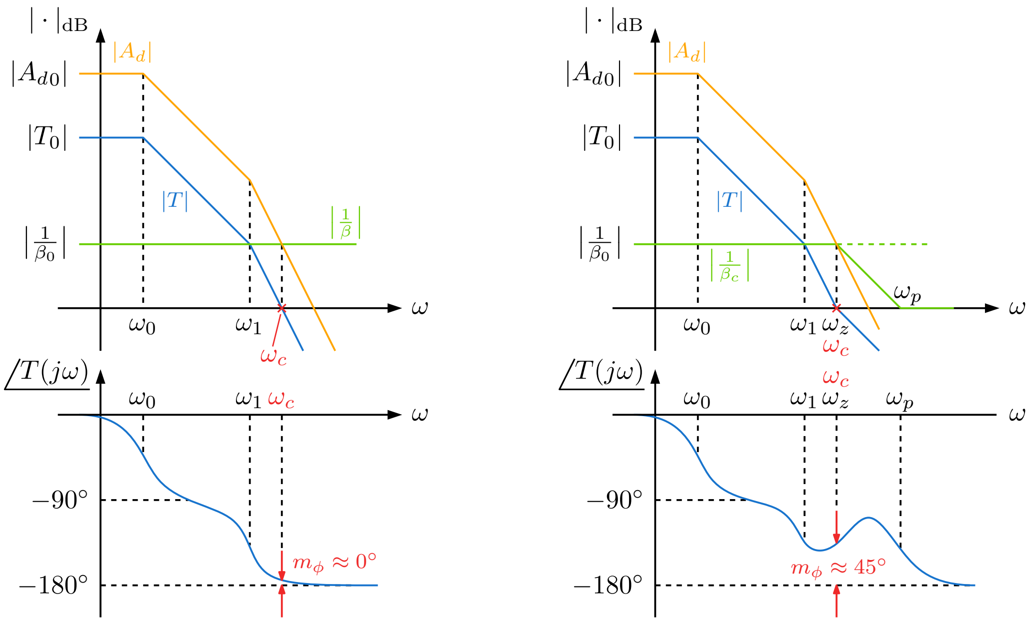


where

$$\alpha = \frac{R_2}{R_1 + R_2} \frac{1}{1 + s(R_1 \| R_2)C_c} \qquad \text{and} \qquad \beta = \frac{R_1}{R_1 + R_2} \frac{1 + sR_2C_c}{1 + s(R_1 \| R_2)C_c}.$$

It can be noticed that the β-network includes a pole together with the zero, but it is a higher frequency pole needed to bring $|\beta_\infty|$ to 1:

$$\omega_p = \frac{1}{(R_1 \| R_2)C_c} > \omega_z = \frac{1}{R_2C_c}.$$



Usually a phase margin boost of about 45° is requested. A possible solution is to place the zero of the β-network at the crossing frequency ω_c, which therefore corresponds to the new crossing frequency ω_c^*.

A less used (and less useful) approach is to place β's pole at $A_d(s)$'s crossing frequency, which has the same effect on phase margin, but modifies the loop gain also for frequencies lower than $T(s)$'s crossing frequency ω_c^*. ■

Example 2. Consider the inverting configuration with dominant pole compensation, implemented through a capacitor C_c in parallel with R_2. Let $R_1 = 1\,\mathrm{k\Omega}$, $R_2 = 33\,\mathrm{k\Omega}$ and

$$A_d(s) = \frac{A_{d0}}{\left(1+\frac{s}{\omega_0}\right)\left(1+\frac{s}{\omega_1}\right)},$$

where $A_{d0} = 10^5$, $\omega_0 = 10^2\,\mathrm{rad/s}$ and $\omega_1 = 5\cdot 10^3\,\mathrm{rad/s}$.

1. Phase margin of the uncompensated amplifier.

   From loop inspection, the β-network has transfer function

   $$\beta(s) = \beta_0 = \frac{R_1}{R_1+R_2} = \frac{1}{34}$$

   and the crossing frequency can be found with gain-bandwidth products:

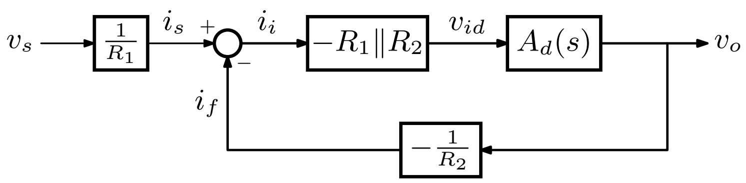
   $$\begin{cases} \beta_0 A_{d0}\omega_0 = \beta_0 |A_d(j\omega_1)|\omega_1 \implies |A_d(j\omega_1)| = A_{d0}\frac{\omega_0}{\omega_1} = 2000 > 1 \quad \checkmark \\ \beta_0 |A_d(j\omega_1)|\omega_1^2 = \omega_c^2 \implies \omega_c = \omega_1\sqrt{\beta |A_d(j\omega_1)|} = 3.84\cdot 10^4\,\mathrm{rad/s}. \end{cases}$$

   The uncompensated phase margin is then very low:

   $$m_\phi = 180^\circ - \overbrace{\arctan\frac{\omega_c}{\omega_0}}^{89.85^\circ} - \overbrace{\arctan\frac{\omega_c}{\omega_1}}^{82.58^\circ} \simeq 7.57^\circ.$$

2. Block diagram according to feedback theory.

   The inverting amplifier has a transresistance topology, so the coherent block diagram is



3. Compensation capacitance C_c to increase phase margin by 45°.

   As seen previously, the β-network has a zero ω_z and a higher frequency pole ω_p. Placing the zero exactly in ω_c gives an additional 45° phase margin:

   $$\omega_z = \frac{1}{R_2 C_c} = \omega_c \implies C_c = \frac{1}{\omega_c R_2} = \frac{1}{33\,\mathrm{k\Omega}\cdot 3.84\cdot 10^4\,\mathrm{rad/s}} \simeq 789\,\mathrm{pF}.$$

Noise gain compensation (pole-zero). The dominant pole compensation approach *fails* when the *theoretical gain* $\frac{1}{\beta}$ is *small*; in that case noise gain compensation must be used.

The β-network expression can be obtained in a very fast way by using time constant methods. First we compute the low-frequency β_0 and the high-frequency β_∞:

$$\beta_0 = \frac{R_1}{R_1 + R_2}, \qquad \beta_\infty = \frac{R_1 \| R_c}{R_1 \| R_c + R_2}$$

then, by zeroing the output of the operational amplifier v_o (which is the input port of the β-network), the resistance seen by C_c is $R_c + R_1 \| R_2$, producing a pole in

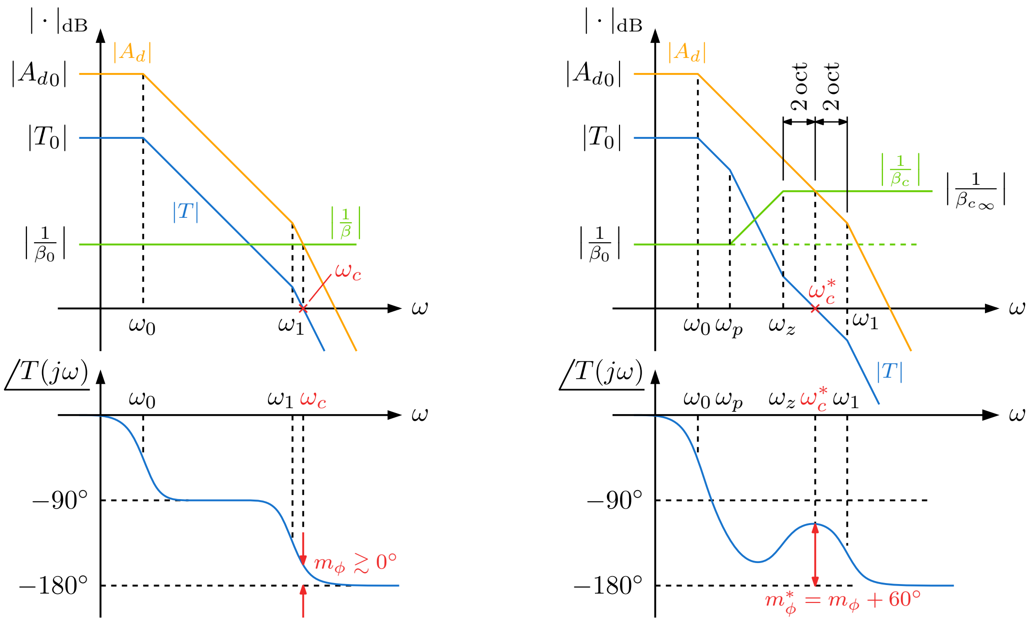
$$\omega_p = \frac{1}{(R_c + R_1 \| R_2) C_c}.$$

To find the zero, instead, the output of the β-network (i. e. the voltage v_- of the inverting input) must be zeroed; this way C_c sees only the resistance R_c and the zero is

$$\omega_z = \frac{1}{R_c C_c}.$$

Therefore,

$$\beta(s) = \frac{R_1}{R_1 + R_2} \cdot \frac{1 + s R_c C_c}{1 + s (R_c + R_1 \| R_2) C_c}.$$



For $m_\phi^* \gtrsim 60°$, the zero of the β-network can be placed two *octaves* before the new ω_c^* (i. e. $\omega_c^*/4$), symmetrically to the high frequency pole ω_1 of $A(s)$, which is two *octaves* after ω_c^* (i. e. $4\omega_c^*$). As a consequence, $\omega_z = \omega_1/16$. ■

Example 3. Consider the inverting configuration of the operational amplifier, with $R_1 = 5\,\mathrm{k\Omega}$, $R_2 = 10\,\mathrm{k\Omega}$ and

$$A_d(s) = \frac{A_{d0}}{\left(1+\frac{s}{\omega_0}\right)\left(1+\frac{s}{\omega_1}\right)},$$

where $A_{d0} = 10^5$, $\omega_0 = 10\,\mathrm{rad/s}$ and $\omega_1 = 10^5\,\mathrm{rad/s}$.

1. Phase margin of the uncompensated amplifier.

   The β factor is constant and equal to $\frac{R_1}{R_1+R_2} = \frac{1}{3}$. To find the crossing frequency:

$$\begin{cases} A_{d0}\omega_0 = |A_d(j\omega_1)|\omega_1 \implies |A_d(j\omega_1)| = A_{d0}\frac{\omega_0}{\omega_1} = 10 > 1 \quad \checkmark \\ \beta_0|A_d(j\omega_1)|\omega_1^2 = \omega_c^2 \implies \omega_c = \omega_1\sqrt{\beta_0|A_d(j\omega_1)|} \simeq 1.83 \cdot 10^5\,\mathrm{rad/s}. \end{cases}$$

   Therefore, the phase margin is

$$m_\phi = 180° - \overbrace{\arctan\frac{\omega_c}{\omega_0}}^{90.00°} - \overbrace{\arctan\frac{\omega_c}{\omega_1}}^{61.35°} \simeq 28.65°.$$

2. Design of the compensation network to have 60° additional phase margin, using noise gain compensation.

   Since we want ω_c^* to be placed two octaves before ω_1, i. e. $\omega_c^* = \omega_1/4 = 2.5 \cdot 10^4\,\mathrm{rad/s}$, the required β_∞ can be obtained:

$$\omega_c^* = \beta_\infty A_{d0}\omega_0 \implies \beta_\infty = \frac{\omega_c^*}{A_{d0}\omega_0} = \frac{2.5\cdot 10^4\,\mathrm{rad/s}}{10^5 \cdot 10\,\mathrm{rad/s}} \simeq 0.025,$$

   from which the R_c value is

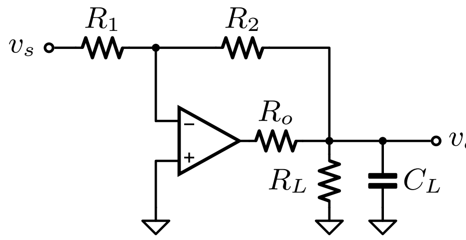
$$\beta_\infty = \frac{R_1\|R_c}{R_1\|R_c + R_2} \implies R_c = \frac{R_2}{\frac{1}{\beta_\infty} - \left(1+\frac{R_2}{R_1}\right)} = \frac{10\,\mathrm{k\Omega}}{40-(1+2)} \simeq 270\,\Omega.$$

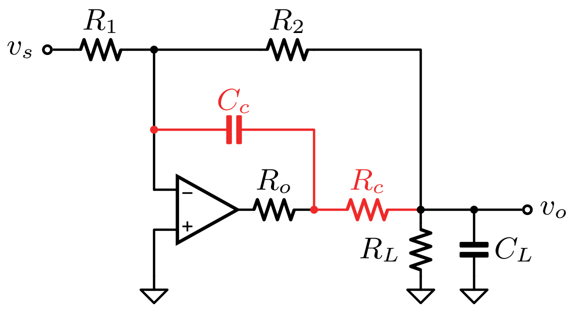
   Finally, we want to place the zero ω_z of the β-network two octaves before ω_c^* (i. e. four octaves before ω_1), therefore

$$\omega_z = \frac{\omega_c^*}{4} = \frac{1}{R_cC_c} \implies C_c = \frac{4}{\omega_c^* R_c} \simeq 593\,\mathrm{nF}.$$

Capacitively loaded amplifier compensation (pole-zero-zero-pole). The «normal» feedback loop formed by R_1 and R_2 is slowed down by the presence of the $\frac{1}{R_o C_L}$ pole at the output, possibly causing $\pm 40\,\mathrm{dB/dec}$ opening/closing ratios and thus instability.

To avoid this, a «fast» feedback loop is added by means of a decoupling resistor R_c and of a feedback capacitor C_c (*in-the-loop* compensation), which only acts at high frequency.





At low frequency ($\omega = 0$), the addition of the in-loop compensation does not change $\beta \triangleq \frac{v_-}{v_o'}$ as long as $R_c \ll R_o$:

$$\beta_{c0} = \frac{R_L}{R_o + R_c + R_L} \cdot \frac{R_1}{(R_o + R_c)\|R_L + R_2 + R_1}.$$

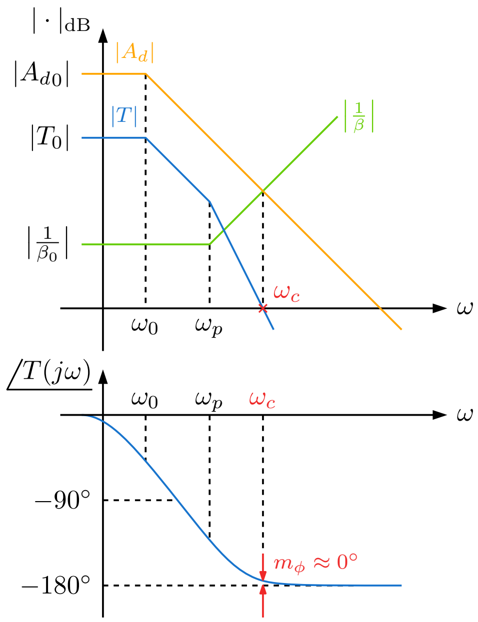
At high frequency, instead, capacitors behave as short circuits and thus

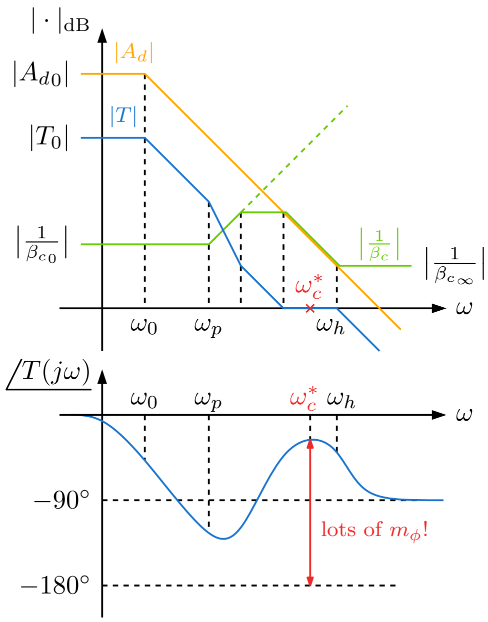
$$\beta_{c\infty} = \frac{R_c\|R_1\|R_2}{R_o + R_c\|R_1\|R_2} \overset{R_c \ll R_1, R_2}{\approx} \frac{R_c}{R_o + R_c} \overset{R_c \ll R_o}{\approx} \frac{R_c}{R_o};$$

from this relation we can calculate R_c to obtain the desired high-frequency β_c. Equivalently, we can calculate R_c for the requested high-frequency pole ω_h:

$$\omega_h = \beta_{c\infty} A_{d0} \omega_0 \approx \frac{R_c}{R_o} A_{d0} \omega_0 \implies R_c = \frac{R_o}{A_{d0}} \cdot \frac{\omega_h}{\omega_0}.$$

Overall, the β-network has two poles and two zeros; however, as long as the latter two are placed between the two poles, their position is not of vital importance. In fact, since opening and closing ratios are around $0\,\mathrm{dB/dec}$, the phase margin boost is very high:



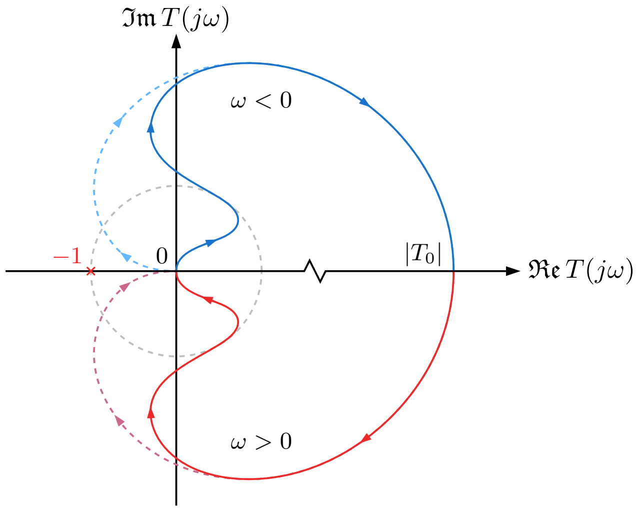


The mid frequency pole ω_p associated to C_L does not quite move, assuming $R_c \ll R_o$:

$$\omega_p = \frac{1}{[R_L \| (R_o + R_c) \| (R_1 + R_2)] C_L} \approx \frac{1}{(R_o + R_c) C_L}.$$

Then, after having calculated R_c, to find C_c we need to estimate the position of the associated high-frequency pole ω_h, by computing the resistance seen from it considering C_L a short circuit:

$$\omega_h \approx \frac{1}{(R_1 \| R_2 + R_o \| R_c) C_c}.$$



The Nyquist diagram shows how the loop gain is modified to avoid the critical point $-1 + j0$. The uncompensated frequency response is dashed, while the compensated one is solid; in red is shown the part from $\omega = 0^+$ to $+\infty$ and in blue the part from $\omega = -\infty$ to 0^-. ■

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