For \(|T| \gg 1\)
\[
A_0 = \frac{V_0}{V_S} = -\frac{R_2/R_2 + \frac{R_1}{R_2}}{R_2 + \frac{R_1}{R_2}} = -\frac{R_2}{R_1}
\]

**Hint:**
\[
\frac{R_2/R_2}{R_2 + \frac{R_1}{R_2}} = \frac{R_2/R_2 + \frac{R_1}{R_2}}{R_2 + \frac{R_1}{R_2}} \cdot \frac{1}{R_2}
\]

**Design of Compensation Network**

**Target** \( \Delta \Phi M = +60^\circ \)

This means we are designing \( \frac{1}{R_c} \) so that:

To find \( \omega_p \) we can find \( \Delta \frac{A}{A_{cm}} \).

\[
A(\omega_0) \cdot \omega_0 = \frac{1}{R_{cm}} \cdot \frac{\omega_0}{A} \quad \Rightarrow \quad \frac{1}{R_{cm}} = \frac{A(\omega_0) \cdot \omega_0}{\omega_0} \quad \text{This is known from OPAMP characteristics}
\]

From this, we find \( \omega_p \) imposing

\[
1 + \frac{1}{\omega_p \cdot \frac{1}{R_c}} = 1 + \frac{1}{\omega_p \cdot \frac{1}{R_{cm}} \cdot \frac{1}{R_0}} = 1 + \frac{1}{\omega_p \cdot \frac{1}{R_{cm}} \cdot \frac{1}{R_0}} = \frac{\omega_p}{\omega_p} \quad \Rightarrow \quad \omega_p = \frac{R_{cm} \cdot \omega_0}{16 R_c} = \frac{\omega_0}{64 A_{cm} \cdot \omega_0 \cdot R_c}
\]

For this circuit:

\[
\frac{1}{R_c}(s) = 1 + \frac{R_2}{R_2} = 1 + R_2 \cdot \left( \frac{1}{R_2} + \frac{1}{R_1} \frac{1}{sC} \right) = 1 + \frac{R_2}{R_1} + \frac{sR_cC}{1 + sR_cC}
\]

\[
= \frac{R_4 + R_2}{R_1} \cdot \frac{1 + sC}{1 + sR_cC}
\]
From which we see that:

\[
\frac{1}{\beta} = 1 + \frac{R_2}{R_1} \quad \text{as we expected it is equal to} \quad \frac{1}{\beta_0}
\]

\[\omega_c = \frac{1}{RC} \quad \text{from here we set} \quad C_C = \frac{1}{RC \omega_c}
\]

\[\omega_p = \frac{1}{C_C (R_C + \frac{R_1}{R_2})} \quad \text{from which, using this, we find}
\]

\[\omega_p = \frac{R_C \omega_c}{R_C + \frac{R_1}{R_2}} = \frac{\omega_c}{1 + \frac{R_1}{R_C}} \quad \leftrightarrow
\]

\[R_C = \frac{\frac{R_1}{R_2}}{\frac{\omega_c^2 - \omega_p^2}{\omega_c}} \quad \leftrightarrow \quad C_C = \frac{1}{R_C \omega_c}
\]

From which, finally, we can find

\[
\text{Faster procedure}
\]

From the circuit we see that:

\[
\frac{1}{\beta_0} = 1 + \frac{R_2}{R_1/R_C} = 4 \cdot \frac{\omega_0}{\omega_c} \quad \text{known}
\]

From here we can directly find \(R_C\)

Then, \(C_C\) is found as \(\frac{1}{R_C \omega_c}\)

In the non-inverting case, the noise gain compensation looks like this:

\[
\text{when } |T| >> 1 \quad A_0 = \frac{V_0}{V_S} = \frac{1 + \frac{R_2}{R_1}}{R_2}
\]

But the amplifier will be compensated.

The design procedure for \(R_C\) and \(C_C\) is the same as in the inverting configuration case.

Verify this as an exercise!
Oscillators

- Non-sinusoidal (Schmitt trigger)
- Sinusoidal

We consider the latter case. All begins with a feedback amplifier

\[
\frac{X_o}{X_i} = \frac{A}{1+T}
\]

N.B. \(\angle \frac{X_o}{X_i} = \angle T + 90^\circ\) due to the "+" sign at the summing node.

In sinusoidal oscillators, we intentionally create the conditions to have the circuit oscillate at some frequency \(\omega_0\).

These conditions are named **Barkhausen** conditions

\[
\begin{align*}
|T(j\omega)| &= 1 \\
\angle T(j\omega) &= \pm (2m+1)90^\circ \quad m \in \mathbb{N}
\end{align*}
\]

The simplest way to satisfy Barkhausen conditions is to have

\(T = -1\) \(\implies\) \(\omega = \omega_0\) \(\implies\) \(\frac{X_o}{X_i}(j\omega_0) = +1\) \(\implies\)

This happens anytime \(1+T\) includes a term as \((s^2 + \omega_0^2)\) which is \(\neq 0\) \(\implies\) \(\omega = \omega_0\).

A typical oscillator circuit is the one called the **Wien bridge** oscillator that is based on a simple opamp configuration.
Let's find $T(s)$ for this amplifier:

$$T(s) = A \cdot \frac{R}{1 + sRC} = A \cdot \frac{R}{R + R + \frac{4}{sC}} = A \cdot \frac{SCR}{SCR(2 + SCR) + 1 + SCR} = A \cdot \frac{SCR}{1 + 3SCR + s^2(SCR)^2}$$

Let's consider $T(j\omega_0)$ where $\omega_0 = \frac{1}{RC}$

$$|T(j\omega_0)| = A \cdot \left| \frac{j}{\omega_0^2 + 1} \right| = A \cdot \frac{1}{3}$$

$$\angle T(j\omega_0) = 0^\circ$$

N.B. In this case $\frac{X_R(s)}{X_T(s)} = T(s)$ because there is no summing node and no sign inversion.

We can satisfy Barkhausen conditions by making

$$A = 3 \Rightarrow 1 + \frac{R_2}{R_1} = 3 \Rightarrow R_2 = 2R_1$$

In practice we choose $R_2 > 2R_1$ to start the oscillation rapidly.
AND THEN WE PREVENT SATURATION BY ACTIVELY LIMITING THE OSCILLATION AMPLITUDE. A SIMPLE IMPLEMENTATION IS:

A nonlinear network reduces the circuit gain any time $n_0(t)$ gets close to $\pm V_f$ (the node terminated voltage drop).

This is the preferred output signal sensing point.

THANKS TO THE SELECTIVITY OF THE RESONANT $L$-NETWORK $n_0^1 = n_0$, THE FUNDAMENTAL HARMONIC COMPONENT OF $n_0$.

ANOTHER WAY IS TO MAKE $R_2$ A PTC RESISTOR. A SIMPLE EXAMPLE OF PTC RESISTOR IS A LIGHT BULB. REPLACING $R_4$ WITH A PTC AGAIN STABILIZES THE OSCILLATOR.

A SINUSOIDAL OSCILLATOR CAN BE MADE ALSO USING A SIMPLE C-E (ON C-S) AMPLIFIER.

COLPITTS OSCILLATOR
(COMMON EMITTER BASED)

HYP. @ $\omega_T = \omega_0$

1. $C_L$ AND $C_E$ ARE PRACTICALLY SHORTED
2. $\omega_0 < \omega_T$ IS THAT $C_E$ AND $C_L$ ARE PRACTICALLY OPEN.
Phase variation is very "fast". A signal with frequency different from $\omega_0$ (even by a little amount) will not meet Barkhausen conditions and will not propagate around the loop.

Selectivity = the oscillation frequency will be very stable.