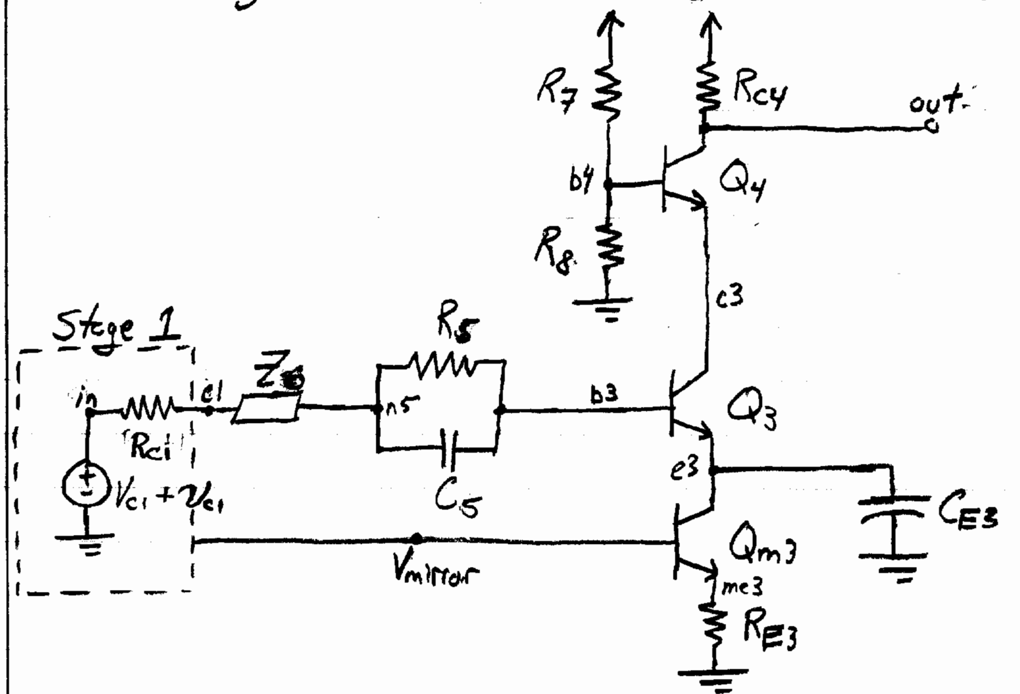
More Consistent Labeling of Components and Design of Inductance Z_6 for Noise Reduction



From prior design work (s2-BJT-d1 and s2-BJT-d2), we had the following AC characteristics:

$$G_{HIGH}^{-1} = \frac{R_{C1}}{R_{C4}\|Z_L}\left[1-\left(1-\frac{V_T}{R_{C1}I_{C3}}\right)\sqrt{1+\left(\frac{\omega}{\omega_3}\right)^2}\, e^{j2\tan^{-1}\left(\frac{\omega}{\omega_3}\right)}\right]$$

$$G_{LOW}^{-1} = \frac{-(V_{C1}-2V_{BE}-V_{margin})}{I_{C3}(R_{C4}\|Z_L)}\cdot\left[\frac{e^{-j\tan^{-1}(\omega/\omega_5)}}{\sqrt{1+(\omega/\omega_5)^2}}-j\frac{\omega_4}{\omega}\right]$$

$$\omega_4 = \frac{I_{C3}}{(V_{C1}-2V_{BE}-V_{margin})C_{E3}},\quad \omega_5=\frac{1}{R_5C_5},\quad \omega_3=\frac{I_{C3}}{V_T C_{jc}},$$

and five components were fixed by DC operating point:

$$R_{E3}=R_{E1}\frac{I_{C3}}{I_{C1}},\quad R_5=\frac{\beta(V_{C1}-2V_{BE}-V_{margin})}{I_{C3}},\quad R_{C4}=\frac{3V_{BE}+2V_{margin}}{2I_{C3}},$$

$$R_7=\frac{V_T\beta}{10I_{C3}}\left(\frac{V_{CC}}{3V_{BE}+2V_{margin}}\right),\ \text{and}\ R_8=\frac{V_T\beta}{10I_{C3}}\left(\frac{V_{CC}}{V_{CC}-3V_{BE}-2V_{margin}}\right)$$

I have no intention of repeating the network analysis, but we do need some basis for choosing L such that it does not interfere in the band pass region.

Looking at the equations above, every component and parameter (even the flat band magnitude!) is set by

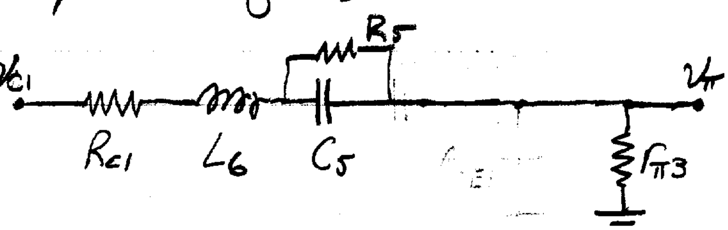
choosing 3 independent variables: ω_3, ω_4, and ω_5.
The high frequency pole ω_3, associated with the miller capacitance, does not have very broad domain because it must be greater than the max operating frequency but at the same time, increasing it adversely affects the gain,

$$G_{HIGH} \propto \frac{1}{I_{C3}} \propto \frac{1}{\omega_{3(\text{high, corner})}}$$

The addition of Z_4 will create a new high frequency pole ω_6. For sanity, we should try to keep ω_6 from interfering with ω_3, which should definitely be the lesser of the two weavils. Therefore

$$\omega_6 = 10\,\omega_3$$

The AC coupling circuit with output resistance of S1 and input impedance of Q_3 is shown below. (for high freq)



The series reactances sum to

$$X_{56} = j\omega L_6 + \frac{1}{j\omega C_5} = j\omega L_6\left(1-\frac{1}{\omega^2 C_5 L_6}\right) = \frac{j}{\omega C}\left(\omega^2 LC - 1\right)$$

$$|X_{56}| = \omega L_6\left[1-\left(\frac{\omega_6}{\omega}\right)^2\right] = \frac{1}{\omega C}\left[\left(\frac{\omega}{\omega_6}\right)^2 - 1\right] \quad \text{where } \omega_6 = \frac{1}{\sqrt{LC}}.$$

If we ignore R_{C1} and R_5 as negligible, then the transfere function is a voltage divider:

$$G_{Z6} = \frac{r_\pi}{r_\pi + \frac{j}{\omega C}\left[\left(\frac{\omega}{\omega_6}\right)^2-1\right]} = \frac{1}{1+\frac{j}{\omega r_\pi C_5}\left[\left(\frac{\omega}{\omega_6}\right)^2-1\right]}$$

[1] $$G_{Z6} = \frac{1}{1+\frac{j\omega_\pi}{\omega}\left[\left(\frac{\omega}{\omega_6}\right)^2-1\right]} \quad \text{where} \quad \omega_\pi = \frac{1}{r_\pi C_5}, \quad \omega_6 = \frac{1}{\sqrt{L_6 C_5}}$$

So the transfer ratio is dependent on two time constants, but we actually have information about these from prior equations.

$$\omega_\pi = \frac{1}{r_\pi C_5} = \frac{I_{C3}}{V_T \beta C_5} = \frac{I_{C3}}{V_T\beta}(\omega_5 R_5) = \frac{I_{C3}\omega_5}{V_T\beta}(R_5)$$

$$= \frac{I_{C3}\omega_5}{V_T\beta}\left[\frac{\beta(V_{c1} - 2V_{BE} - V_{margin})}{I_{C3}}\right] = \omega_5\left[\frac{(V_{c1} - 2V_{BE} - V_{margin})}{V_T}\right]$$

So $\omega_\pi \approx 43\omega_5$; ~~another low frequency pole~~. We are designing ω_6 to be very high, so these two do not interact.

It is actually probably the same pole, but it shifts due to L_6.

$$G_{76} = \frac{1}{1 + j\frac{43\omega_5}{\omega}\left[\left(\frac{\omega}{\omega_6}\right)^2 - 1\right]}$$

So the gain will be ≈ 1 in the bandpass region, although B is shortened by $43 \cdot f_{LP}$.

Now all that remains is to design ω_6 beyond ω_3 and to redesign ω_5 to account for the $\times 43$ shift.

$$\boxed{\omega_5 = \frac{1}{43}\omega_5^*}$$

$$C_5 = \frac{1}{\omega_5 R_5} = \frac{43}{\omega_5^* R_c} = \frac{43 I_{C3}}{\omega_5^*\beta(V_{c1} - 2V_{BE(sat)} - V_{margin})} = \frac{43\,\omega_3\, V_T\, C_{jc}}{\omega_5^*\beta(\cdots)}$$

$$\boxed{C_5 = \left(\frac{43}{\beta}\right)\left(\frac{\omega_3}{\omega_5}\right)\left(\frac{V_T}{V_{c1} - 2V_{BE} - V_{margin}}\right)C_{jc}} \approx 4nF$$

$$\omega_6 = \frac{1}{\sqrt{L_6 C_5}}$$

$$\boxed{L_6 = \frac{1}{C_5\omega_6^2}}$$

$$\cancel{\omega_6 = 10\omega_3}$$

$$\cancel{\boxed{L_6 \le \frac{1}{100 C_5 \omega_3^2}}} \approx 0.25nH$$

But really try to do equal to.

Now the S2 design is still in terms of just $\{\omega_3, \omega_4, \omega_5^*\}$

$$I_{C3} = V_T C_{jc} \cdot \omega_3$$

$$R_{E3} = (R_{E1}) \frac{I_{C1}}{I_{C3}}$$

$$R_{C4} = \frac{3V_{BE} + 2V_{margin}}{2I_{C3}}$$

$$R_5 = (\beta) \frac{(V_{C1} - 2V_{BE} - V_{margin})}{I_{C3}}$$

$$R_7 = \frac{V_T \beta}{10 I_{C3}} \left(\frac{V_{cc}}{3V_{BE} + 2V_{margin}} \right)$$

$$R_8 = \frac{V_T \beta}{10 I_{C3}} \left(\frac{V_{cc}}{V_{cc} - 3V_{BE} - 2V_{margin}} \right)$$

$$C_{E3} = \frac{I_{C3}}{(V_{C1} - 2V_{BE} - V_{margin})\, \omega_4}$$

$$\cancel{G_{S2\_BP} = \frac{R_{C4} \| Z_{Load}}{R_{C1}} \approx \frac{3V_{BE} + 2V_{margin}}{2R_{C1} I_{C3}} = \frac{3V_{BE} + 2V_{margin}}{2R_{C1}(V_T C_{jc} \omega_3)}}$$

$$\cancel{G_{S2\_Bandpass} \approx \left(\frac{3V_{BE} + 2V_{margin}}{2V_T} \right) \frac{1}{R_{C1} C_{jc} \omega_3} \approx}$$