

# Source impedance and op-amp stability

Version 2, Marcel van de Gevel, 19 November 2024

Update history:

Version 1, 15 November 2024: first version

Version 2, 19 November 2024: loop gain calculation made more explicit, relation between loop gain and stability made more explicit, calculation about undamped LC parallel tanks added, document split up in sections

## 1. Introduction

It is obvious to me that for an amplifier with series feedback at the input and shunt feedback at the output, a too high source impedance with the wrong phase has the same kind of impact as too low load impedance with the wrong phase, because they are in the loop gain equation in a similar way. On a diyAudio thread, there were people stating the opposite, namely that source impedance has nothing to do with loop stability and that an RC network from input to ground (or an RC filter between the source and the amplifier) therefore cannot have a positive effect on stability. My point of view is explained in more detail in this document.

The small-signal stability of a feedback circuit as a function of the source impedance can be analysed in several ways, such as:

- A. Calculating the loop gain as a function of source impedance and applying the Nyquist stability criterion (or one of its simpler special cases, as are often used with Bode plots)
- B. Calculating the characteristic polynomial of the circuit (without even using the fact that it is a feedback circuit) as a function of source impedance and checking when the characteristic polynomial is Hurwitz, using the Routh-Hurwitz stability criterion
- C. Proving stability for an ideal source impedance (zero for a voltage amplifier), calculating the input impedance and checking the stability of the combination of the source impedance and the input impedance

In section 2, loop gain calculations are done because these show quite simply and quite elegantly that the effect of source impedance is more or less the dual of the effect of the load impedance, for an amplifier with series feedback at the input and shunt feedback at the output. The countermeasures are therefore also more or less each other's duals. No attempt is made to apply the Nyquist stability criterion, because I find it easier to use method C.

In section 3, method C is applied to the special case of a CMOS op-amp driven from a load with a resonance peak in its impedance. For simplicity, the calculation will be done for a simple LC parallel tank, but a shorted transmission line of a quarter wavelength long (for example, an audio cable of which the other end is shorted by a muting switch) has a rather similar impedance peak.

Some countermeasures are listed in section 4.

## 2. Loop gain calculation

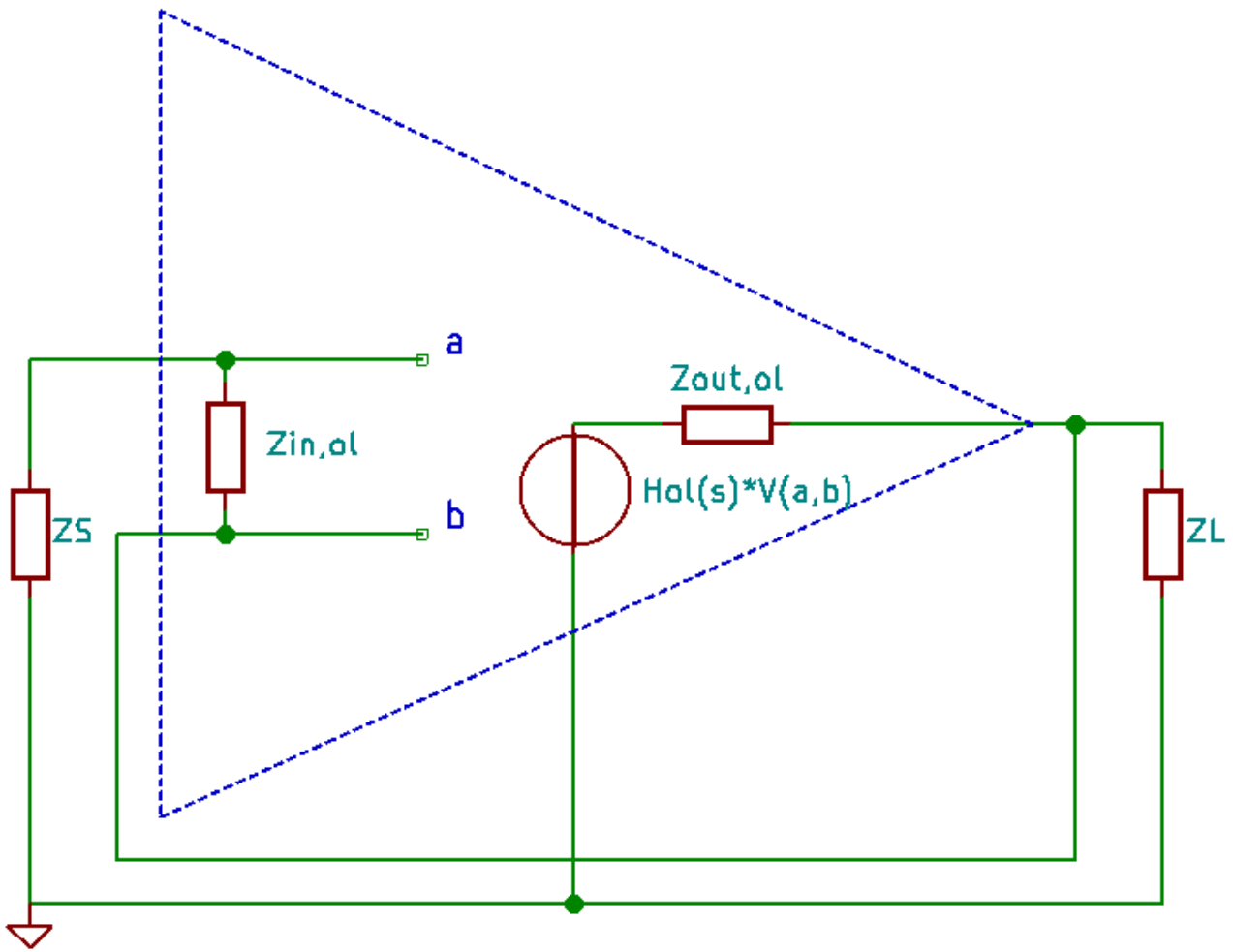


Figure 1: Simple model for an op-amp voltage follower with source and load impedance

Figure 1 shows a simple model for an op-amp, inside the dashed triangle, connected as a voltage follower and used with a source impedance  $Z_S$  and a load impedance  $Z_L$ . (A simple phono preamplifier with active RIAA correction is essentially a voltage follower at the frequency or frequencies where the loop gain passes through unity, as the feedback capacitors normally have negligible impedances at such frequencies.)

To keep the equations simple, I introduce

$$Z_{Lt} = \frac{Z_L (Z_{in,ol} + Z_S)}{Z_L + Z_{in,ol} + Z_S},$$

where  $Z_{Lt} \approx Z_L$  when  $Z_{in,ol} + Z_S \gg Z_L$ .

If you now see by inspection that the loop gain is simply

$$-H_{ol}(s) \frac{Z_{Lt}}{Z_{Lt} + Z_{out,ol}} \frac{Z_{in,ol}}{Z_{in,ol} + Z_S}$$

you can continue to section 2.7. If not, the next subsections may help.

## 2.1. Loop cutting method

There are several definitions and methods in use for calculating loop gain in feedback circuits. The simplest, but not universally applicable, method involves cutting open the loop at a place where this can be done without problems due to loading effects. In our case, this boils down to cutting open the loop at the output of the controlled source, applying a voltage there and looking at the ratio of the output signal of the controlled source to the applied voltage. It is illustrated in Figure 2.

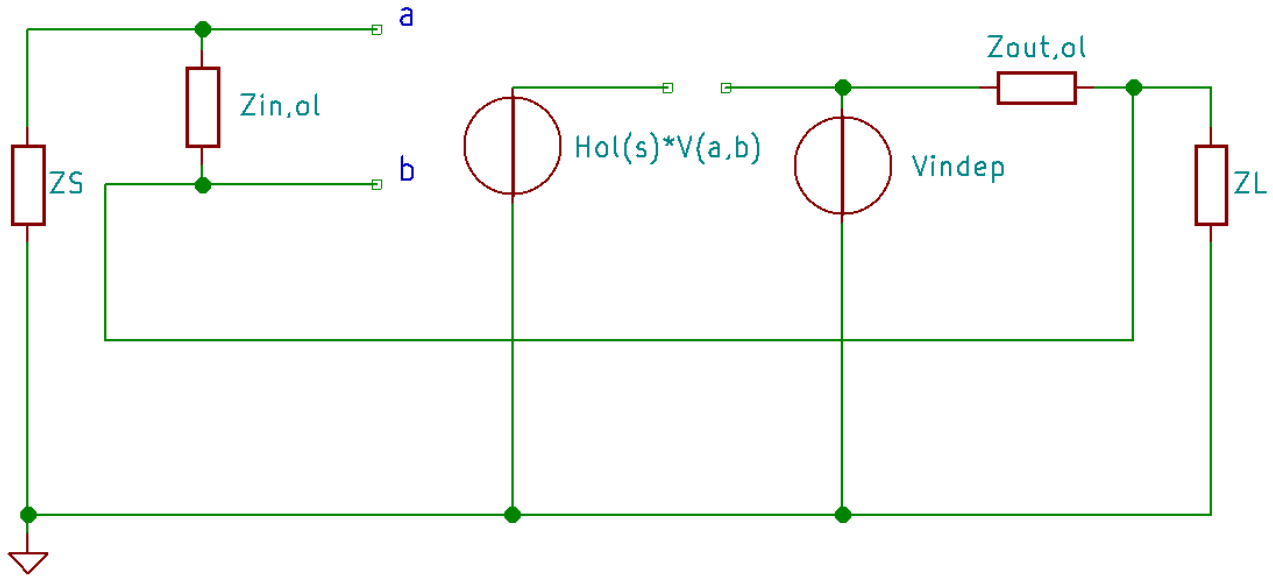


Figure 2: Loop gain calculation using the loop cutting method

It is clear from inspection that the voltage across  $Z_L$  is

$$V_{\text{indep}} \frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}}$$

Hence, the voltage between nodes b and a is

$$V_b - V_a = V_{\text{indep}} \frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}} \frac{Z_{\text{in,ol}}}{Z_{\text{in,ol}} + Z_S}$$

The output voltage of the controlled source is therefore

$$-V_{\text{indep}} \frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}} \frac{Z_{\text{in,ol}}}{Z_{\text{in,ol}} + Z_S} H_{\text{ol}}(s)$$

The loop gain is the ratio of this to the applied voltage, so

$$-\frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}} \frac{Z_{\text{in,ol}}}{Z_{\text{in,ol}} + Z_S} H_{\text{ol}}(s)$$

## 2.2. Asymptotic gain model

In the asymptotic gain model, no loops are cut open, but a suitably chosen controlled source is replaced with an independent one. The transfer from this independent source to what used to be the control voltage (or control current) of the former controlled source is called  $\beta$ . The gain of the

former controlled source is called  $A$ , the loop gain is now  $A\beta$ .

As we have only one controlled source, it is obvious which controlled source to replace with an independent one, see Figure 3.

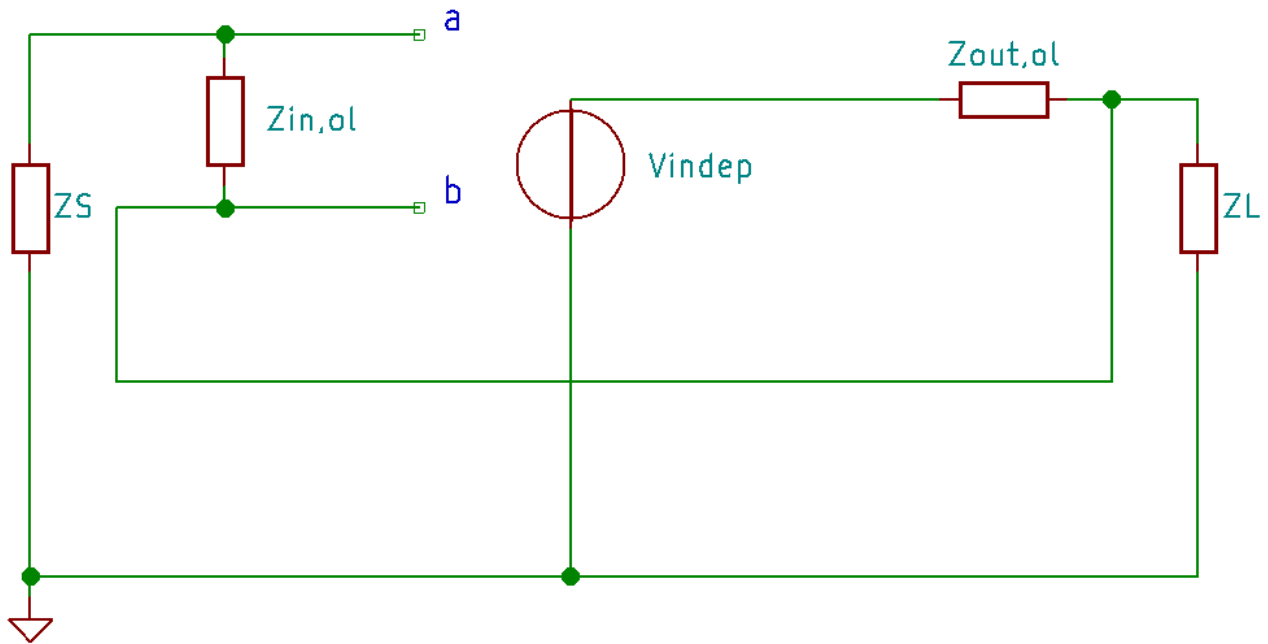


Figure 3: Schematic for loop gain calculation according to the asymptotic gain model

The voltage between nodes b and a is again

$$V_b - V_a = V_{\text{indep}} \frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}} \frac{Z_{\text{in,ol}}}{Z_{\text{in,ol}} + Z_S}$$

The transfer from  $V_{\text{indep}}$  to the voltage between nodes a and b is therefore

$$\beta = - \frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}} \frac{Z_{\text{in,ol}}}{Z_{\text{in,ol}} + Z_S}$$

The gain of the former controlled source was

$$A = H_{\text{ol}}(s)$$

so the loop gain is

$$A\beta = -H_{\text{ol}}(s) \frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}} \frac{Z_{\text{in,ol}}}{Z_{\text{in,ol}} + Z_S}$$

### 2.3. Bode's return ratio

As far as I know, Bode's return ratio is the opposite of the loop gain of the asymptotic gain model. It

is therefore  $H_{\text{ol}}(s) \frac{Z_{L_t}}{Z_{L_t} + Z_{\text{out,ol}}} \frac{Z_{\text{in,ol}}}{Z_{\text{in,ol}} + Z_S}$

### 2.4. Middlebrook method

I am not very familiar with the Middlebrook method, I hope I get the following right anyway. See

[https://www.arrl.org/files/file/QEX\\_Next\\_Issue/Mar-Apr\\_2011/QEX\\_3\\_11\\_Post.pdf](https://www.arrl.org/files/file/QEX_Next_Issue/Mar-Apr_2011/QEX_3_11_Post.pdf) for an explanation of the method.

When a current is injected into the controlled voltage source, all of it will flow into the voltage source, as the voltage source's impedance is zero. The magnitude of Middlebrook's current loop gain  $T_i$  is therefore infinite.

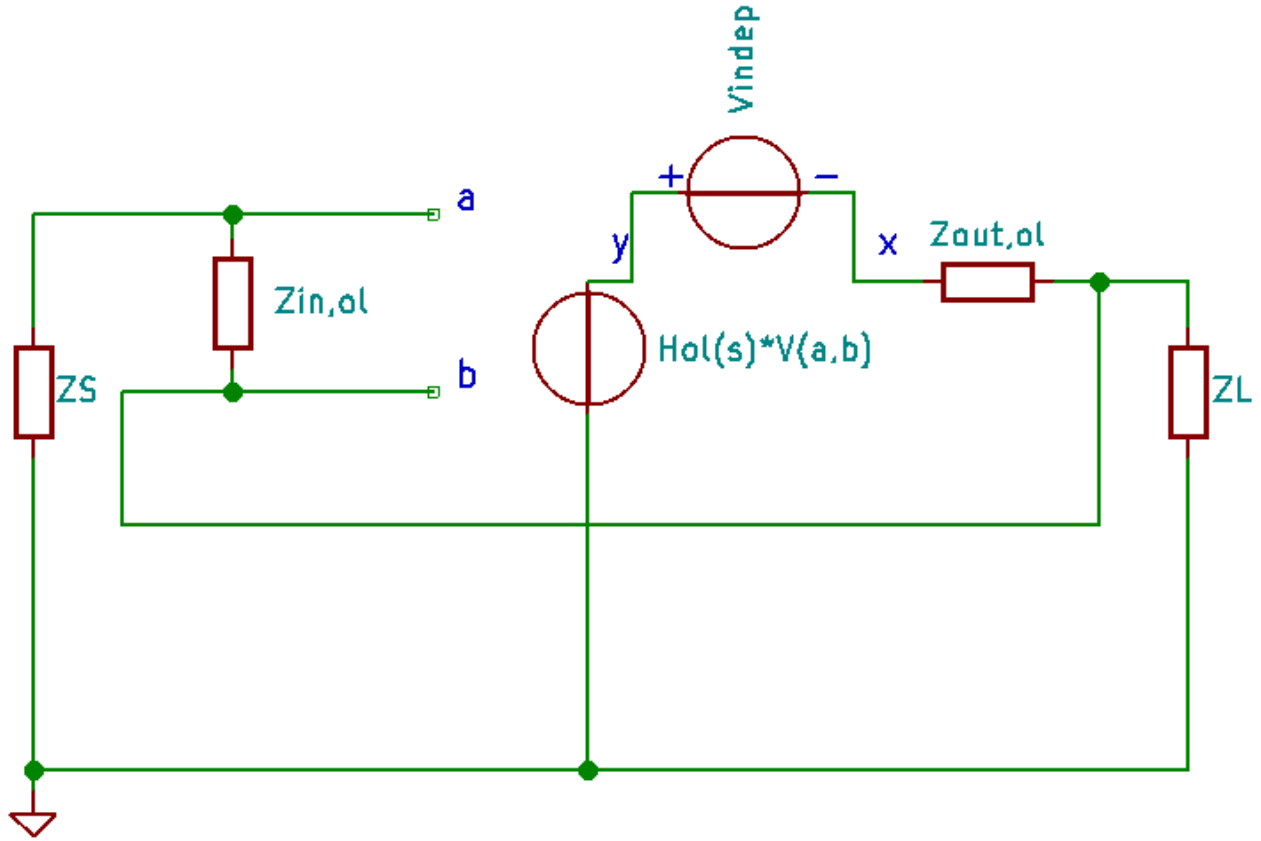


Figure 4: Schematic for calculating Middlebrook's  $T_v$

Middlebrook's voltage loop gain is the ratio of the voltages at nodes y and x in Figure 4.

$$V_b - V_a = V_x \frac{Z_{Lt}}{Z_{Lt} + Z_{out,ol}} \frac{Z_{in,ol}}{Z_{in,ol} + Z_S}$$

$$V_y = H_{ol}(s)(V_a - V_b) = -H_{ol}(s)V_x \frac{Z_{Lt}}{Z_{Lt} + Z_{out,ol}} \frac{Z_{in,ol}}{Z_{in,ol} + Z_S}$$

$$T_v = \frac{V_y}{V_x} = -H_{ol}(s) \frac{Z_{Lt}}{Z_{Lt} + Z_{out,ol}} \frac{Z_{in,ol}}{Z_{in,ol} + Z_S}$$

The total loop gain is  $T = \frac{T_v T_i - 1}{T_v + T_i - 2} \rightarrow T_v$  as  $|T_i| \rightarrow \infty$ . Hence,

$$T = T_v = -H_{ol}(s) \frac{Z_{Lt}}{Z_{Lt} + Z_{out,ol}} \frac{Z_{in,ol}}{Z_{in,ol} + Z_S}$$

## 2.5. Tian method

I'm even less familiar with the Tian method, which is supposed to be an improved Middlebrook

method. I will not calculate the Tian loop gain, but if anyone else wants to, they are very welcome.

## 2.6. Your pet loop gain definition/method

If your preferred loop gain definition leads to a different result, please write an explanation of the method and write out the equations, and send it to me.

## 2.7. Conclusions

All loop gain calculation methods and definitions I am familiar with result in a loop gain of

$$A\beta = -H_{ol}(s) \frac{Z_{Lt}}{Z_{Lt} + Z_{out,ol}} \frac{Z_{in,ol}}{Z_{in,ol} + Z_S}$$

give or take a minus sign. As all the impedances can be functions of frequency, it's actually more accurate to write

$$A(s)\beta(s) = -H_{ol}(s) \frac{Z_{Lt}(s)}{Z_{Lt}(s) + Z_{out,ol}(s)} \frac{Z_{in,ol}(s)}{Z_{in,ol}(s) + Z_S(s)} ,$$

but I won't because it only makes the equations look more complicated. The exact relation between loop gain and stability is given by the Nyquist stability criterion, but it usually boils down to the phase shift around the loop having to be less negative than  $-360^\circ$  at the frequency or frequencies where the magnitude of the loop gain passes through unity (that is, the phase margin has to be positive).

The loop gain equation shows that the load and the source impedance both affect the loop gain. The effect of the load impedance is small when  $Z_{Lt} \gg Z_{out,ol}$  at the frequencies of interest and the effect of the source impedance is small when  $Z_S \ll Z_{in,ol}$  at the frequencies of interest. For stability, "frequencies of interest" means around the frequency or frequencies where the magnitude of the loop gain passes through unity.

A common example of violating the condition  $Z_{Lt} \gg Z_{out,ol}$  around the frequency where the magnitude of the loop gain passes through unity is connecting a big capacitive load to the output of an op-amp. On top of that, the capacitive load causes extra negative phase shift in the loop gain. It is well known that this can cause oscillations. A series LC resonator or an unterminated transmission line with deep impedance dips at a low enough frequency can do the same.

Regarding phase, it is actually the difference in phase between the open-loop output impedance and the load impedance that matters, and the same holds for the open-loop input impedances and the source impedance. For example, when the open-loop output impedance is resistive,  $Z_{out,ol} = R_{out,ol}$  ,

while the load is capacitive,  $Z_{Lt} = \frac{1}{sC_L}$  , the output-impedance-related factor

$$\frac{Z_{Lt}(s)}{Z_{Lt}(s) + Z_{out,ol}(s)} = \frac{1}{1 + sR_{out,ol}C_L}$$

in the loop gain equation gives an extra negative real pole, resulting in extra negative phase shift and eating up phase margin. Similarly, when the open-loop input impedance is capacitive,

$Z_{in,ol} = \frac{1}{sC_{in}}$  and the source impedance is resistive,  $Z_S = R_S$ , the input-impedance-related factor

$$\frac{Z_{in,ol}}{Z_{in,ol} + Z_S} = \frac{1}{1 + sC_{in}R_S}$$

in the loop gain equation gives an extra negative real pole, resulting in extra negative phase shift and eating up phase margin. That is, you don't even need the source impedance to have a reactive part to lose phase margin when the open-loop input impedance is capacitive.

A practical example of violating the condition  $Z_S \ll Z_{in,ol}$  around the frequency where the magnitude of the loop gain passes through unity can be found in the thread

<https://www.diyaudio.com/community/threads/opa1656-phonopreamplifier-split-from-opa1656-thread.377331/post-6785883>, or at least that is my hypothesis. The thread starter had made the TI-

recommended moving-magnet preamplifier around an OPA1656, a low-noise CMOS op-amp with 53 MHz typical gain-bandwidth product, but without the 150 pF capacitor across the input, because the cartridge had a low recommended load capacitance.

The turntable features a switch that shorts the cartridge when the record has ended. At the end of the record, when the switch closed, there was always an enormous bang. My suspicion was that the shorted cable from the turntable to the phono preamplifier acted as a quarter-wave transmission line resonator, causing a too high impedance peak, causing the op-amp to oscillate. Some rectification effects then did the rest. Connecting a 220  $\Omega$ -15 pF RC series network across the input, the dual of the LR parallel network at an amplifier output, solved it: no bangs anymore. The time constant of the RC network was based on the estimated resonant frequency of the transmission line, the required capacitance was determined experimentally. Shorting switches are also used as mute switches in some DACs and preamplifiers.

The stability analysis for this case will be done in section 3, using negative conductances rather than loop gains. It's an equally valid method, that I find more convenient in this specific case.

### 3. Stability and negative resistances and conductances

Assuming that the circuit is designed to be stable when driven from a 0  $\Omega$  source, one can analyse the effect of the source impedance on the stability in terms of impedances and admittances. When you calculate the input impedance of the voltage follower, for simplicity assuming  $Z_{out,ol} = 0$ , you find that the real part of the input impedance can easily go negative, particularly for a fast low-noise MOS op-amp. This will be shown in section 3.1. The effect on the stability of an LC parallel tank will be treated in section 3.2, a parallel LC tank being a lumped model for a quarter-wave transmission line resonator. A more extensive analysis can be found in section 3.3.

### 3.1. Closed-loop input impedance at the positive op-amp input

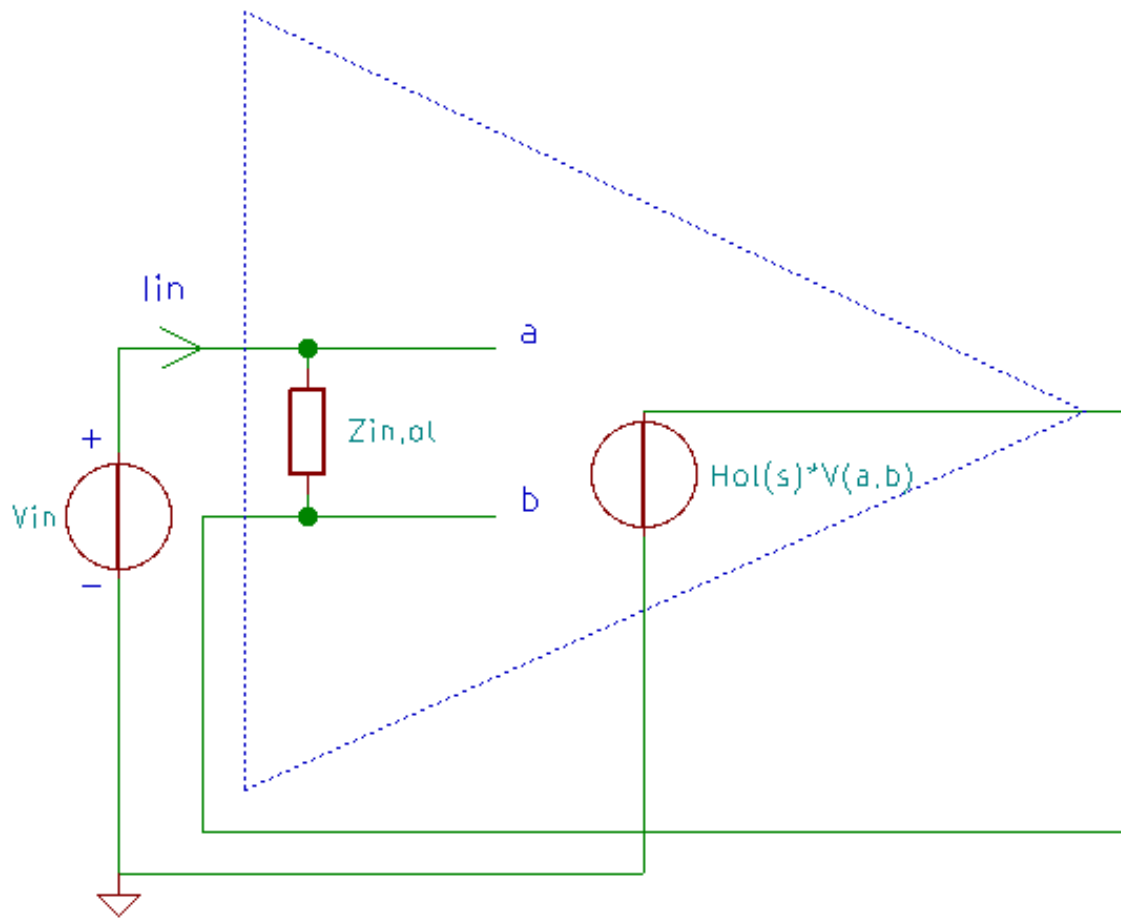


Figure 5: Schematic for calculating the input impedance

In Figure 5,

$$V_{in} = (H_{ol}(s) + 1) V_{ab}$$

$$V_{ab} = \frac{V_{in}}{H_{ol}(s) + 1}$$

$$I_{in} = \frac{V_{ab}}{Z_{in,ol}} = \frac{V_{in}}{(H_{ol}(s) + 1) Z_{in,ol}}$$

$$Z_{in, \text{ closed loop}} = \frac{V_{in}}{I_{in}} = (H_{ol}(s) + 1) Z_{in,ol}$$

When the open-loop gain drops off with frequency at a first-order rate and when the open-loop input impedance is capacitive:

$$H_{ol}(s) = \frac{\omega_{GBP}}{s}$$

$$Z_{in,ol} = \frac{1}{sC_{in,ol}}$$



$$Z_{\text{in, closed loop}} = \left( \frac{\omega_{\text{GBP}}}{s} + 1 \right) \frac{1}{s C_{\text{in,ol}}} = \frac{\omega_{\text{GBP}}}{s^2 C_{\text{in,ol}}} + \frac{1}{s C_{\text{in,ol}}}$$

When  $s = j\omega$ :

$$Z_{\text{in, closed loop}} = -\frac{\omega_{\text{GBP}}}{\omega^2 C_{\text{in,ol}}} - \frac{j}{\omega C_{\text{in,ol}}}$$

The first term represents a so-called frequency-dependent negative resistance, the second a series capacitance.

The differential open-loop input capacitance of an OPA1656 is specified to be 9.1 pF, see <https://www.ti.com/document-viewer/opa1656/datasheet> The large capacitance is no doubt due to the use of very large input MOSFETs to keep the  $1/f$  noise small.

The gain-bandwidth product is specified to be 53 MHz, so the theoretical input impedance is

4.45 MHz:  $-46810 \Omega - 3930 j\Omega$ , admittance  $-21.21 \mu\text{S} + 1.781 j \mu\text{S}$ , equivalent to  $-47140 \Omega$  in parallel with 0.0637 pF

10 MHz:  $-9269 \Omega - 1749 j\Omega$ , admittance  $-104.2 \mu\text{S} + 19.66 j \mu\text{S}$ , equivalent to  $-9599 \Omega$  in parallel with 0.3128 pF

50 MHz:  $-370.8 \Omega - 349.8 j\Omega$ , admittance  $-1427 \mu\text{S} + 1346 j \mu\text{S}$ , equivalent to  $-700.8 \Omega$  in parallel with 4.285 pF

The relevance of the 4.45 MHz impedance is that 4.45 MHz is roughly the frequency where the frequency-dependent negative resistance resonates with the 47 k $\Omega$  termination resistor of the moving-magnet phono preamplifier. Shunt capacitance would damp this resonance.

50 MHz is the quarter-wave resonant frequency of a 1 metre long, polyethylene-filled cable. 10 MHz is just a point in between. The 50 MHz input impedance is at best a very rough order-of-magnitude estimate, because the OPA1656's loop gain doesn't follow a -20 dB/decade slope anymore at such frequencies.

The estimated transmission line resonance frequency was 50 MHz. At this frequency, 220  $\Omega$  in series with 15 pF has an impedance of  $220 \Omega - 212.2 j\Omega$ , admittance  $2355 \mu\text{S} + 2271 j \mu\text{S}$ . To the extent that these calculations can be trusted, the conductance of the RC network is large enough to make the total conductance at 50 MHz positive. There is also the  $1/47 \text{ k}\Omega \approx 21.28 \mu\text{S}$  from the parallel resistor at the RIAA amplifier input, but that is only a small contribution.

### 3.2. Effect on the stability of an LC parallel tank

To simplify the stability calculation, an LC parallel tank will be used as a lumped approximation of a quarter-wave transmission line that is shorted at its end. They have quite similar behaviour around the resonance frequency.

Also to simplify the calculation, the input impedance of the amplifier will be represented by the parallel connection of a negative conductance and a capacitance. For a given frequency, it is straightforward to transform a frequency-dependent negative resistance with a series capacitance into this form, as was already done in section 3.1. See section 3.3 for an analysis without this

simplification.

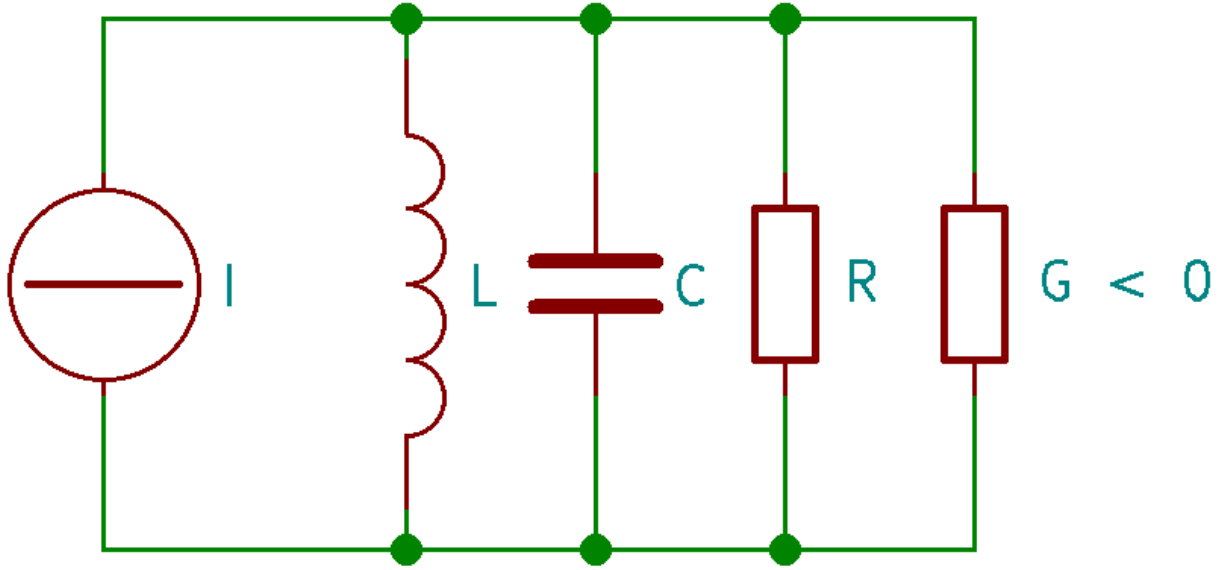


Figure 6: Parallel connection of an LC tank, a resistor and a negative conductance.  $L$ ,  $R$  and  $C$  are all greater than 0, while  $G < 0$ .

In Figure 6, the resistor represents the transmission line losses and the 47 k $\Omega$  resistor at the phono preamplifier input, if the amplifier is a phono preamplifier. The LC tank represents the shorted quarter-wave transmission line. The equivalent parallel capacitance at the positive op-amp input adds to  $C$ , the equivalent input conductance at the positive op-amp input is  $G$ , which is negative.

The admittance of the circuit is

$$Y = \frac{1}{sL} + sC + \frac{1}{R} + G = \frac{s^2 LC + sL \left( \frac{1}{R} + G \right) + 1}{sL}$$

Hence, the impedance is

$$Z = \frac{1}{Y} = \frac{sL}{s^2 LC + sL \left( \frac{1}{R} + G \right) + 1}$$

The impedance is the transfer from the current  $I$  to the voltage across the circuit. Its denominator is the circuit's characteristic polynomial, so equating it to zero gives the poles:

$$s^2 LC + sL \left( \frac{1}{R} + G \right) + 1 = 0$$

$$s = \frac{-L \left( \frac{1}{R} + G \right) \pm \sqrt{L^2 \left( \frac{1}{R} + G \right)^2 - 4LC}}{2LC}$$

The poles are both in the right half plane when  $L \left( \frac{1}{R} + G \right) < 0$ , that is,  $G < -\frac{1}{R}$

The circuit is therefore instable when  $G < -\frac{1}{R}$ , so when the parallel connection of  $G$  and  $R$  corresponds to a negative conductance.

### 3.3. More accurate analysis using the Routh-Hurwitz stability criterion

In this subsection, the shorted quarter-wave transmission line is still approximated with its lumped representation, an LC parallel network, but the amplifier input impedance will not be transformed, but will just be a frequency-dependent negative resistance with a series capacitance. The resulting mathematics are more complicated than those in section 3.2, while the conclusion remains the same.

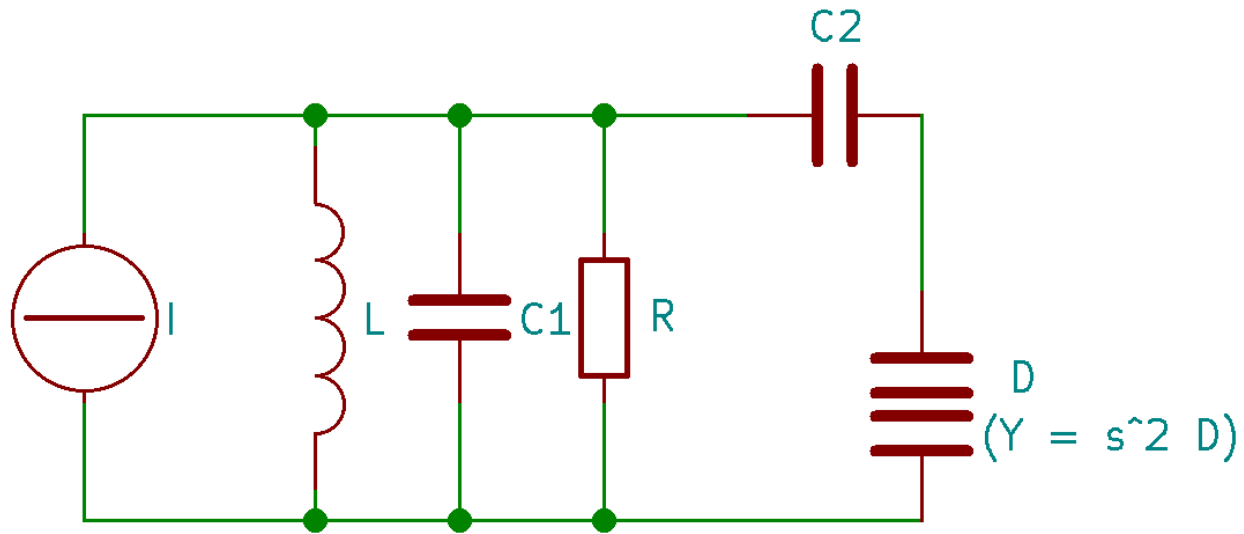


Figure 7: RLC parallel tank connected to a capacitance in series with an FDNR

The admittance of Figure 7 is

$$\frac{1}{sL} + sC_1 + \frac{1}{R} + \frac{1}{\frac{1}{sC_2} + \frac{1}{s^2 D}} = \frac{1}{sL} + sC_1 + \frac{1}{R} + \frac{s^3 C_2 D}{s^2 D + s C_2}$$

which (if I made no mistakes) can be rewritten as

$$\frac{s^3 L D (C_1 + C_2) + s^2 L \left( C_1 C_2 + \frac{D}{R} \right) + s \left( L \frac{C_2}{R} + D \right) + C_2}{s^2 L D + s L C_2}$$

The numerator of the admittance is the denominator of the impedance is the characteristic polynomial. Instead of trying to solve the poles, the Routh-Hurwitz criterion can be used to check if the characteristic polynomial is Hurwitz (stable).

According to [https://en.wikipedia.org/wiki/Routh–Hurwitz\\_stability\\_criterion](https://en.wikipedia.org/wiki/Routh%E2%80%93Hurwitz_stability_criterion), a third-order polynomial is Hurwitz when all its coefficients have the same sign and

$$a_2 a_1 - a_3 a_0 > 0$$

All coefficients are positive, and the other condition corresponds to

$$L\left(C_1 C_2 + \frac{D}{R}\right)\left(L \frac{C_2}{R} + D\right) - L D (C_1 + C_2) C_2 > 0$$

or, as  $L > 0$  and  $R > 0$ ,

$$(R C_1 C_2 + D)(L C_2 + D R) - D R^2 (C_1 + C_2) C_2 > 0$$

It is clear that this criterion is met for  $D = 0$  and for  $D \rightarrow \infty$ . There may be a region in between where instability occurs. The edges of this region are given by

$$(R C_1 C_2 + D)(L C_2 + D R) - D R^2 (C_1 + C_2) C_2 = 0$$

$$R D^2 + (L C_2 - R^2 C_2^2) D + L R C_1 C_2^2 = 0$$

$$D = \frac{(R^2 C_2^2 - L C_2) \pm \sqrt{(R^2 C_2^2 - L C_2)^2 - 4 L R^2 C_1 C_2^2}}{2 R}$$

$L$  and  $C_1$  will be used to model a quarter-wave transmission line with shorted end around its resonant frequency. As shown in

[https://eng.libretexts.org/Bookshelves/Electrical\\_Engineering/Electro-Optics/Book%3A\\_Electromagnetics\\_I\\_\(Ellingson\)/03%3A\\_Transmission\\_Lines/3.16%3A\\_Input\\_Impedance\\_for\\_Open-\\_and\\_Short-Circuit\\_Terminations](https://eng.libretexts.org/Bookshelves/Electrical_Engineering/Electro-Optics/Book%3A_Electromagnetics_I_(Ellingson)/03%3A_Transmission_Lines/3.16%3A_Input_Impedance_for_Open-_and_Short-Circuit_Terminations), the impedance of an ideal shorted transmission line is

$$Z_{\text{line}} = j Z_0 \tan(\beta l)$$

so

$$Y_{\text{line}} = -j \frac{Y_0}{\tan(\beta l)} \approx j Y_0 \left( \beta l - \frac{\pi}{2} \right) = j Y_0 \left( \omega \frac{l}{v} - \frac{\pi}{2} \right)$$

where the approximation holds when the line is approximately a quarter wave long (first-order Taylor approximation). The  $v$  in the last term is the propagation velocity in the line.

$$\frac{\partial Y_{\text{line}}}{\partial \omega} = j Y_0 \frac{l}{v} \quad \text{around resonance.}$$

An LC parallel network with infinite  $Q$  has an admittance

$$Y_{LC} = j \left( \omega C - \frac{1}{\omega L} \right)$$

so

$$\frac{\partial Y_{LC}}{\partial \omega} = j \left( C + \frac{1}{\omega^2 L} \right) \approx j 2 C$$

where the approximation holds around resonance, where  $\omega \approx \frac{1}{\sqrt{LC}}$

Equating the derivatives of the admittance to radian frequency around the resonant frequency  $f_{\text{res}}$  (quarter-wave resonant frequency for the transmission line and LC resonant frequency for the parallel tank):

$$C = \frac{1}{2} Y_0 \frac{l}{v} = \frac{1}{8 Z_0 f_{\text{res}}}$$

$$L = \frac{1}{4 \pi^2 f_{\text{res}}^2 C}$$

In the case of a 1 metre long, 75  $\Omega$  cable with a wave velocity of 200 000 km/s:

$$f_{\text{res}} = 50 \text{ MHz}$$

$$C = 33.3333... \text{ pF}$$

$$L \approx 303.96355 \text{ nH}$$

Hence, assume

$$L \approx 303.96355 \text{ nH}$$

$$C_1 = 33.3333... \text{ pF}$$

$$C_2 = 9.1 \text{ pF (the OPA1656 differential input capacitance)}$$

I'm not sure what a realistic value for  $R$  would be, but I guess it is somewhere between 3 k $\Omega$  and 47 k $\Omega$  depending on the quality of the transmission line / cable of the turntable. Using

$$D = \frac{(R^2 C_2^2 - L C_2) \pm \sqrt{(R^2 C_2^2 - L C_2)^2 - 4 L R^2 C_1 C_2^2}}{2 R}$$

we then find:

$$R = 3 \text{ k}\Omega:$$

oscillatory region from  $D \approx 3.4377 \bullet 10^{-21}$  sF to  $2.4407 \bullet 10^{-19}$  sF, where sF stands for second farad.

$$R = 47 \text{ k}\Omega:$$

oscillatory region from  $D \approx 2.1559 \bullet 10^{-22}$  sF to  $3.8918 \bullet 10^{-18}$  sF.

As follows from the derivation in section 3.1, the actual value of  $D$  is

$$D = \frac{C_{\text{in,ol}}}{\omega_{\text{GBP}}} = \frac{C_{\text{in,ol}}}{2 \pi f_{\text{GBP}}}$$

where  $C_{\text{in,ol}}$  is what is called  $C_2$  in this subsection. That is,

$$D = \frac{9.1 \text{ pF}}{2 \pi 53 \text{ MHz}} \approx 2.73266 \cdot 10^{-20} \text{ sF}$$

which lies in the oscillatory region in both cases.

## 4. Solutions

The usual solution for load impedances affecting the stability is to put a series resistor between the op-amp and the load. If that isn't acceptable, an LR parallel network or a first-order series filter can be used. (The use of a first-order series filter was proposed by A. N. Thiele for audio power amplifiers many years ago, it also makes the amplifiers more robust against RF signals picked up by

the loudspeaker cable.)

The dual can be used for the input, although things are complicated somewhat by the fact that the open-loop input impedance is often capacitive and that the resulting closed-loop input impedance has frequency-dependent negative resistance (FDNR) behaviour. An FDNR can resonate with a resistor, capacitance is needed to damp such a resonance. Simply connecting a resistor across the input, the dual of the output series resistor, is therefore not necessarily enough, you really need an RC network. In the thread about the phono amplifier that went bang, an RC series network was shunted across the input, but it is generally better to use an RC low-pass filter, as that also provides some protection against RF signals. This has in fact been standard practice for decades.