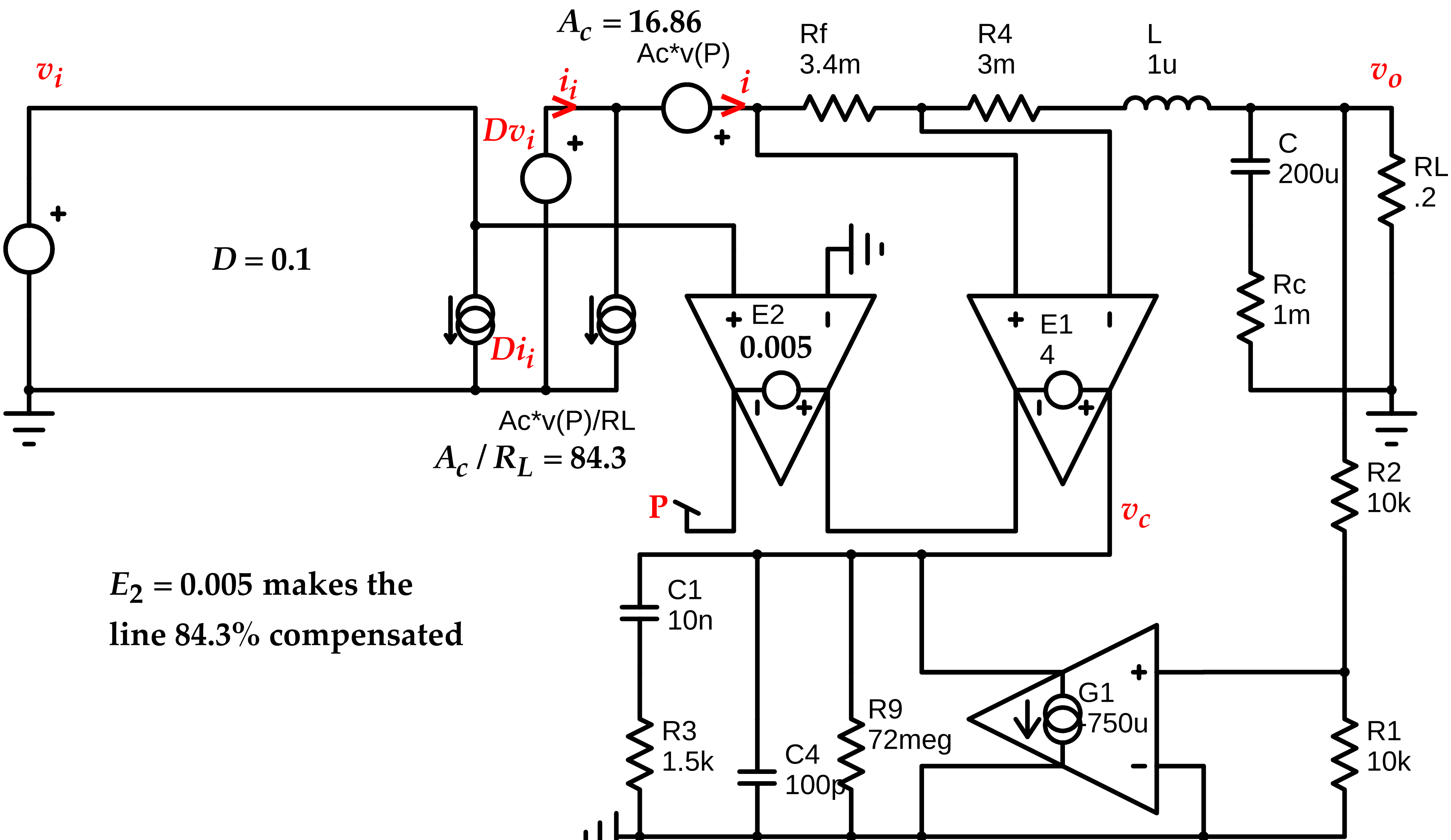
EXAMPLE

18. CURRENT-PROGRAMMED SWITCHED-MODE REGULATOR

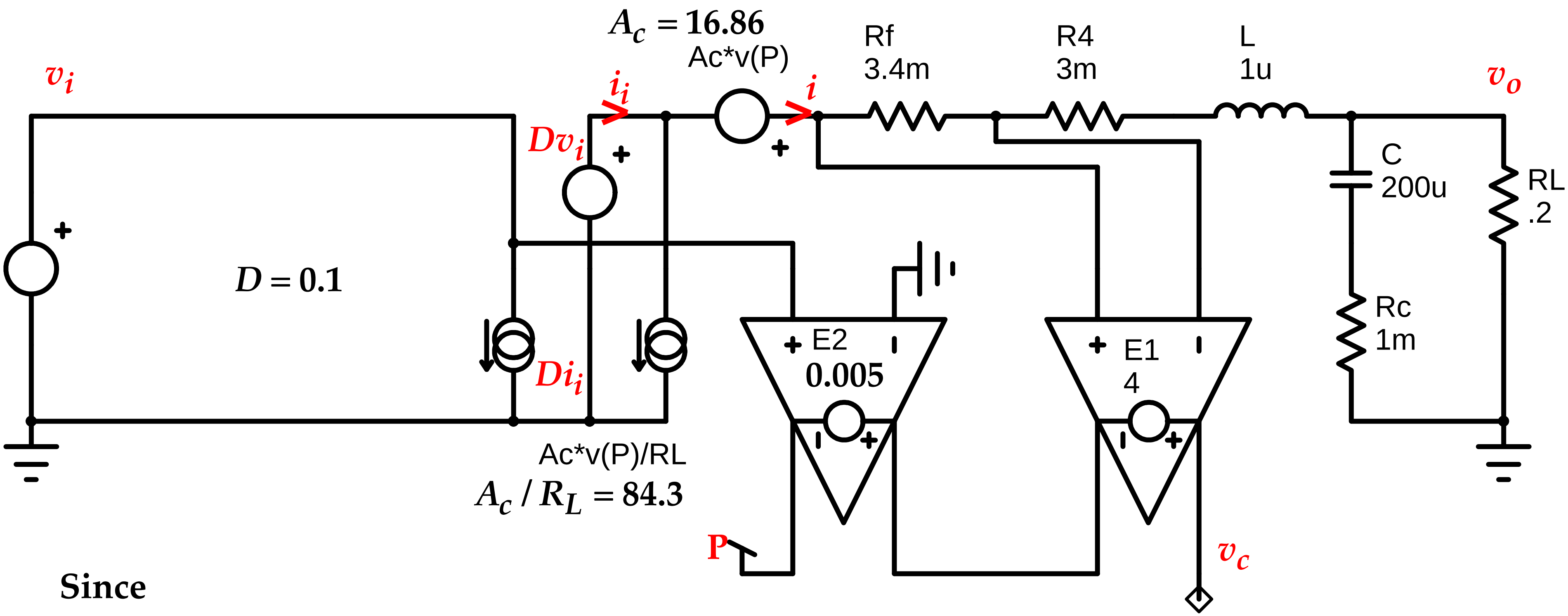


$E_2 = 0.005$ makes the line 84.3% compensated

The power stage has an internal current loop and a voltage feedforward loop.

STEP 1

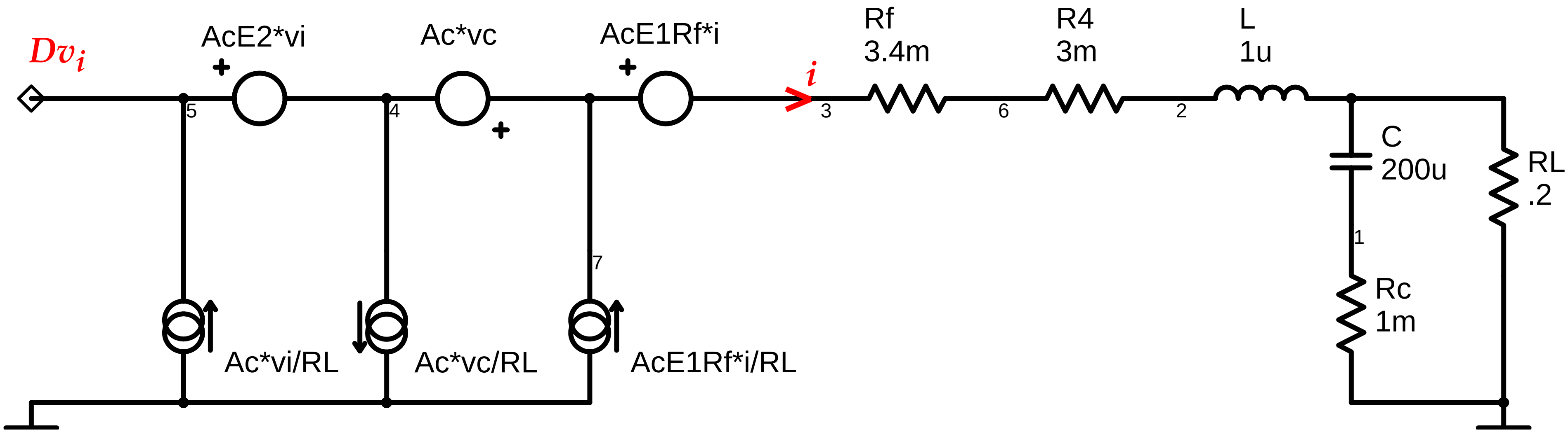
Strategy: absorb both loops into an equivalent power stage model, by Doing Some Algebra on the Circuit Diagram (Ch.2)



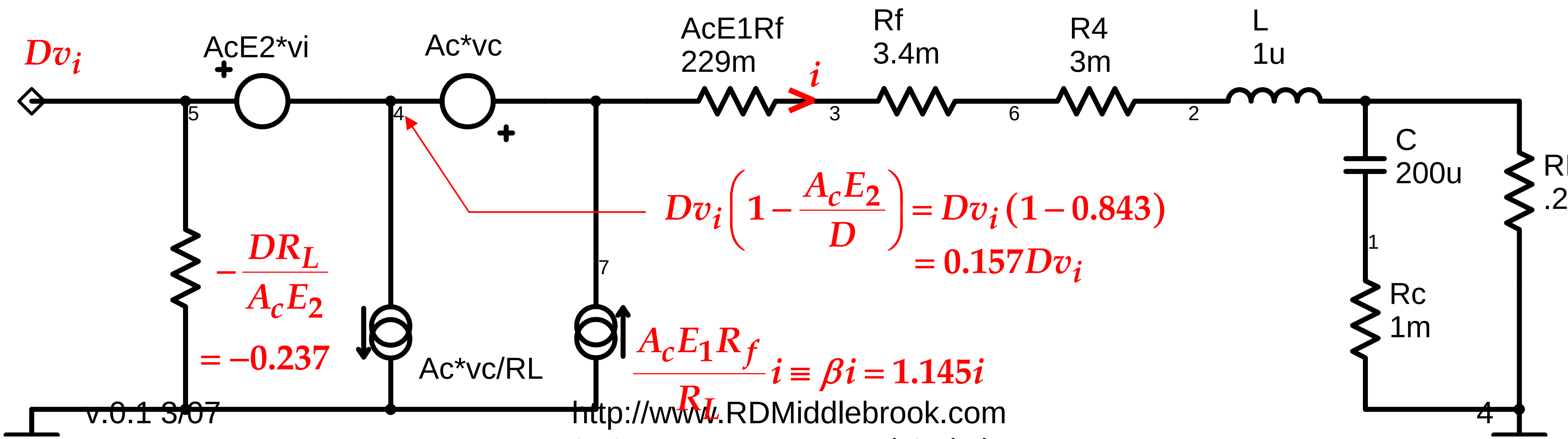
Since

$$v(P) = -E_2 v_i + v_c - E_1 R_f i$$

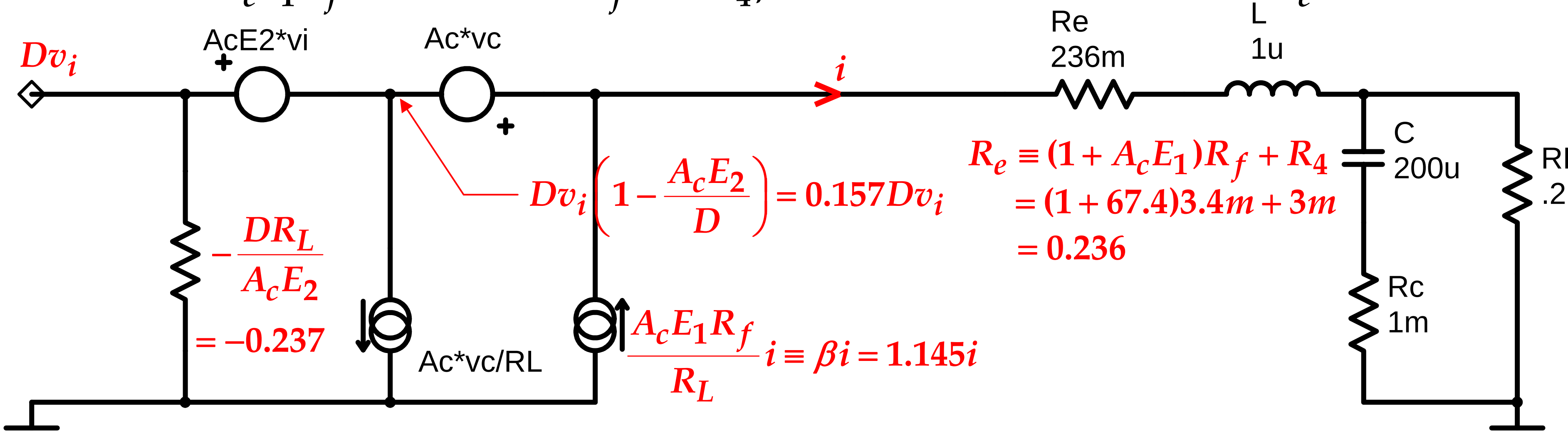
split the voltage and current modulation generators into three:



Since the voltage generator $A_c E_1 R_f i$ is proportional to the current through it, it can be replaced by a resistance $A_c E_1 R_f$. Also, since the current generator $A_c E_2 v_i / R_L$ is proportional to the voltage Dv_i across it, it can be replaced by a (negative) resistance $-DR_L / A_c E_2$.



The consequence of the current loop is that it introduces a (lossless) damping resistance $A_c E_1 R_f$ in series with R_f and R_4, to form an effective resistance R_e :

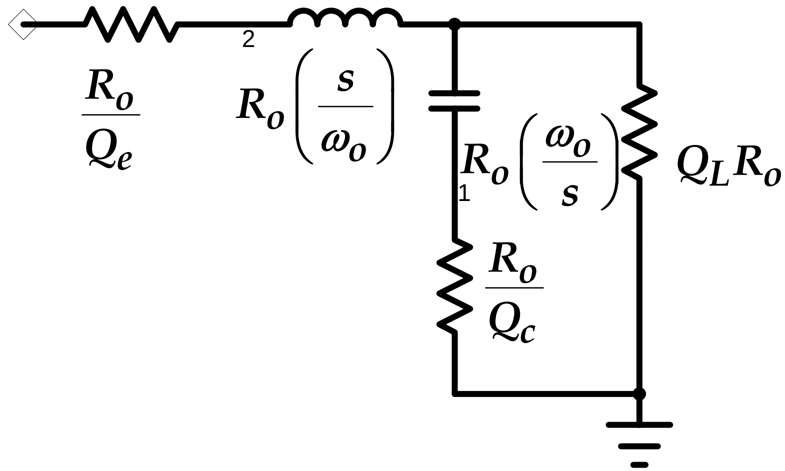


STEP 2

Normalize the element values in the filter (Ch. 5):

$$\omega_o \equiv \frac{1}{\sqrt{LC}} \qquad f_o = 11.2kHz \qquad R_o \equiv \sqrt{\frac{L}{C}} = 70.7m$$

$$Q_e \equiv \frac{R_o}{R_e} = 0.300 \qquad Q_c \equiv \frac{R_o}{R_c} = 70.7 \qquad Q_L \equiv \frac{R_L}{R_o} = 2.83$$



The voltage transfer function H for this triple-damped RLC filter was obtained in Ch. 12:

$\frac{1}{Q_cQ_L}$

$$H = \frac{1}{1+\frac{1}{Q_eQ_L}} \; \frac{1 + \frac{1}{Q_c}\left(\frac{s}{\omega_o}\right)}{1 + \left(\frac{1}{Q_L\left(1+\frac{1}{Q_eQ_L}\right)} + \frac{1}{Q_c} + \frac{1}{Q_e}\,\frac{1}{1+\frac{1}{Q_eQ_L}}\right)\left(\frac{s}{\omega_o}\right) + \frac{1+\frac{1}{Q_eQ_c}}{1+\frac{1}{Q_eQ_L}}\left(\frac{s}{\omega_o}\right)^2}$$

$\frac{1}{Q_cQ_L}$

$$= \frac{1}{1+\frac{1}{Q_eQ_L}} \; \frac{1 + \frac{1}{Q_c}\left(\frac{s}{\omega_o}\right)}{1 + \frac{\frac{1}{Q_e} + \frac{1}{Q_L} + \frac{1+1/Q_eQ_L}{Q_c}}{1+\frac{1}{Q_eQ_L}}\left(\frac{s}{\omega_o}\right) + \frac{1+\frac{1}{Q_eQ_c}}{1+\frac{1}{Q_eQ_L}}\left(\frac{s}{\omega_o}\right)^2}$$

Q_cQ_L

For $Q_eQ_c \gg 1$, $Q_eQ_L \gg 1$, the result reduces to that previously obtained by extrapolation of the result for $Q_c = \infty$:

$$H \approx \frac{1 + \frac{1}{Q_c}\left(\frac{s}{\omega_o}\right)}{1 + \left(\frac{1}{Q_e} + \frac{1}{Q_c} + \frac{1}{Q_L}\right)\left(\frac{s}{\omega_o}\right) + \left(\frac{s}{\omega_o}\right)^2}$$

For the LM3495, however, the inequality $Q_cQ_L = 70.7 * 2.83 = 200 >> 1$ still holds, but the inequality $Q_eQ_L = 0.300 * 2.83 = 0.848 >> 1$ does not.

The result therefore becomes

$$H = \frac{1}{1+\frac{1}{Q_eQ_L}} \frac{1+\frac{1}{Q_c}\left(\frac{s}{\omega_o}\right)}{1+\frac{\frac{1}{Q_e}+\frac{1}{Q_L}+\frac{1+1/Q_eQ_L}{Q_c}}{1+\frac{1}{Q_eQ_L}}\left(\frac{s}{\omega_o}\right)+\frac{1+\frac{1}{Q_eQ_c}}{1+\frac{1}{Q_eQ_L}}\left(\frac{s}{\omega_o}\right)^2}$$

$\frac{1}{Q_cQ_L}$

$$H = \frac{1}{1+\frac{1}{Q_eQ_L}} \frac{1+\frac{1}{Q_c}\left(\frac{s}{\omega_o}\right)}{1+\frac{1}{1+\frac{1}{Q_eQ_L}}\frac{1}{Q_e\|Q_L}\left(\frac{s}{\omega_o}\right)+\frac{1}{1+\frac{1}{Q_eQ_L}}\left(\frac{s}{\omega_o}\right)^2}$$

where $Q_e\|Q_L = 0.300\|2.83 = 0.271$ **and** $\frac{1}{1+\frac{1}{Q_eQ_L}} = \frac{1}{2.18} = 0.459$

The Q_t of the denominator quadratic is (Ch. 4)

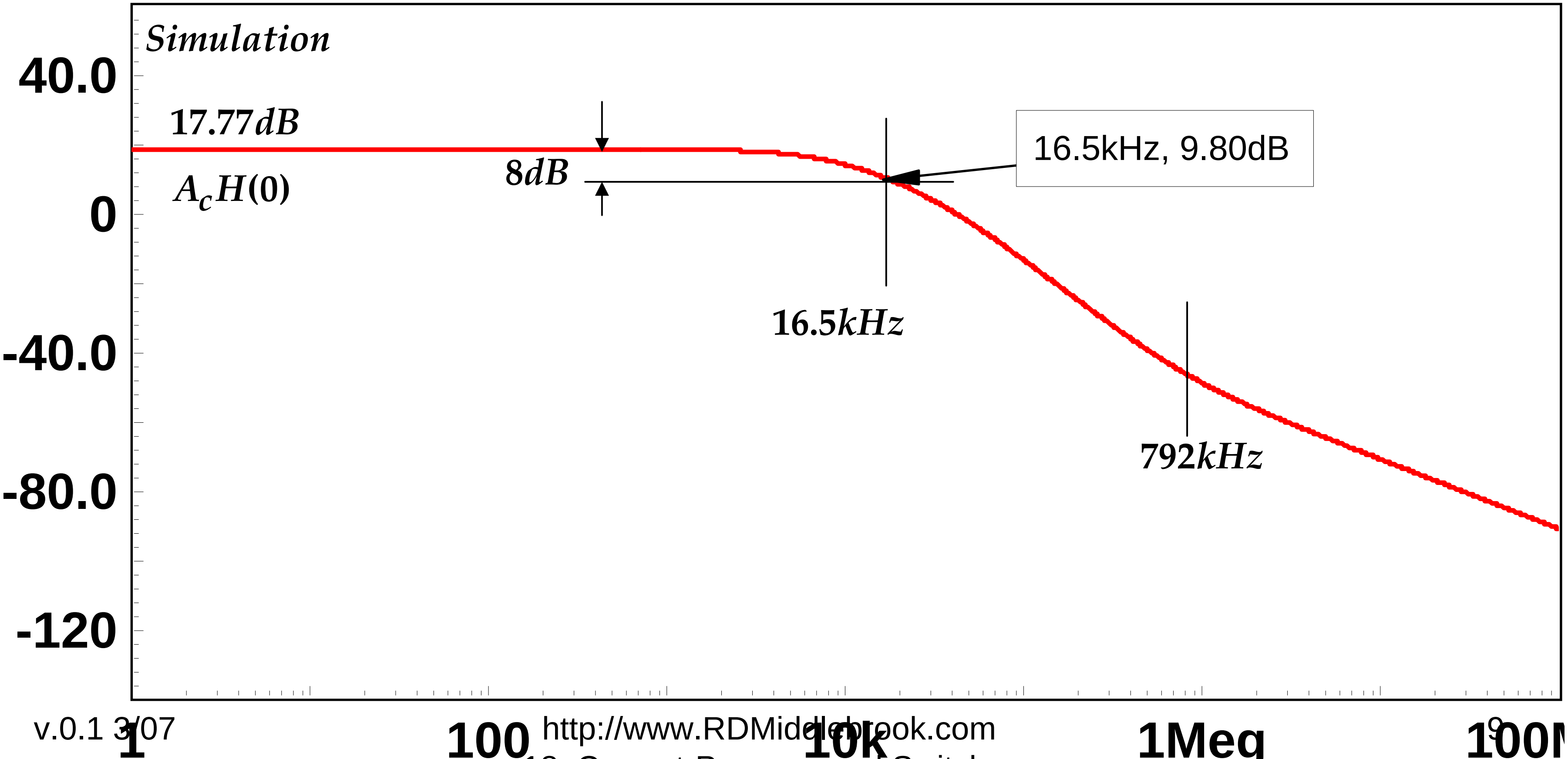
$$Q_t \equiv \frac{\sqrt{ac}}{b} = \sqrt{\left(1+\frac{1}{Q_eQ_L}\right)}\left(Q_e\|Q_L\right) = 0.400,\ Q_t = -7.96dB$$

$$\omega_o' \equiv \sqrt{\left(1+\frac{1}{Q_eQ_L}\right)}\omega_o, \qquad f_o' = \sqrt{2.18} * 11.2kHz = 16.5kHz$$

The power stage control-to-output voltage gain is $v_o / v_c \big|_{v_i=0} = A_c H$ **where** $A_c = 16.86$.

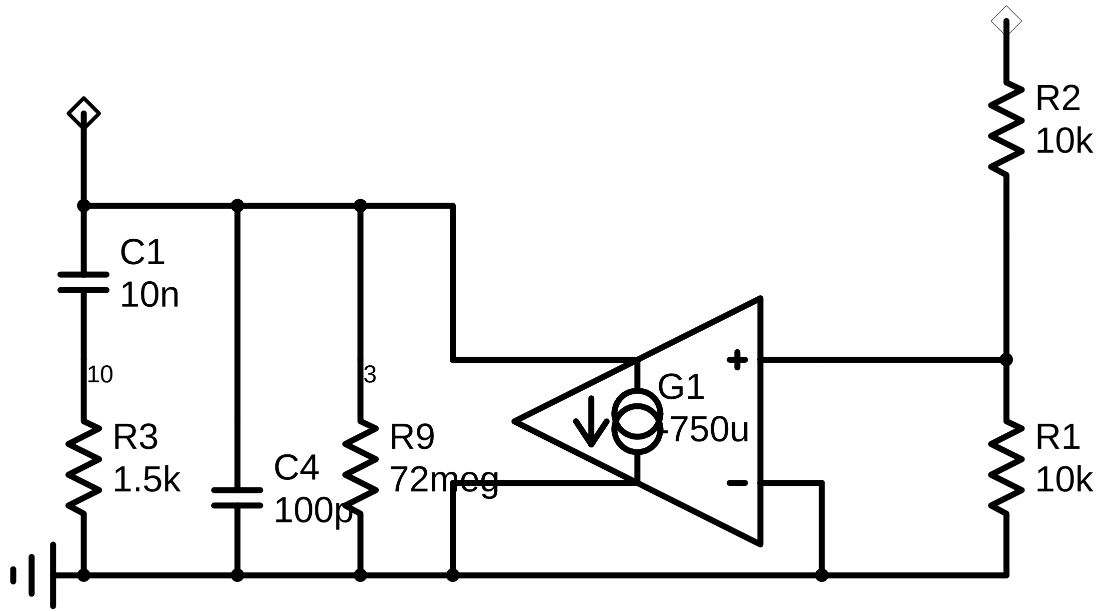
Insertion of numbers gives

$$A_c H = 17.77dB \frac{1 + \frac{s/2\pi}{792kHz}}{1 + \frac{1}{0.400}\left(\frac{s/2\pi}{16.5kHz}\right) + \left(\frac{s/2\pi}{16.5kHz}\right)^2}$$



STEP 3

Error Amplifier



The flat gain that sets the loop gain crossover occurs when C_1 is short, and C_4 open, so take this gain as reference (Ch. 3). The two poles are obviously well separated, so the gain can be written by inspection as

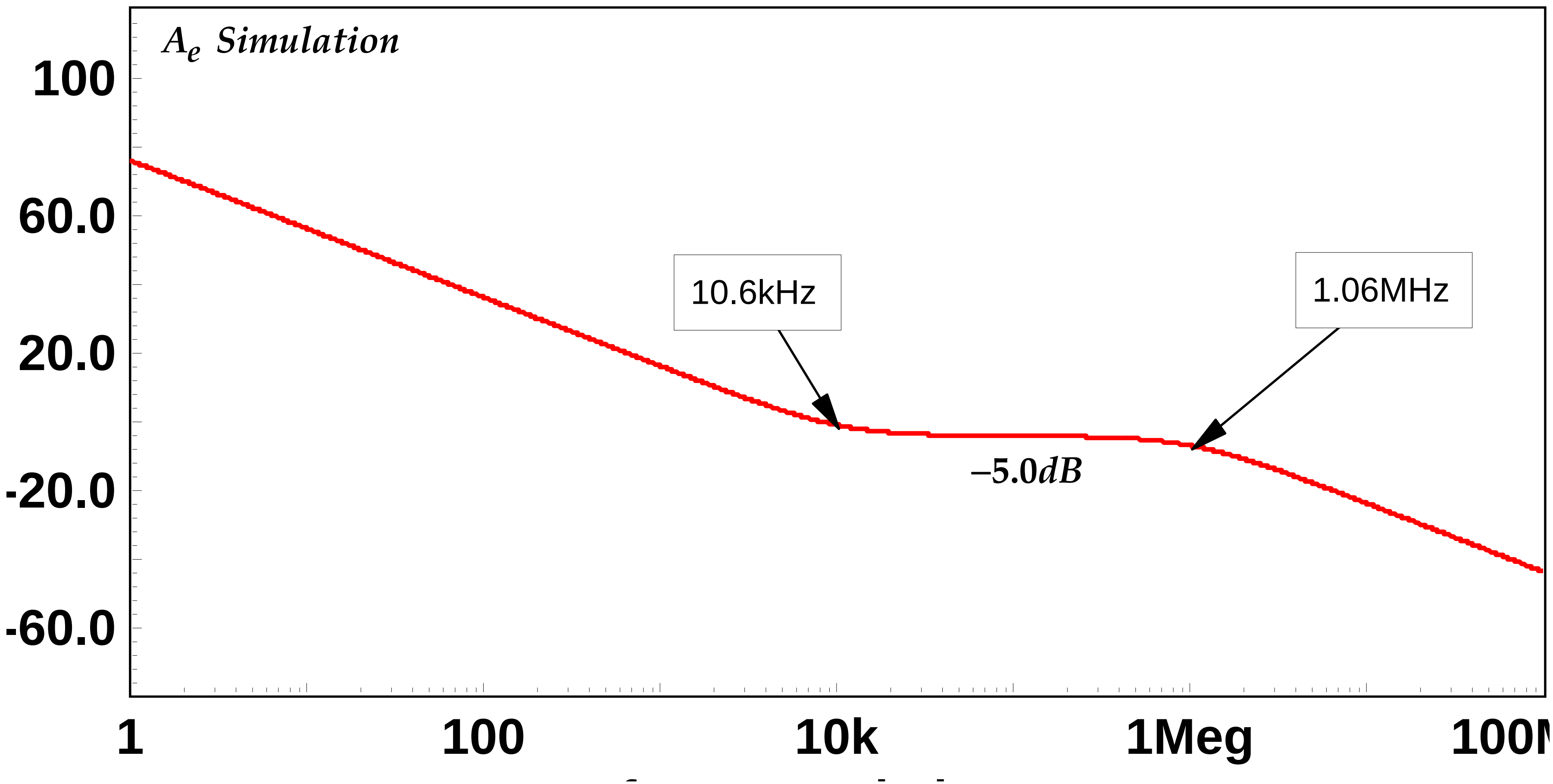
$$A = A_{em} \frac{1}{\left(\frac{1}{sC_1R_3}+1\right)\left(1+sC_4R_3\right)}$$

where

$$A_{em} = \frac{R_1}{R_1+R_2} G1 R_3 = 0.563 \Rightarrow -5.0dB$$

With substitution of numbers,

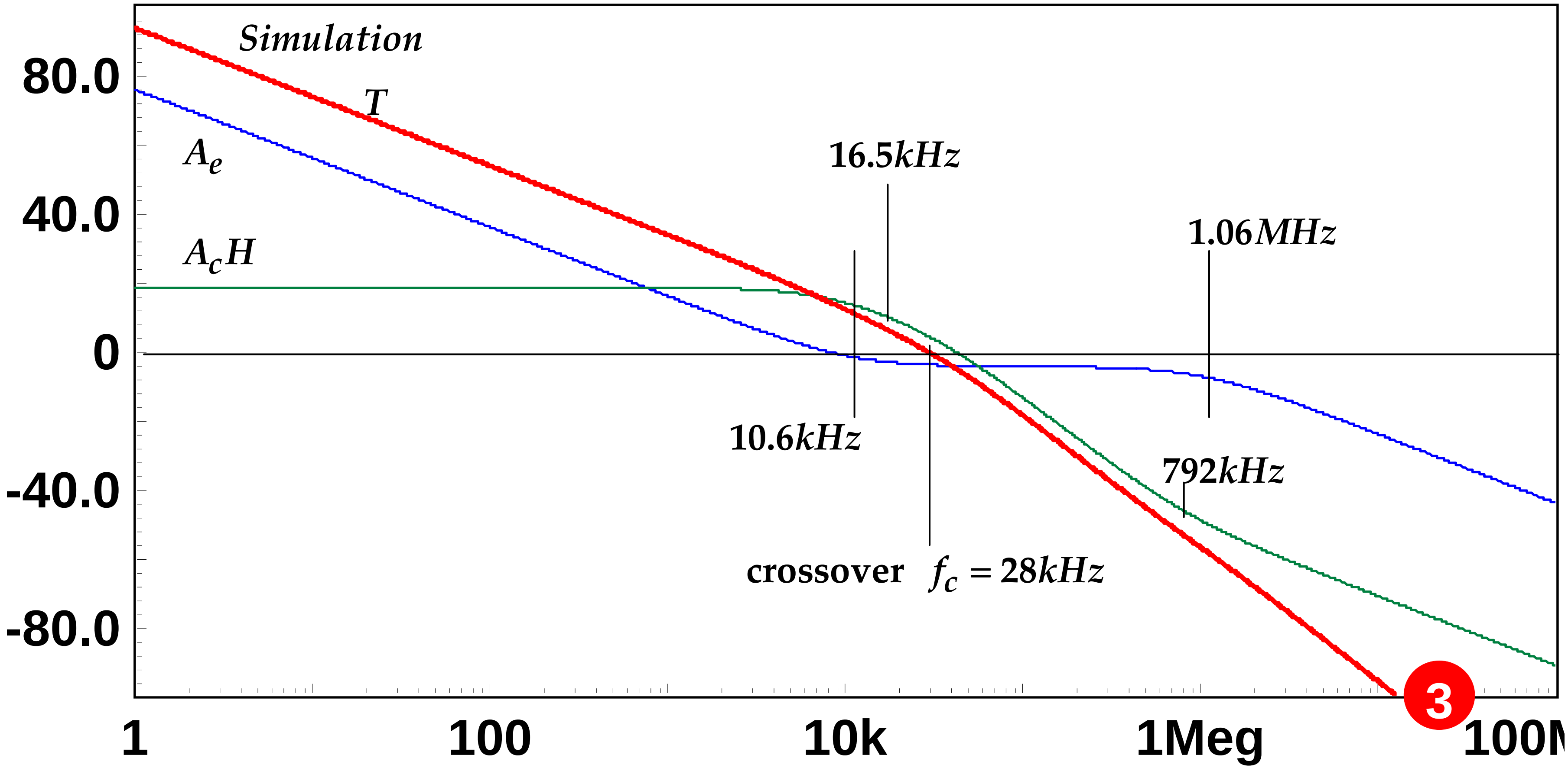
$$A_e = -5dB \frac{1}{\left(\frac{10.6kHz}{s/2\pi}+1\right)\left(1+\frac{s/2\pi}{1.06MHz}\right)}$$


A_e Simulation
100
60.0
20.0
-20.0
-60.0
10.6kHz
1.06MHz
$-5.0dB$
1
100
10k
1Meg

STEP 4

The loop gain T is:

$$T = A_e A_c H = 12.77dB \frac{1+\frac{s/2\pi}{792kHz}}{\left(\frac{10.6kHz}{s/2\pi}+1\right)\left[1+\frac{1}{0.400}\left(\frac{s/2\pi}{16.5kHz}\right)+\left(\frac{s/2\pi}{16.5kHz}\right)^2\right]\left(1+\frac{s/2\pi}{1.06MHz}\right)}$$



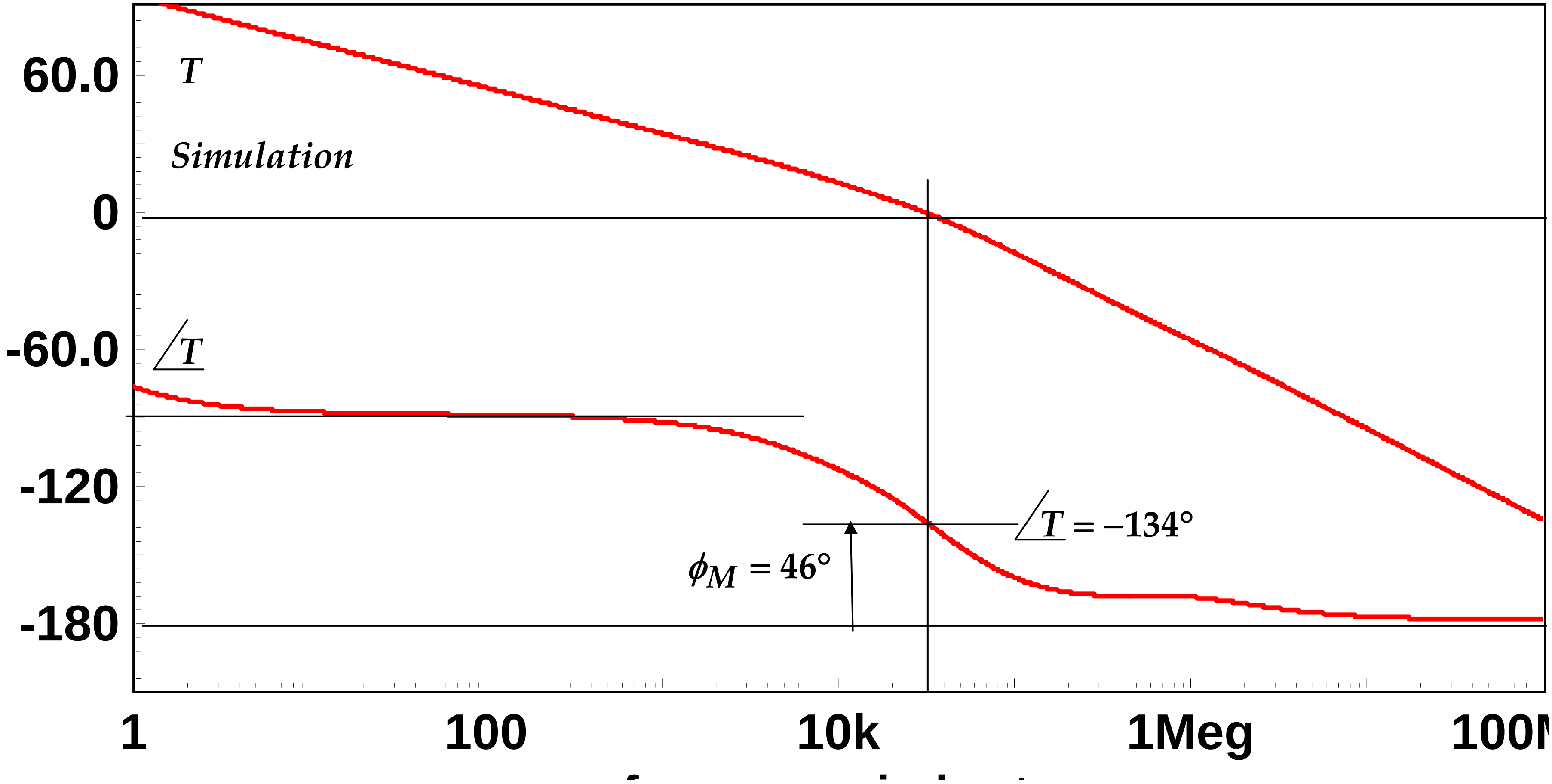
Since three poles close together determine the crossover frequency f_c, there is no point in trying to predict its value. Instead, the simulation shows it to be $f_c = 28kHz$.
From the predicted corner frequencies of T, $\angle T$ can then be calculated as:

$$\angle T = -90° + \tan^{-1}\frac{28}{10.6} - \left(180° + \tan^{-1}\frac{\frac{1}{0.4}\frac{28}{16.5}}{1-\left(\frac{28}{16.5}\right)^2}\right) + \tan^{-1}\frac{28}{792} - \tan^{-1}\frac{28}{1060}$$

$$= -90° + 70° - (180° - 66°) + 2° - 2°$$

$$= -134°$$

This agrees with the simulation:



STEP 5: Closed-loop gain G

The closed-loop gain $G = G_\infty D$ where $G_\infty = \dfrac{R_1 + R_2}{R_1} = 2 = 6.02dB$ is the ideal closed-loop gain, and $D = \dfrac{T}{1+T}$ is the discrepancy factor (Ch. 10).

Since D is a unique function of T, D can be evaluated at the loop gain crossover frequency in terms of the phase margin φ_M.

In polar form,

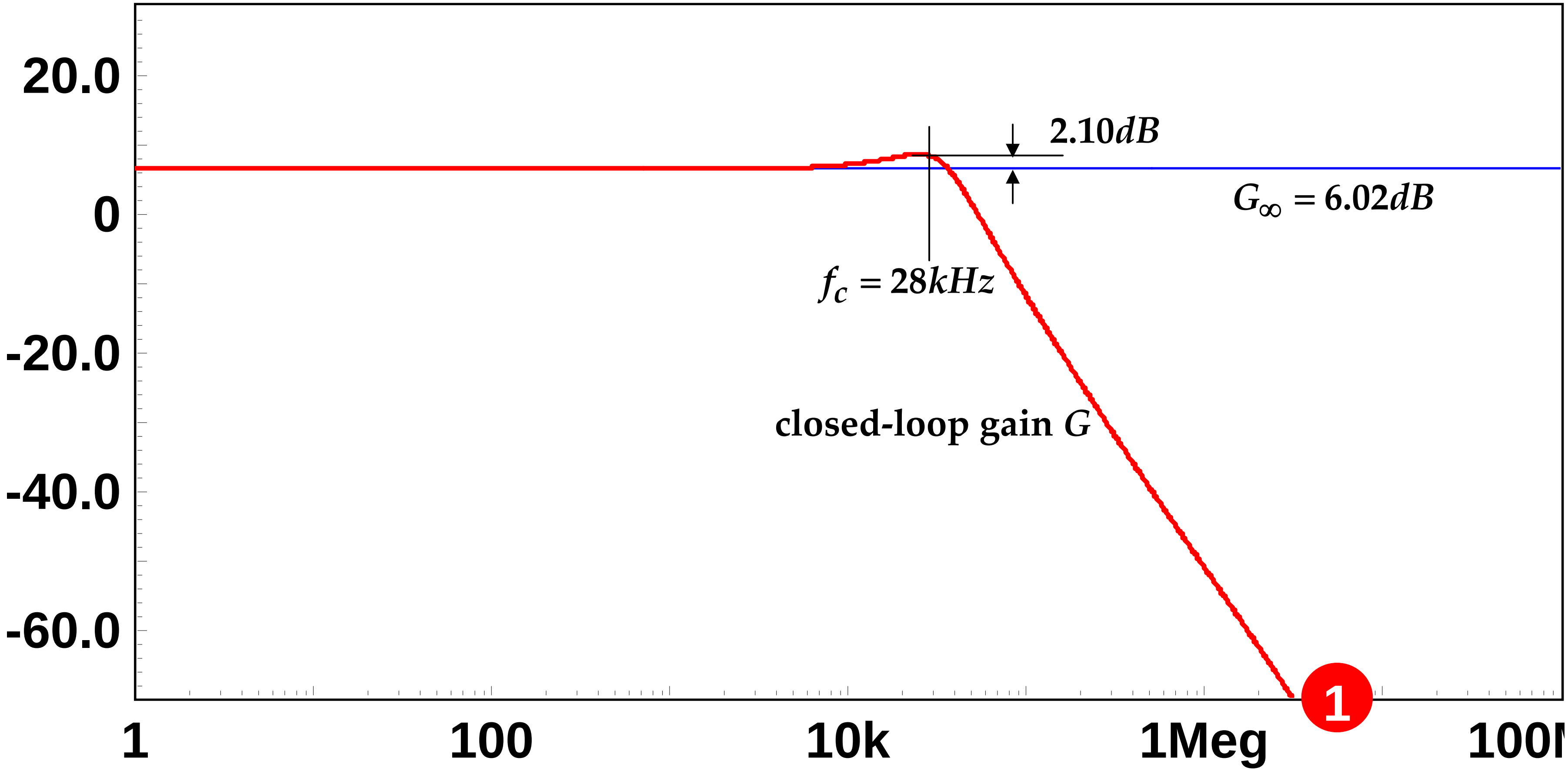
$$T = |T| e^{j\angle T}, \quad |D| = \left| \frac{T}{1+T} \right| = \left| \frac{1}{1+\frac{1}{T}} \right| = \left| \frac{1}{1+\frac{1}{|T|} e^{-j\angle T}} \right|$$

If the phase margin is φ_M, then $-\angle T = (\pi - \varphi_M)$ at the crossover frequency where $|T| = 1$. Substitute in $|D|$:

$$|D|_{f_c} = \left| \frac{1}{1+e^{j(\pi-\varphi_M)}} \right| = \left| \frac{1}{1+e^{-j\varphi_M}} \right| = \left| \frac{1}{1-(\cos\varphi_M - j\sin\varphi_M)} \right| = \frac{1}{\sqrt{(1-\cos\varphi_M)^2 + \sin^2\varphi_M}}$$

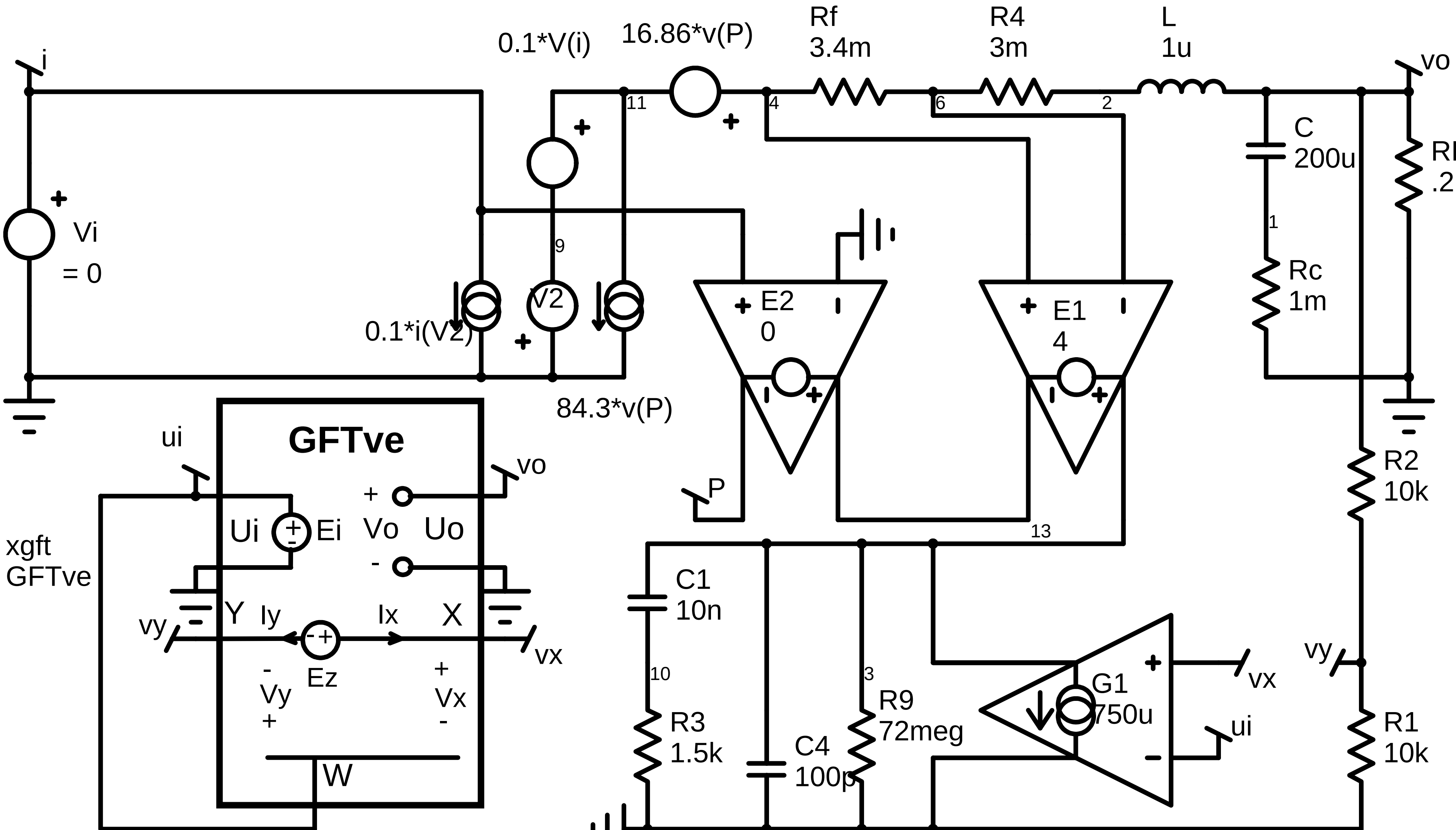
$$= \frac{1}{\sqrt{2(1-\cos\varphi_M)}} = \frac{1}{2\sin\frac{\varphi_M}{2}}$$

In the LM3495, $G_\infty = 2 = 6.02dB$ and $\varphi_M = 46°$ at the crossover frequency $f_c = 28kHz$. Hence, $|D|_{28kHz} = 1.28 \Rightarrow 2.10dB$. These results are in agreement with the simulation using the Intusoft ICAP/4 GFT Template

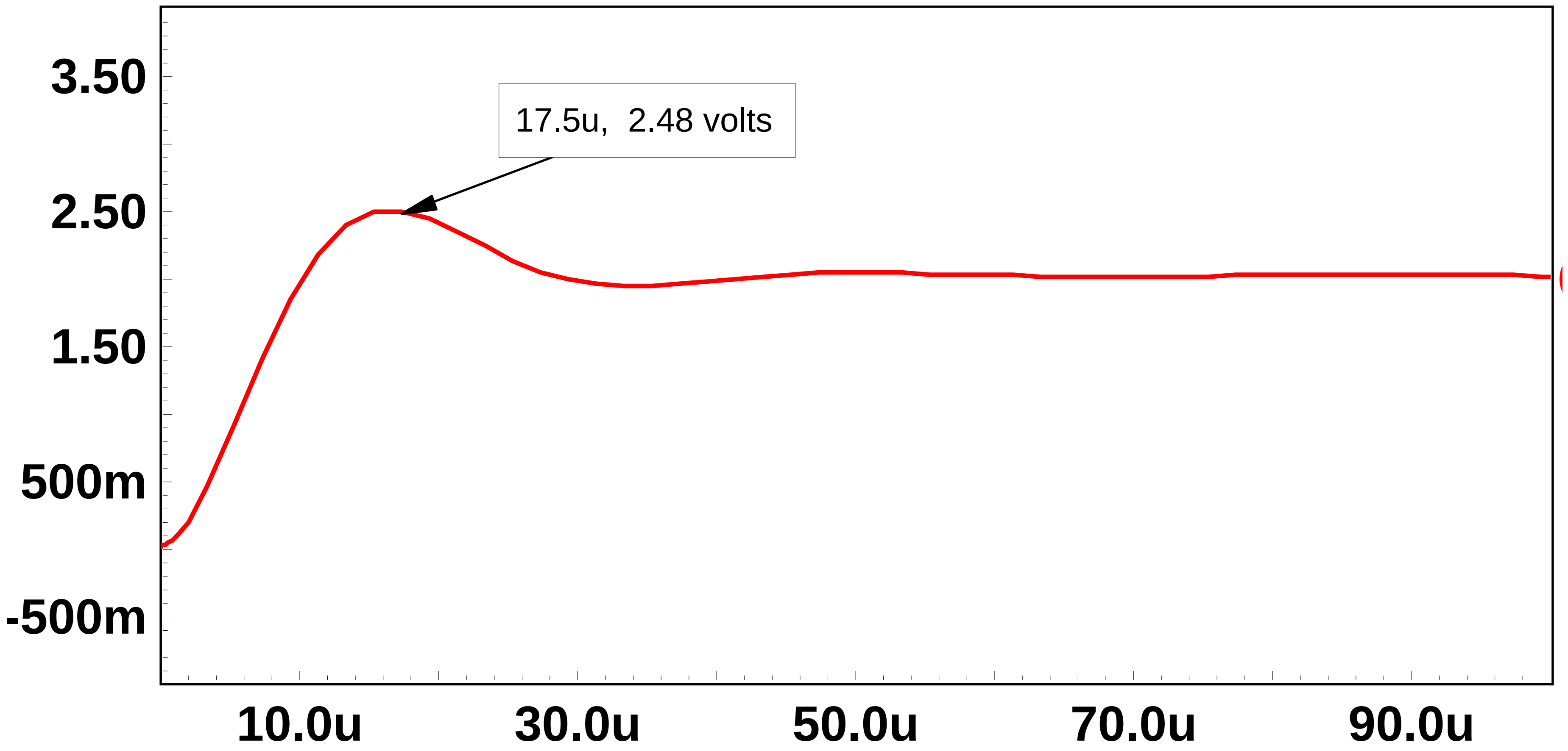

20.0
0
-20.0
-40.0
-60.0
1
100
10k
1Meg
100M
2.10dB
$G_\infty = 6.02dB$
$f_c = 28kHz$
closed-loop gain G
1

Model for simulation using the Intusoft ICAP/4 GFT Template.

(No input filter)

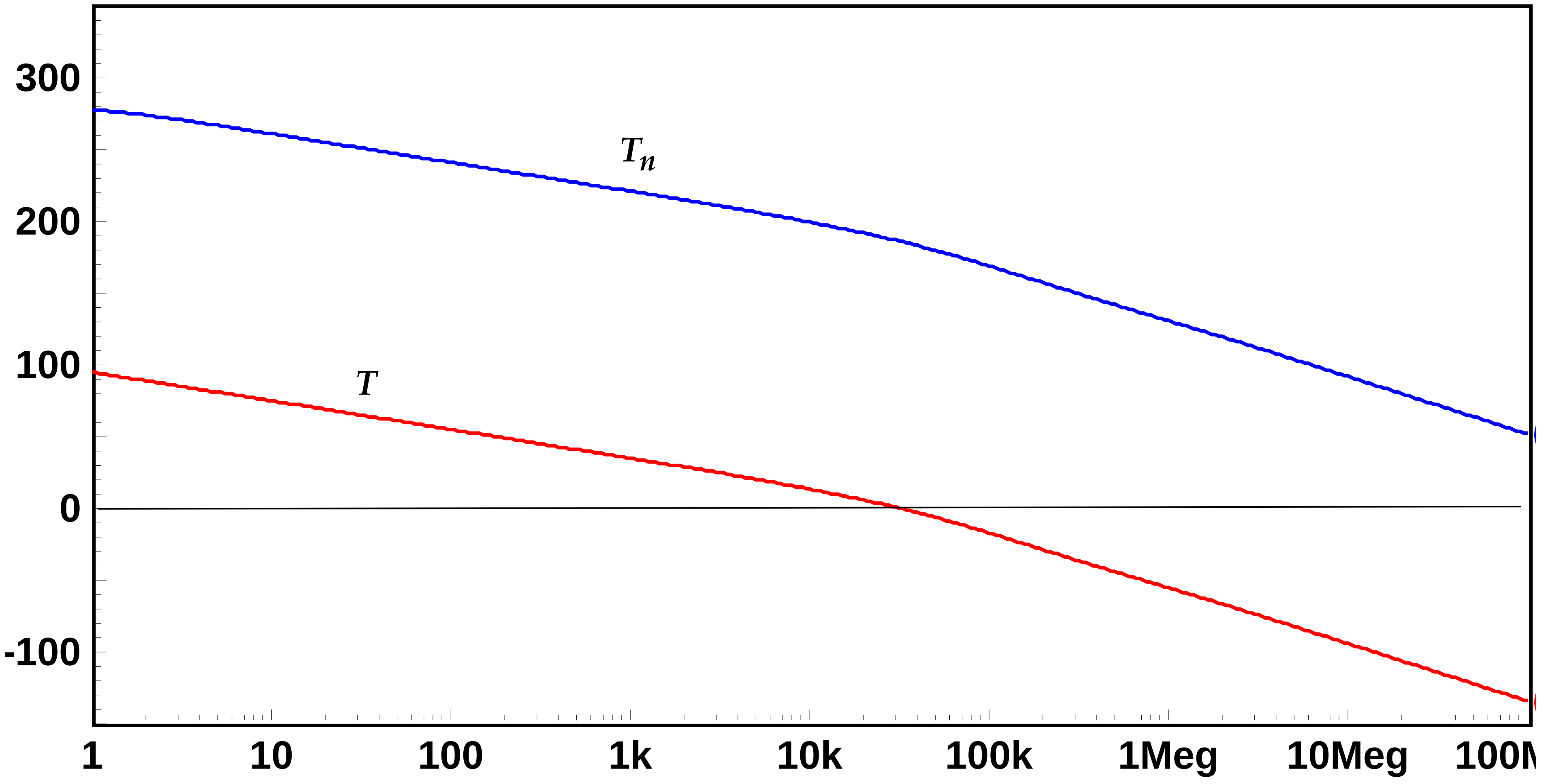


Output voltage response to a 1v step in reference voltage



STEP 6: Check for nonidealities

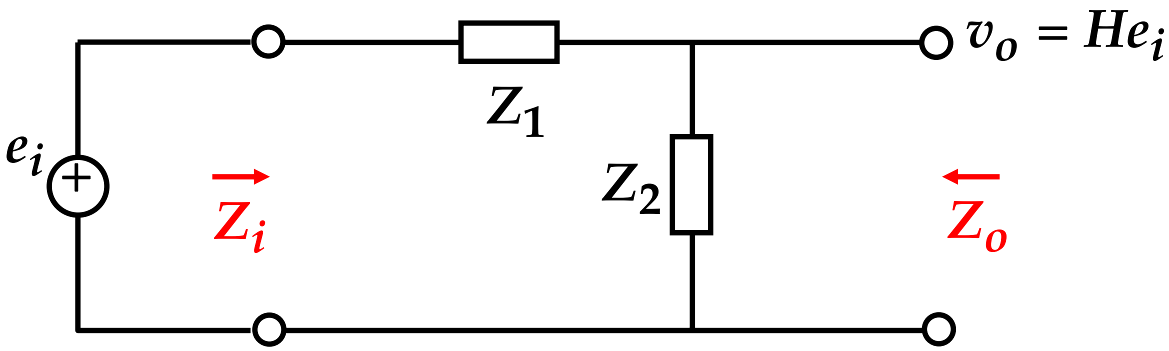
The ICAP/4 simulation also delivers the null loop gain T_n :



Since $T_n >> 1$ at all frequencies of interest, the nonidealities are negligible.

STEP 7: Output Impedance

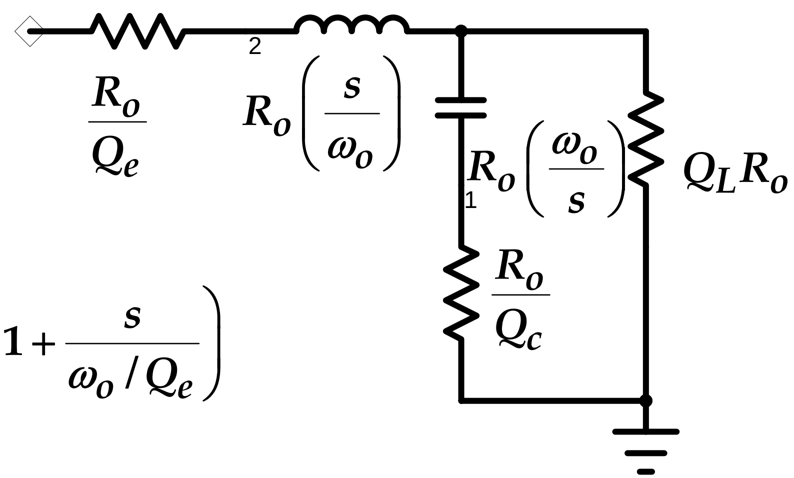
As seen in Ch. 7, for a ladder network such as



the output impedance is

$$Z_o = Z_1 H$$

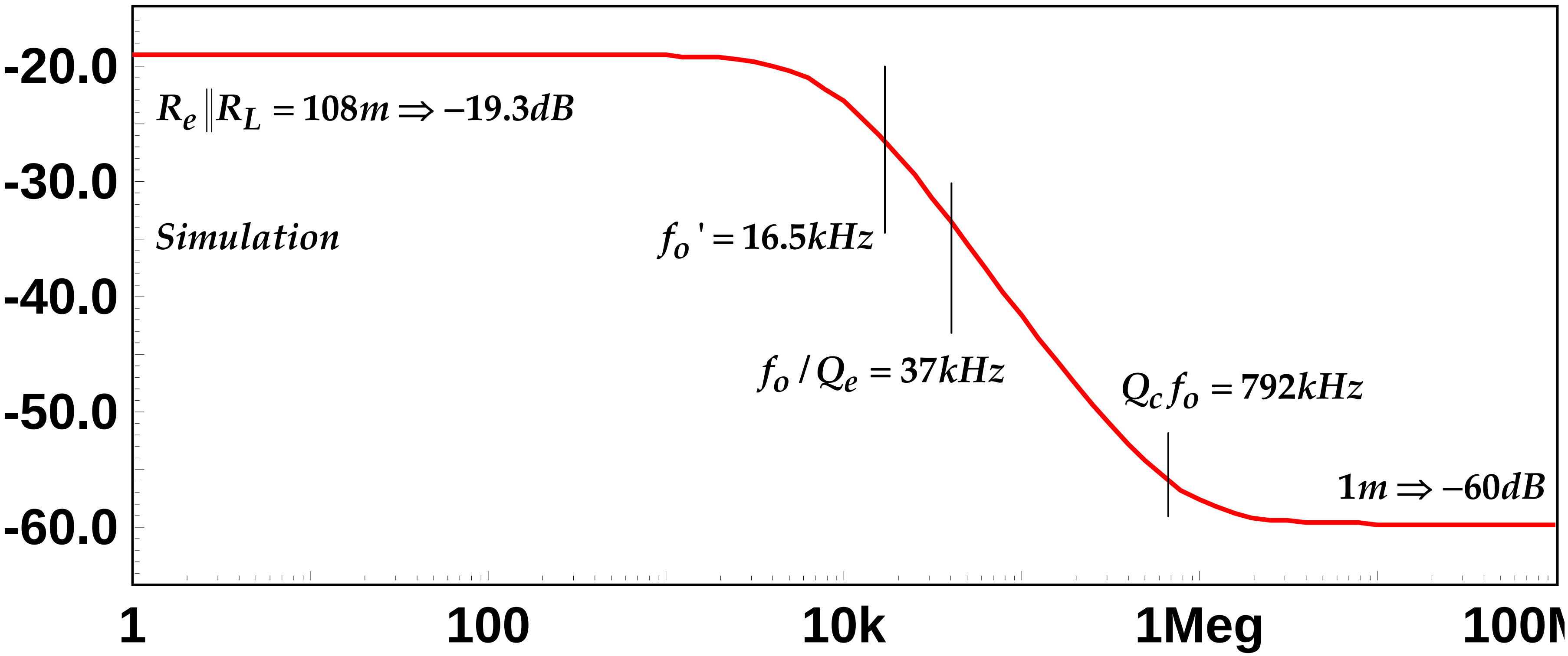
Here,



$$Z_1 = R_o \frac{1}{Q_e}\left(1 + \frac{s}{\omega_o / Q_e}\right)$$

and H is already known, so

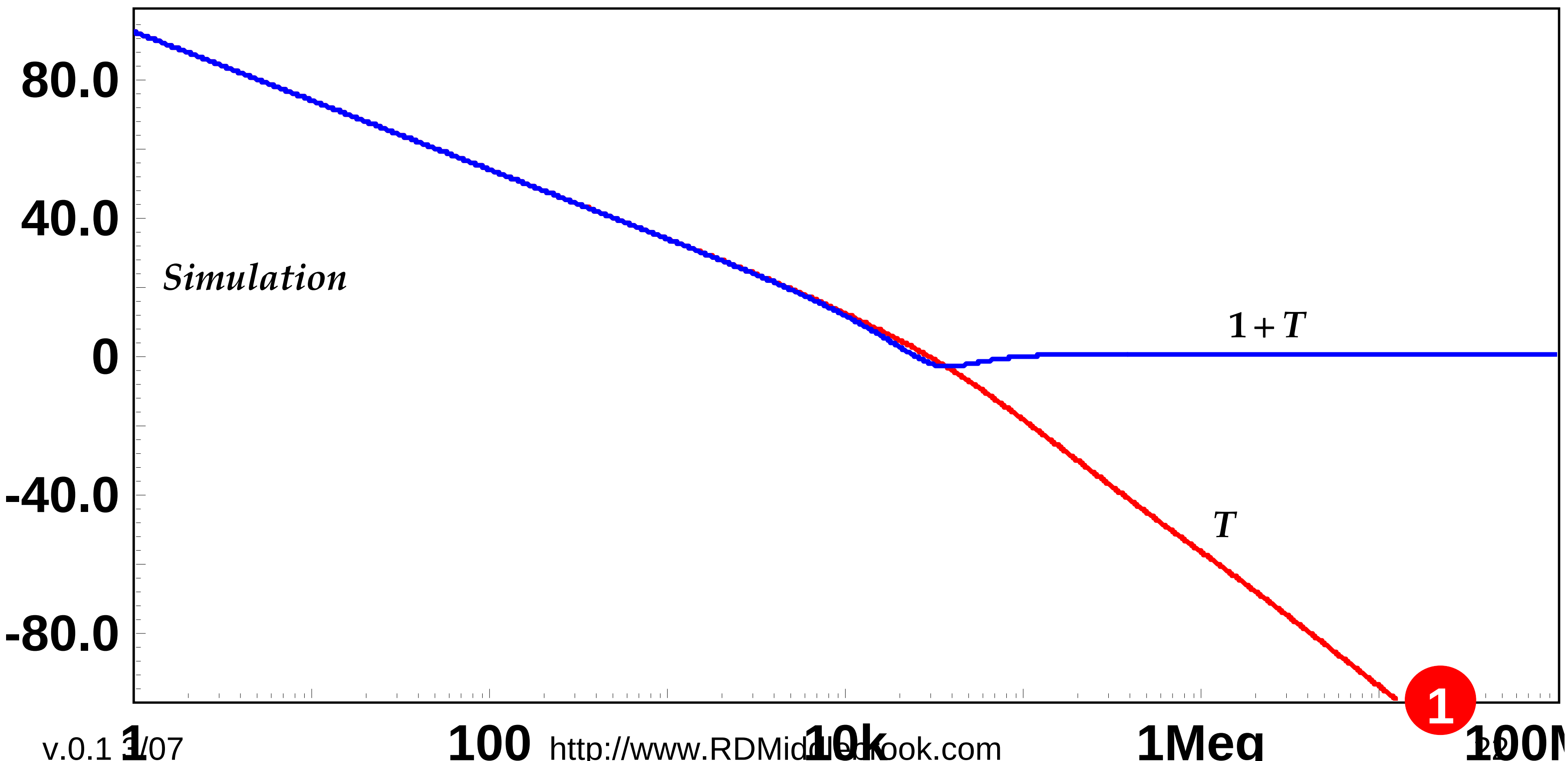
$$Z_o = \left(R_e \| R_L\right) \frac{\left(1+\frac{s}{\omega_o / Q_e}\right)\left(1+\frac{s}{Q_c \omega_o}\right)}{1+\frac{1}{1+\frac{1}{Q_e Q_L}} \frac{1}{Q_e \| Q_L}\left(\frac{s}{\omega_o}\right)+\frac{1}{1+\frac{1}{Q_e Q_L}}\left(\frac{s}{\omega_o}\right)^2}$$

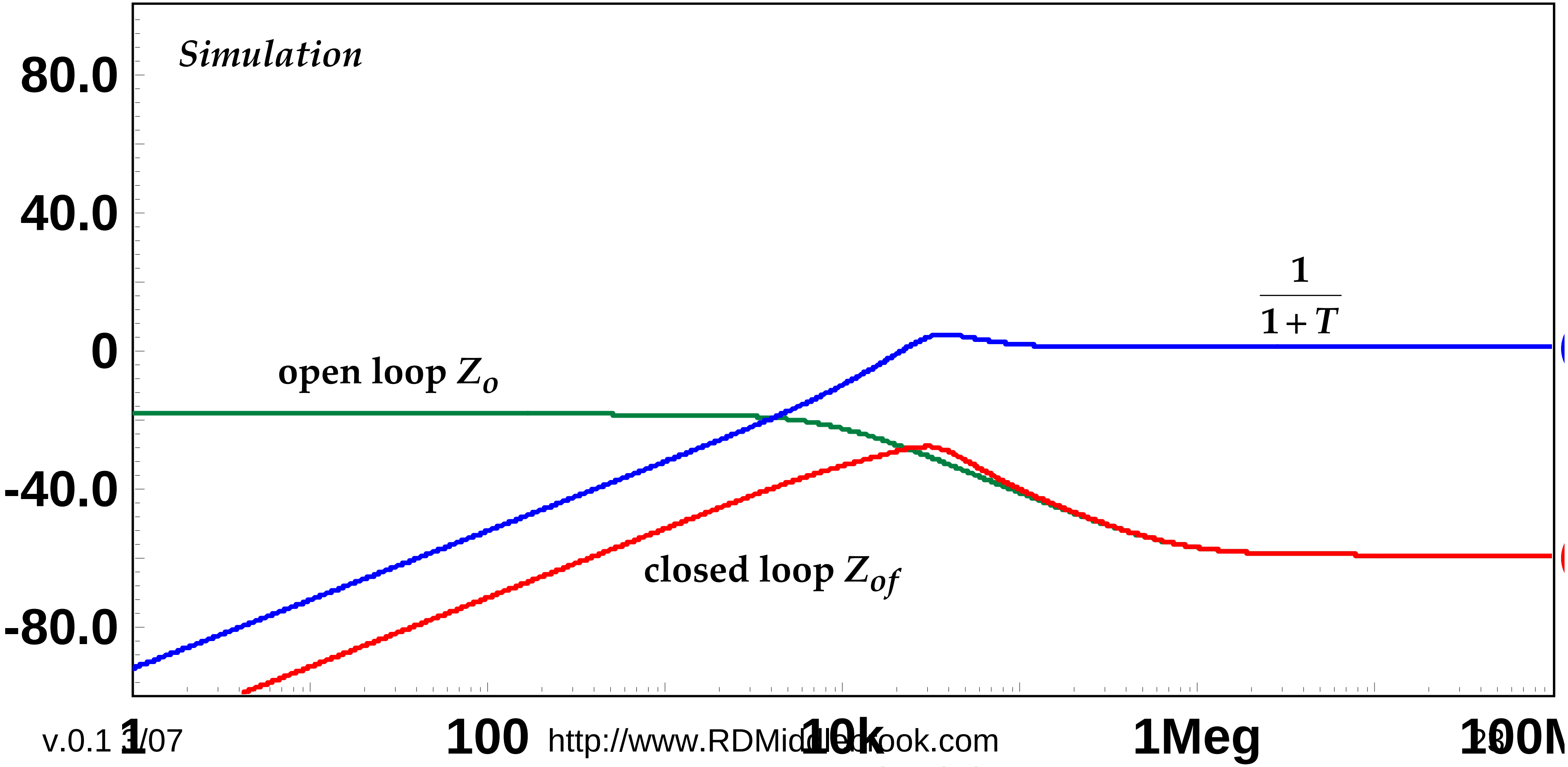


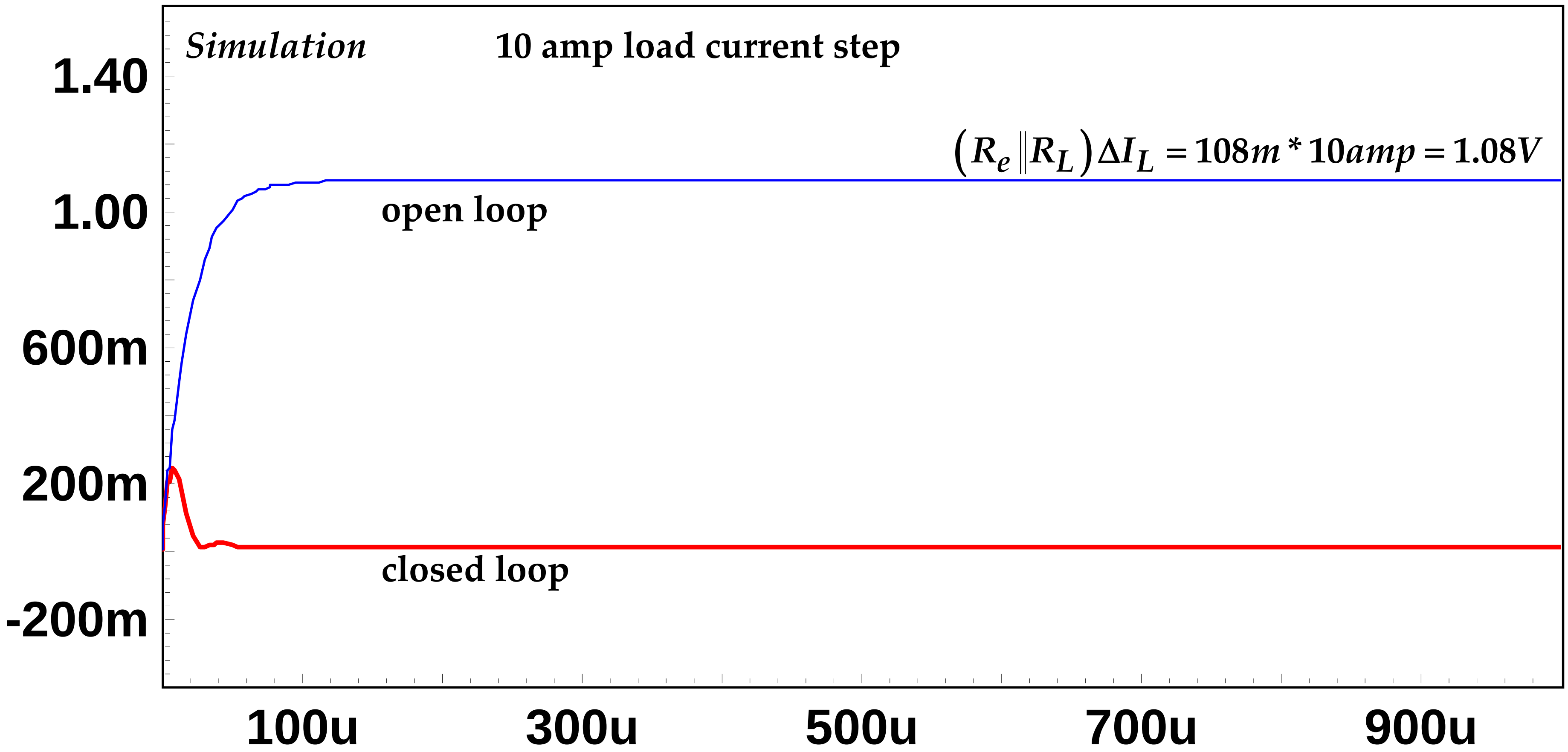
Closed loop output impedance Z_{of}

$$Z_{of} = \frac{Z_o}{1+T}$$

Since T is already known, $1+T$ and $\dfrac{1}{1+T}$ can be found by the methods of Ch. 6.

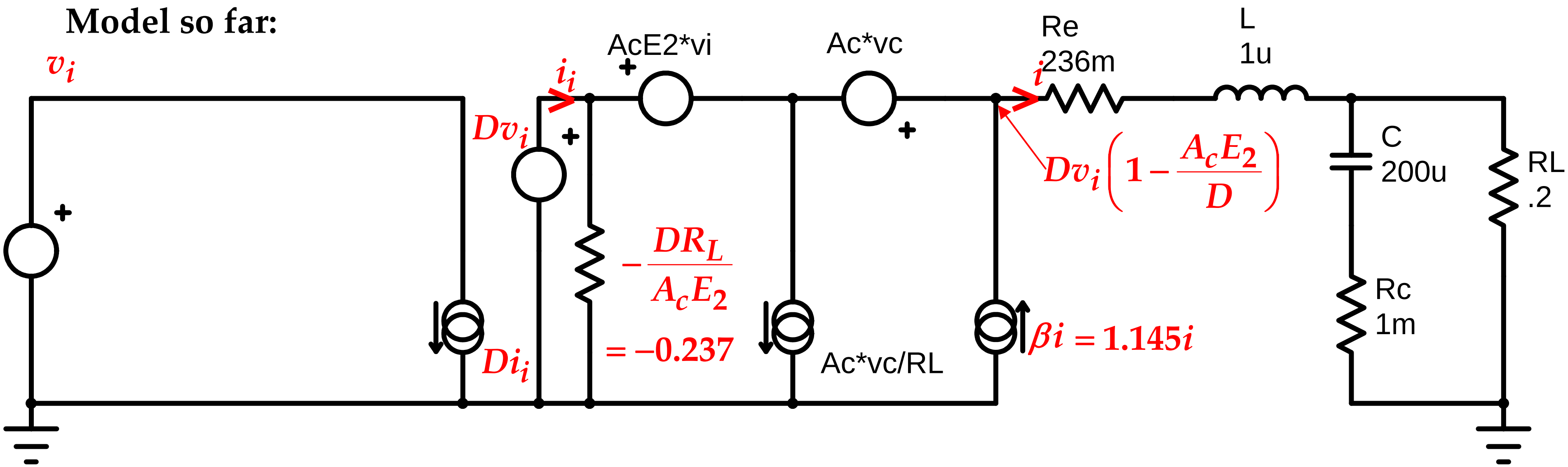



Simulation
80.0
40.0
0
-40.0
-80.0
open loop Z_o
closed loop Z_{of}
$\frac{1}{1+T}$
1
100
10k
1Meg
100M

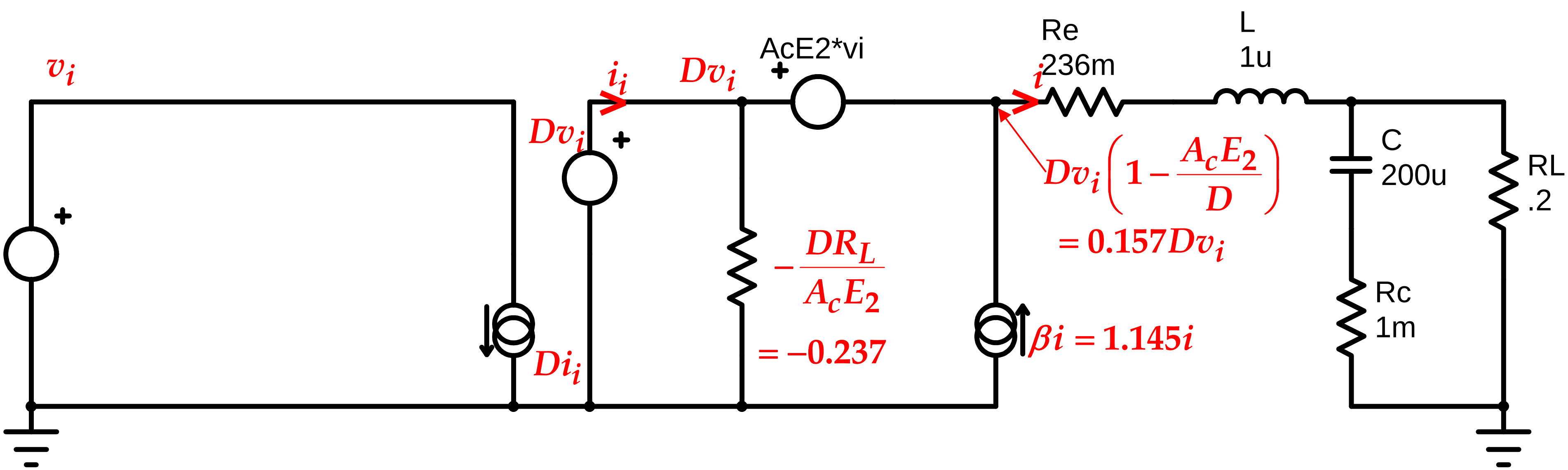

Simulation
10 amp load current step
$\left(R_e \| R_L\right) \Delta I_L = 108m * 10amp = 1.08V$
open loop
closed loop
1.40
1.00
600m
200m
-200m
100u
300u
500u
700u
900u

STEP 8: Line-to output TF

Model so far:



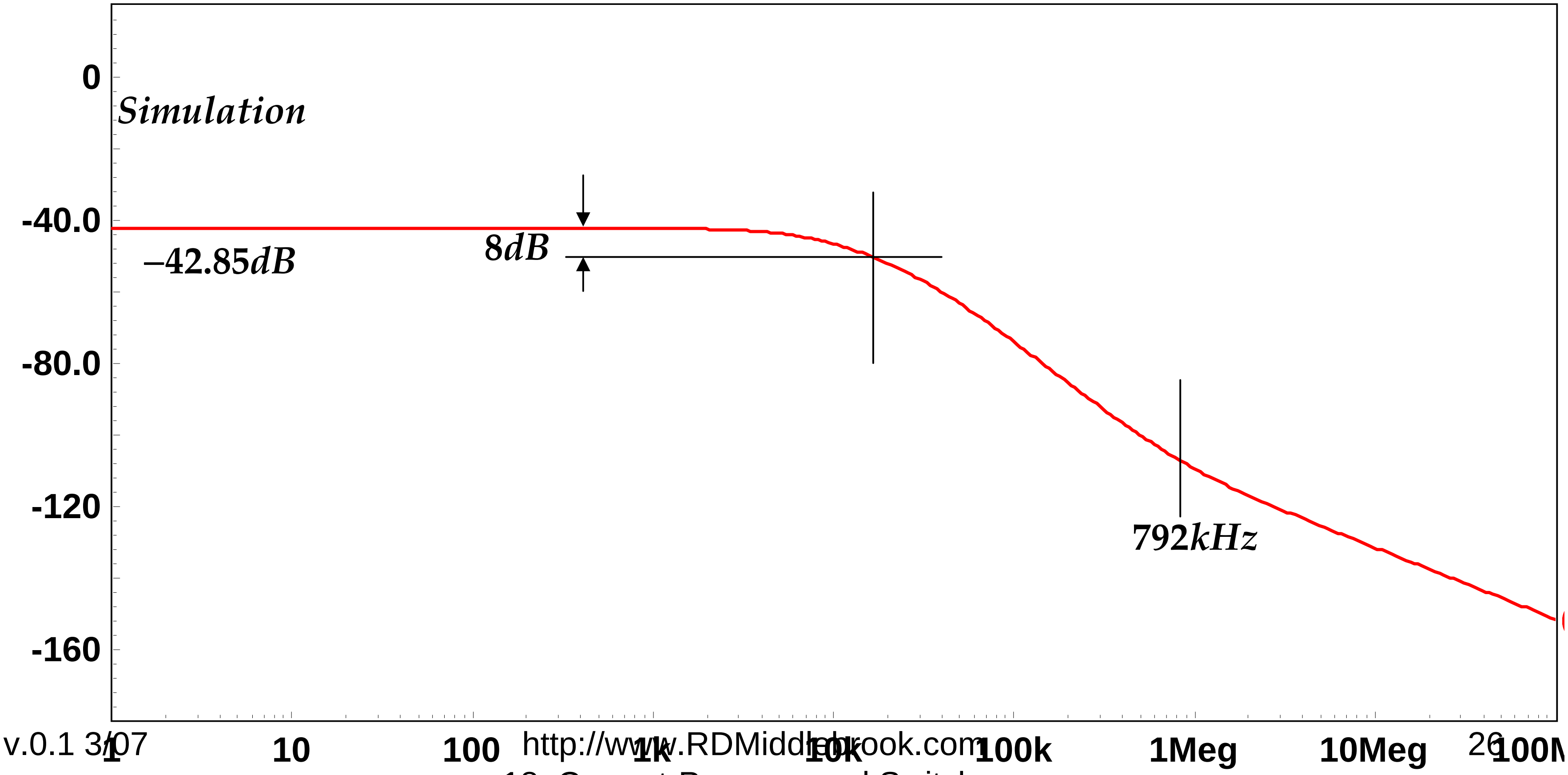
Since $v_c = 0$, the voltage and current modulation generators are zero:



The power stage line-to-output voltage gain is $v_o / v_i \Big|_{v_c=0} = Dv_i \left(1 - \frac{A_c E_2}{D}\right) H = 0.157 DHv_i$

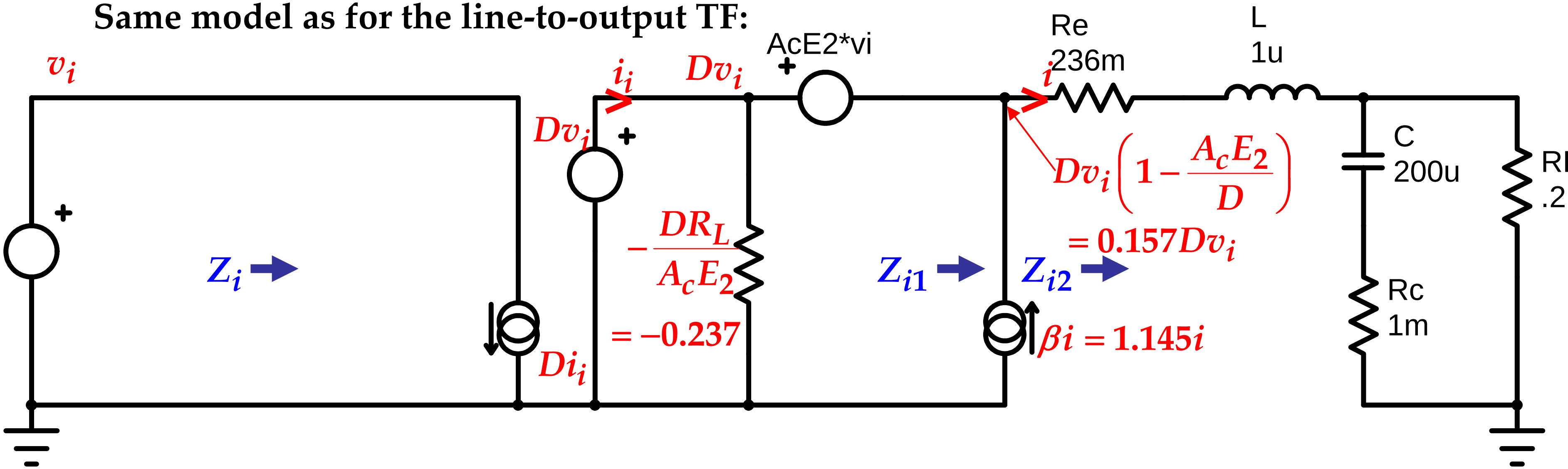
Insertion of numbers gives

$$v_o / v_i \Big|_{v_c=0} = -42.85dB \frac{1 + \frac{s/2\pi}{792kHz}}{1 + \frac{1}{0.400}\left(\frac{s/2\pi}{16.5kHz}\right) + \left(\frac{s/2\pi}{16.5kHz}\right)^2}$$



STEP 9: Line input impedance TF Z_i

Same model as for the line-to-output TF:



$$Z_i = \frac{1}{D^2}\left[-\frac{DR_L}{A_cE_2}\middle\|\frac{Z_{i1}}{1-\frac{A_cE_2}{D}}\right]$$

where

$$Z_{i1} = \frac{1}{1-\beta}Z_{i2}$$

in which

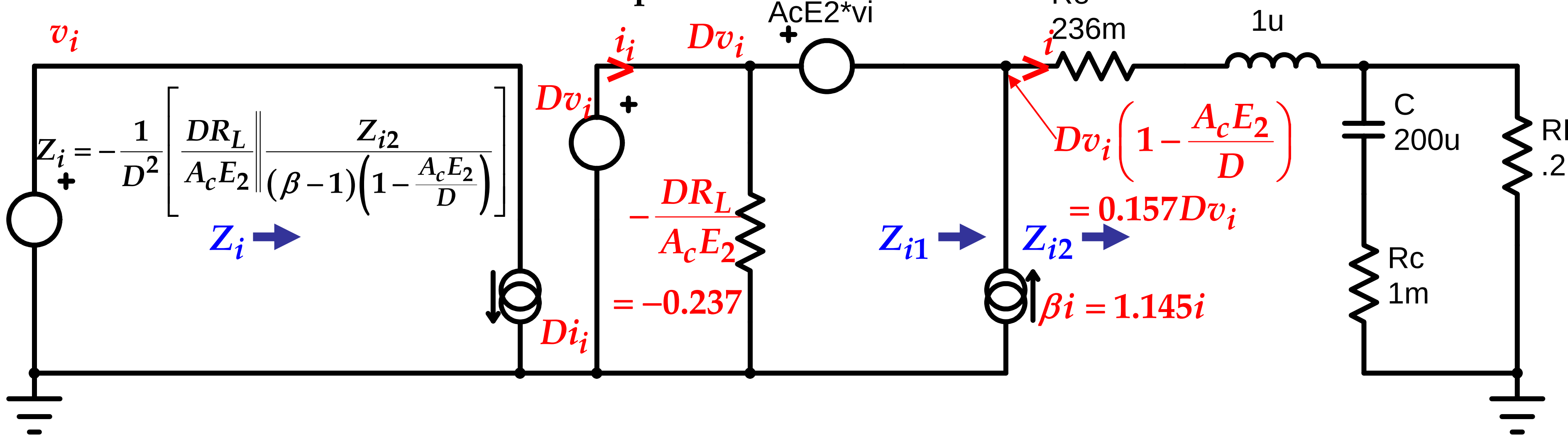
Z_{i2} = input impedance of the filter, which is already known.

Since β>1, the result can be written

$$Z_i = -\frac{1}{D^2}\left[\frac{DR_L}{A_cE_2}\middle\|\frac{Z_{i2}}{(\beta-1)\left(1-\frac{A_cE_2}{D}\right)}\right]$$

STEP 9: Line input impedance TF Z_i

Same model as for the line-to-output TF:



Since $Z_{i2}(0) = R_e + R_L = 0.236 + 0.2 = 0.436$,

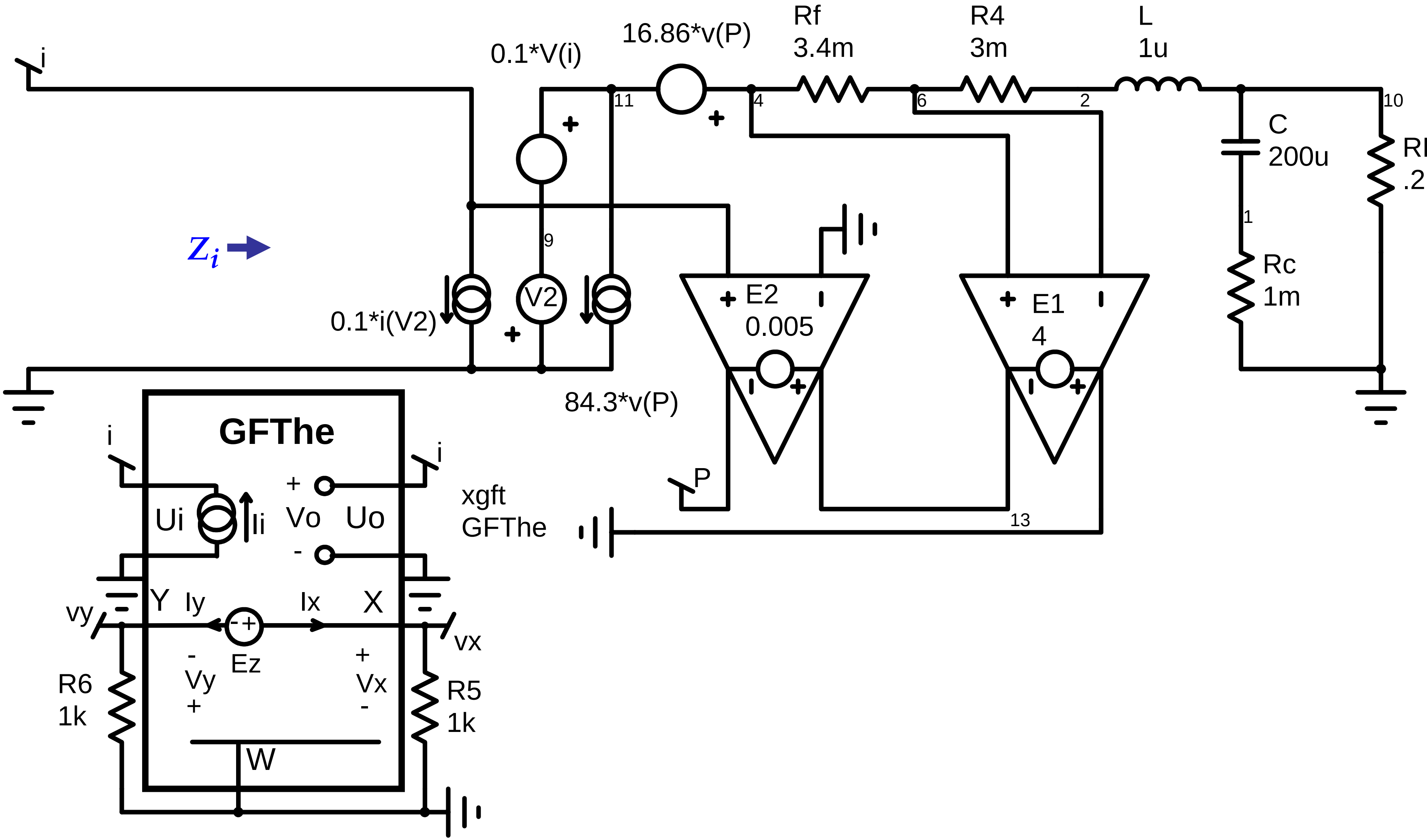
$$Z_i(0) = -\frac{1}{0.1^2}\left[0.237 \middle\| \frac{0.436}{(1.146-1)0.157}\right] = -100[0.237 \| 19] = -100[0.234] = -23.4 \Rightarrow 27.38dB$$

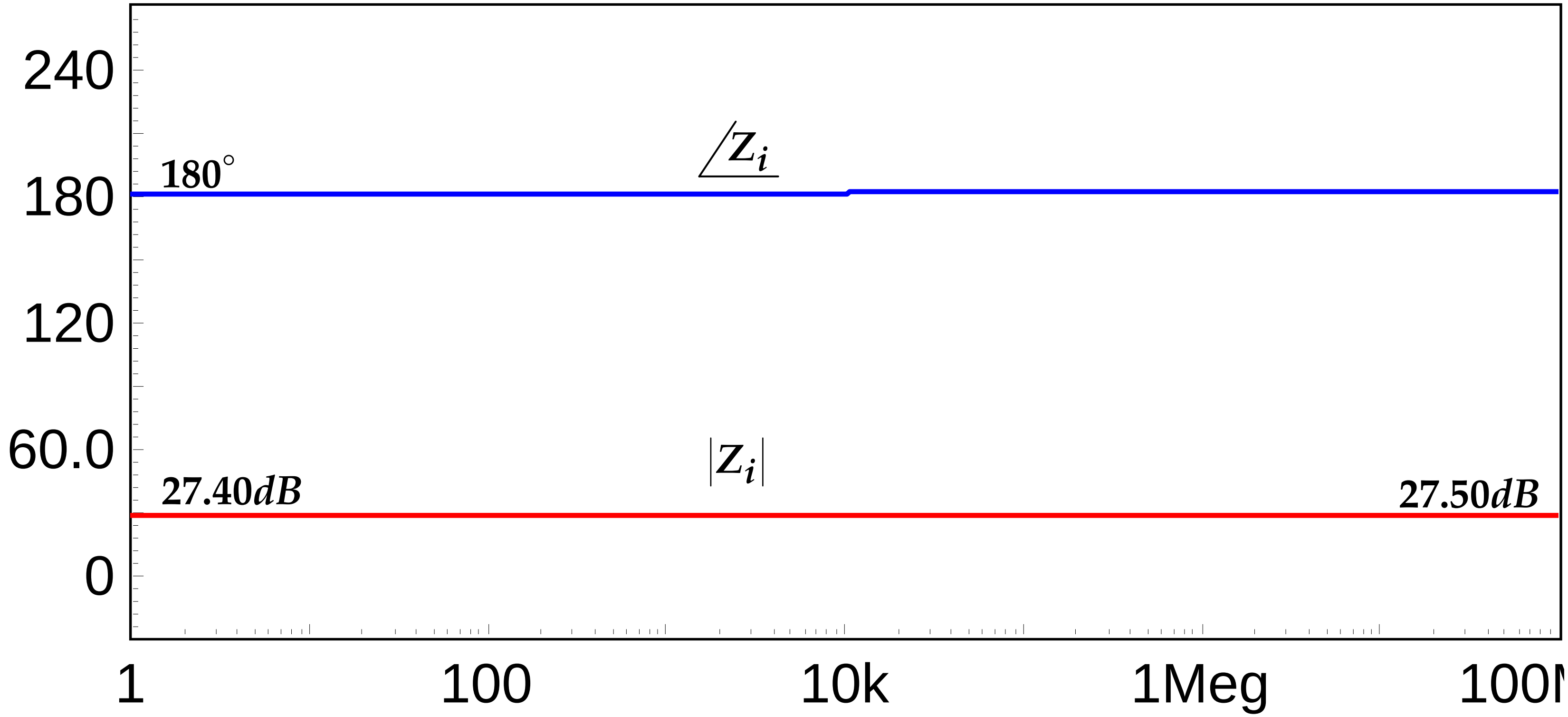
Since $Z_{i2}(\infty) = \infty$,

$$Z_i(\infty) = -\frac{1}{0.1^2}[0.237 \| \infty] = -100[0.237] = -23.7 \Rightarrow 27.49dB$$

Conclusion: the current-programming loop makes the filter input impedance look high, and the (negative) line input impedance is dominated by the term due to the line feedforward.

Input impedance Z_i by the ICAP/4 GFT Template:





The result is in agreement with that predicted by the modified model.