STEP_UP_5_9VDC_12V_3A_CM_CCM

Specifications

$V_{in,min}$ = 5V

$V_{in,max}$ = 9 V

V_{out} = 12 V

V_{ripple} = ΔV = 50 mV

V_{out} drop = 0.5 V max from I_{out} = 0.5 A to 3 A in 1us

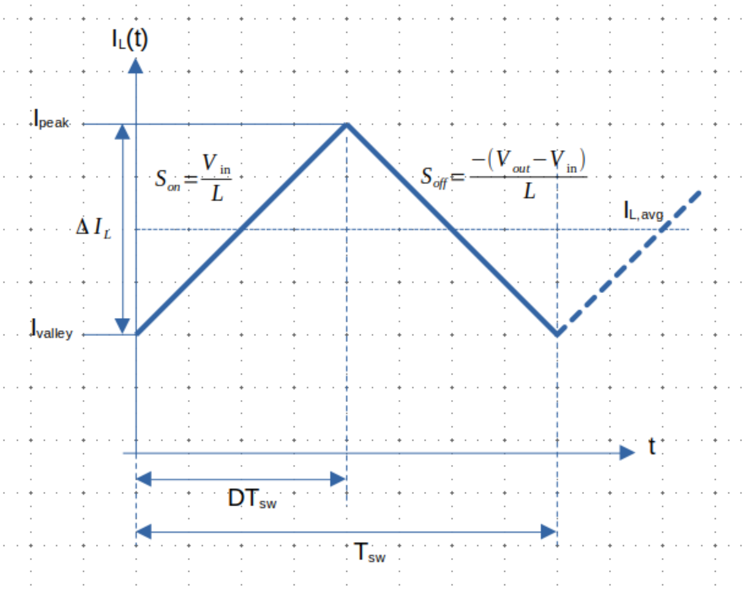
$I_{out,max}$ = 3 A

f_{sw} = 300 kHz

$I_{ripple,peak}$ = 1 mA, input current maximum ripple

CM CCM

Design



(boost inductor current waveform – needed to derive the inductor values)

$$I_{peak} = I_{\text{in,avg}} + \frac{\Delta I_L}{2}$$

$$\Delta I_L = \frac{V_{\text{in}} D T_{sw}}{L}$$

$$\frac{\Delta I_L}{I_{\text{in,avg}}} = \delta I_r = \frac{V_{\text{in}} D}{I_{\text{in,avg}} F_{sw} L}$$

We know

$$\frac{P_{out}}{\eta}=V_{\text{in}}I_{\text{in,avg}}\rightarrow I_{\text{in,avg}}=\frac{P_{out}}{\eta V_{\text{in}}}$$

then

$$\frac{\Delta I_L}{I_{\text{in,avg}}}=\delta I_r=\frac{\eta V_{\text{in}}^2 D}{P_{out}F_{sw}L}$$

$$L=\frac{\eta V_{\text{in}}^2 D}{P_{out}F_{sw}\delta I_r}$$

$$D_{min}=\frac{V_{out}-V_{\text{in,max}}}{V_{out}}=\frac{12-9}{12}=0.25$$

$$D_{max}=\frac{V_{out}-V_{\text{in,min}}}{V_{out}}=\frac{12-5}{12}=0.583$$ (note that we will need slope compensation)

$$L=\frac{\eta V_{\text{in}}^2 D}{P_{out}F_{sw}\delta I_r}=\frac{0.9\cdot 5^2\cdot 0.583}{36\cdot 300\text{e}3\cdot 0.3}=4\,uH$$

$$I_{peak,L}=\frac{P_{out}}{\eta V_{\text{in,min}}}+\frac{V_{\text{in,min}}D_{max}}{2LF_{sw}}=\frac{36}{0.9\cdot 5}+\frac{5\cdot 0.583}{2\cdot 4\text{e-}6\cdot 300\text{e}3}=9.21\,A$$

$$R_{critical}=\frac{2LF_{sw}}{D_{min}(1-D_{min})^2}=\frac{8\text{e-}6\cdot 300\text{e}3}{0.25(1-0.25)^2}=17\,\Omega$$

→ this load corresponds to 0.705 A load current, slightly above the design specification of 0.5 A. As we switch from DCM to CCM the transient response will degrade.

→CAPACITOR SELECTION

$$C_{out}\geqslant\frac{D_{min}V_{out}}{F_{sw}R_{load}\Delta V}\geqslant\frac{0.25\cdot 12}{300\text{e}3\cdot 4\cdot 0.05}\geqslant 50\,uF$$

→ the rms current value represents the final capacitor criterion for its selection

$$I_{\text{Cout,rms}}=I_{out}\sqrt{\frac{D}{D`}+\frac{D^2D`}{12}\left(\frac{D`}{\tau_L}\right)^2}\ where\ \tau_L=\frac{L}{R_{load}T_{sw}}$$

$$\tau_L=\frac{4\text{e-}6\cdot 300\text{e}3}{4}=300\text{e-}3$$

$$I_{\text{Cout,rms}}=I_{out}\sqrt{\frac{D}{D`}+\frac{D^2D`}{12}\left(\frac{D`}{\tau_L}\right)^2}=3\sqrt{\frac{0.58}{0.42}+\frac{0.58^2\cdot 0.42}{12}\left(\frac{0.42}{300\text{e-}3}\right)^2}=3.55\,ARMS$$

→ if we select 3 x 22 uF, the required crossover frequency to fulfill the step load requirement will be:

$$f_c \approx \frac{\Delta I_{out}}{2\pi \Delta V_{out} C_{out}} = \frac{2.5}{6.28 \cdot 0.5 \cdot 66\text{e-}6} = 12\text{e}3$$

-RHPz (this zero limits the selection of the crossover frequency)

$$f_{z2} = \frac{R_{load} D^{`2}}{2\pi L} = \frac{4 \cdot (1-0.58)^2}{6.28 \cdot 4\text{e-}6} = 28\text{e}3\, Hz$$

→ Allowing for some margin we will select a crossover frequency of 6 kHz, below the RHPz. With this new fc, let us compute the needed output capacitance:

$$C_{out} \approx \frac{\Delta I_{out}}{2\pi \Delta V_{out} f_c} = \frac{2.5}{6.28 \cdot 0.5 \cdot 6\text{e}3} = 132\, uF$$

→ we can select 2 x 68 uF

→MOSFET

$V_{DS,Max} = V_{out} = 12$ V

$$I_{\text{sw,rms}} = I_{out} \sqrt{\frac{D}{(1-D)^2} + \frac{1}{3}\left(\frac{1}{2\,\tau_L}\right)^2 D^3 (1-D)^2} = 4\sqrt{\frac{0.583}{(1-.583)^2} + \frac{1}{3}\left(\frac{1}{2 \cdot 300\text{e-}3}\right)^2 0.583^3 (1-0.583)^2} = 7.3\, A$$

→ DIODE

PIV = V_{out} = 12 V

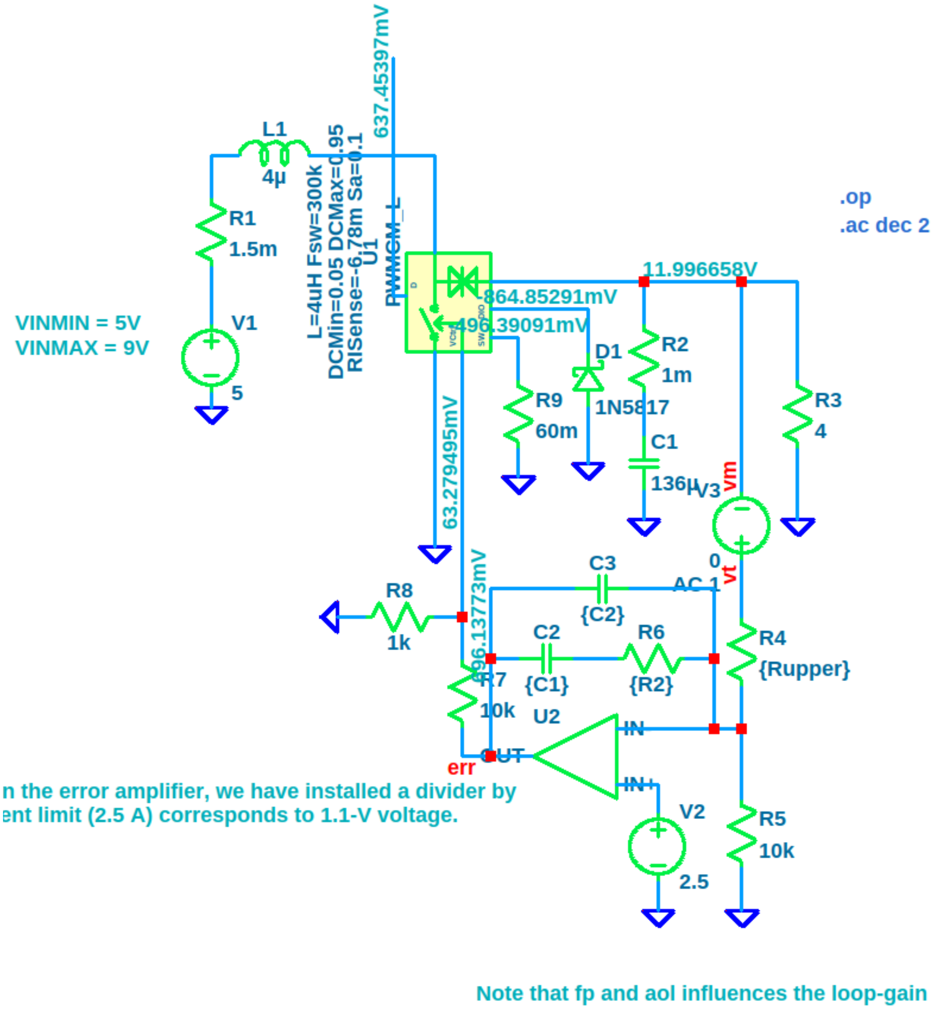
$$I_{\text{d,rms}} = I_{out} \sqrt{\frac{1}{(1-D)} + \frac{1}{3}\left(\frac{1}{2\,\tau_L}\right)^2 D^2 (1-D)^3} = 4\sqrt{\frac{1}{(1-.25)} + \frac{1}{3}\left(\frac{1}{2 \cdot 300\text{e-}3}\right)^2 0.25^2 (1-0.25)^3} = 4.6\, A$$

AC ANALYSIS (using specifications from LM5121)

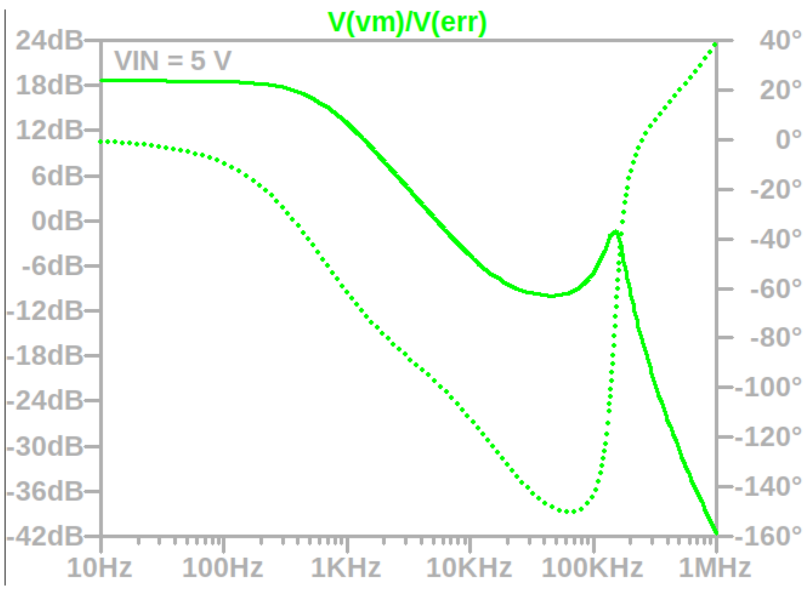
$$R_S = \frac{75\text{e-}3}{9.21 \cdot 1.2} = 6.78\, m\Omega,\ accounting\ for\ 20\,\%\ margin$$

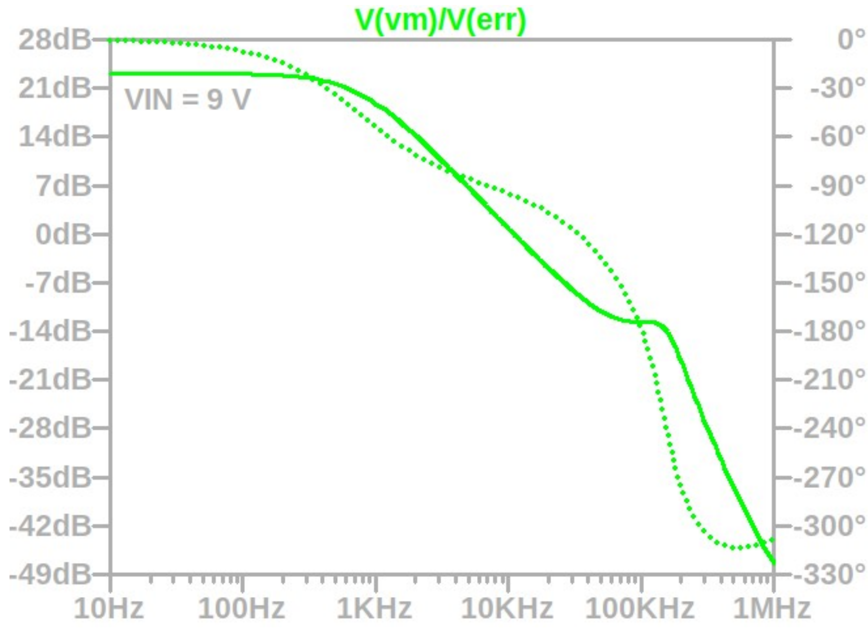
(to provide enough swing on the error amplifier, we have to install a divider, such that the upper current limit 9.21 A, corresponds to a higher voltage)

→ using a lossy model, featuring MOSFET conduction losses (60m), and forward drop of the diode, the duty cycle ratio will differ. Here at VIN = 9 V, Dmin = 0.294, at VIN = 5 V we have Dmax = 0.637.



→ open-loop (control-to-output) transfer function (at VIN = 5 VDC and 9 VDC)





The previous figures shows the ac sweep for the open-loop transfer function. The sub-harmonic peaking clearly appears at 150 kHz, so we will need to damp it by slope compensation.

The traditional slope compensation method requires it to be 50 – 70% of the inductor downslope:

$$|S_{off}| = \frac{V_{out} - V_{\text{in}}}{L} = \frac{12-5}{4\text{e-}6} = 1.75\text{e}6\,A/s = 1.75\,A/us$$

As we reflect it over the sense resistor, we have

$$S_{off}` = S_{off} R_{sense} = 1.75 \cdot 6.78\text{e-}3 = 11.86\,\frac{mV}{us}$$

$$S_e = \frac{S_{off}`}{2} = 5.93\,mV/us = 5.93\,kV/s$$

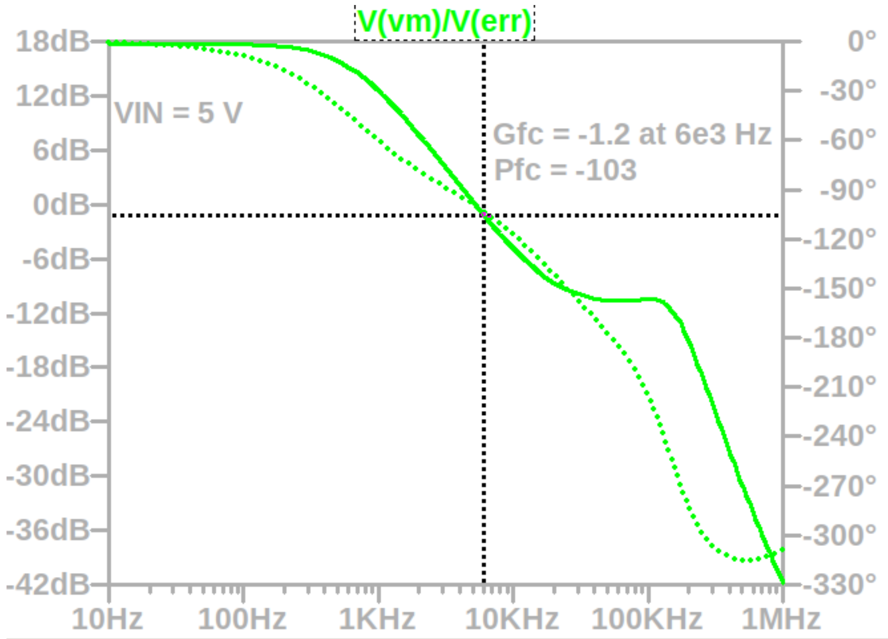
From simulations, a value of Se of 9 kV/s is better suited.

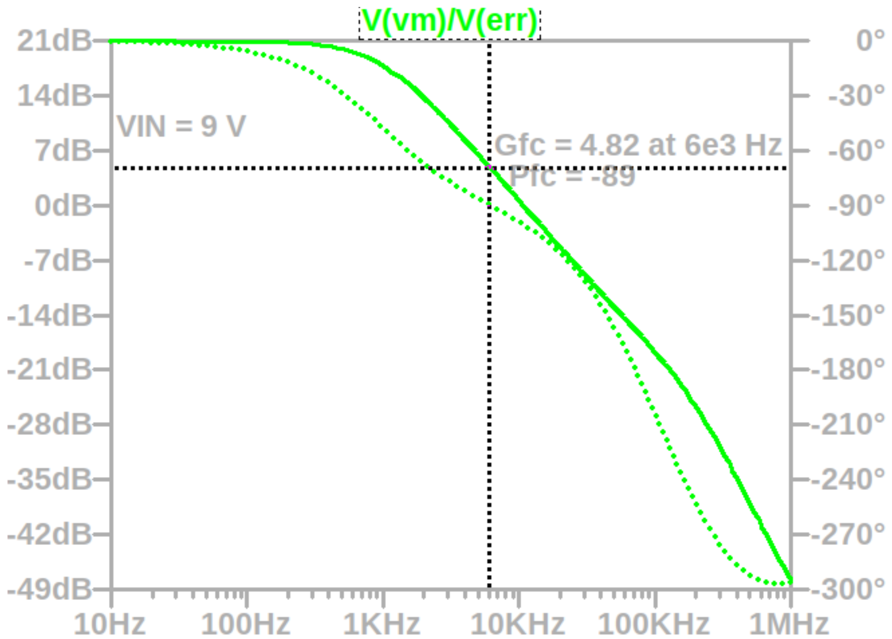
From the CM CCM analysis we know an approximate value of the domninant pole:

$$f_{pd} = \frac{2}{2\pi R_{load} C_{out}} = \frac{2}{6.28 \cdot 4 \cdot 136\text{e-}6} \approx 600\,Hz$$

We can now update the model with this slope compensation and see the final open-loop transfer function.

→ (with slope compensation) open-loop (control-to-output) transfer function (at VIN = 5 VDC and 9 VDC)





We deal with a current mode converter system running in CCM at 3 A. This leads us to implement a type 2 compensation factor. To stabilize the circuit via a type 2 circuit, we use the k factor, which gives adequate results with first-order systems. (If the results does not suit the specifications, we can use the manual placement)

->COMPENSATION

For a 12 V output from a 1.2V reference, we select a 250-uA bridge current. The lower and upper resistor values are therefore:

$$R_{lower} = \frac{1.2}{250\text{e-}6} = 4.8\,k\Omega$$

$$R_{upper} = \frac{12 - 1.2}{250\text{e-}6} = 43.2\,k\Omega$$

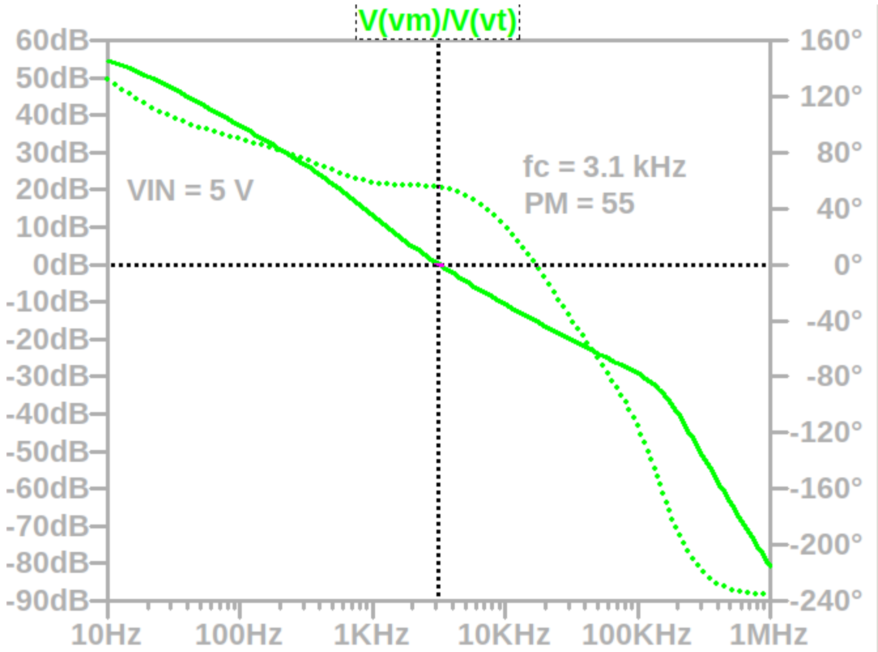
From the open-loop (control-to-output) transfer function, we can see that the deficit of gain at 6e3 Hz worst case is approx 5 dB.

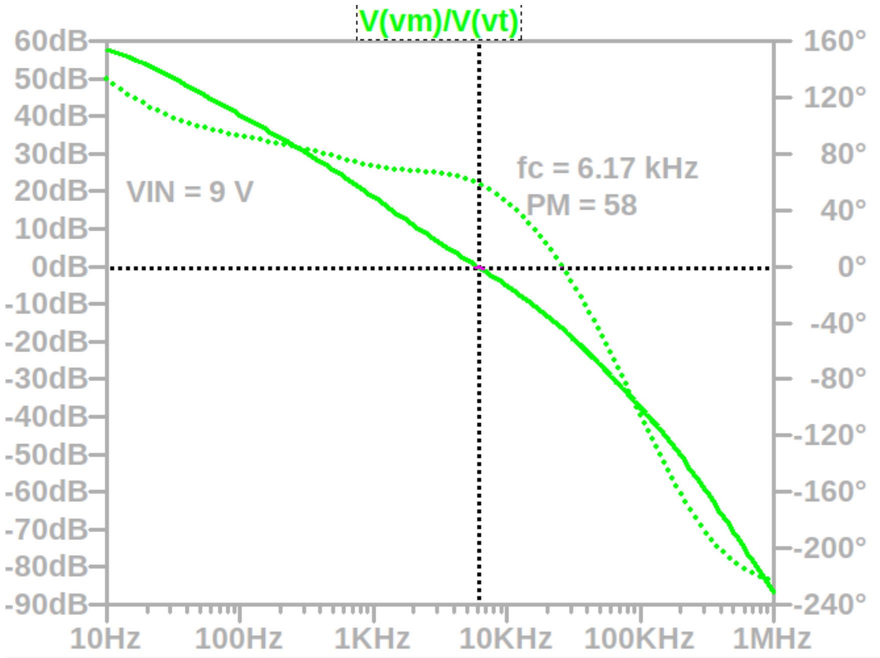
The k factor macro computes a zero placed at 1.6 kHz and a pole at 19.97 kHz (k = 3.6) for a phase margin of 60°. The values passed to the simulator are the following:

C1: 3.63445750386483e-09=3.63446e-09
C2: 3.02813798573986e-10=3.02814e-10
R2: 26317.1886120818=26317.2

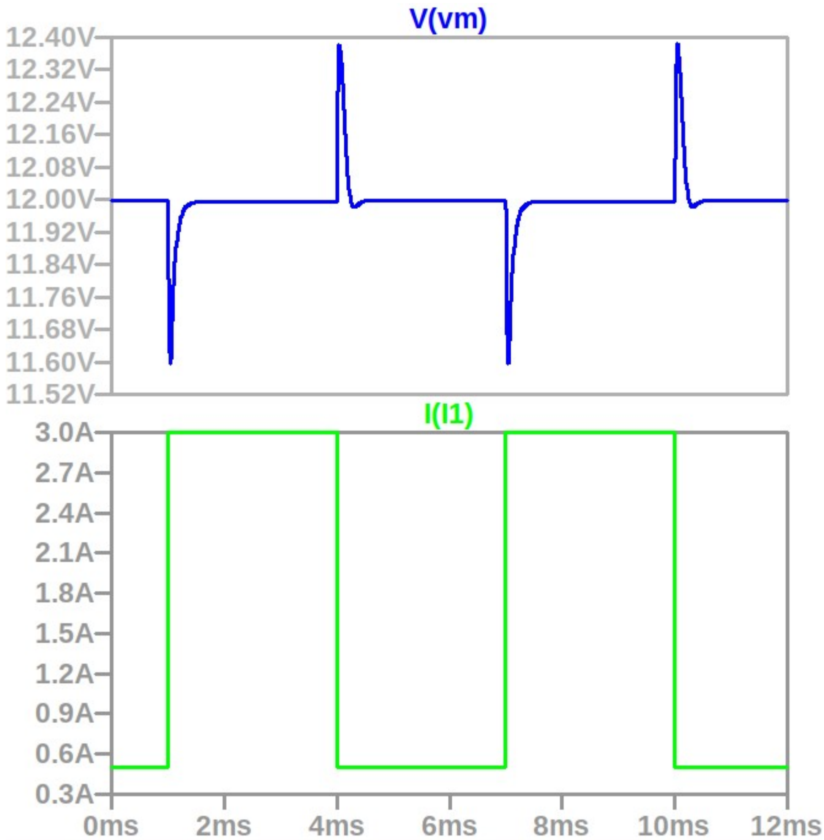
Once these elements are plugged into the simulator, a final run can be performed at both input levels. The results confirm a phase margin of ~ 60° together with a gain margin around 16 dB.

→ loop-gain transfer function at VIN = 5 VDC and 9 VDC





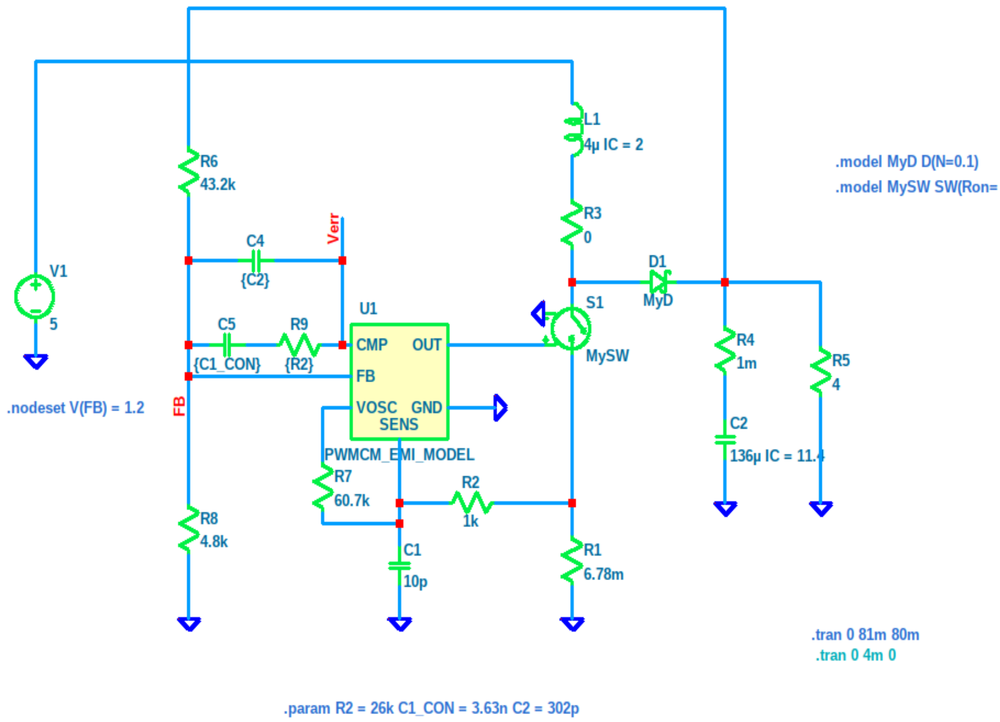
→ transient analysis on the averaged model



The transient test uses the averaged model together with a load evolving between 0.5 A and 3 A. The previous figure shows the results and confirms the stability. The drop stays within 0.5 V (approx 0.372 V) thus confirming our design specifications.

TRANSIENT ANALYSIS

Transient simulation template



The controller parameters are the following (extracted from LM5121):

Period = 3.33us Switching period (300 kHz)

DISABLE_OPAMP = 0

Vhigh = 5 Represantative of the output of EA

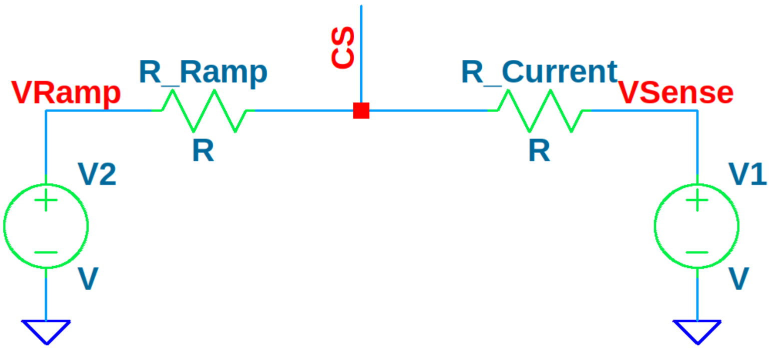
VOSC_H = 1.2 Oscillator max limit

REF = 1.2 Internal reference voltage of 1.2 V

RATIO = 0.033 Feedback signal is internally divided by 10

SLOPE COMPENSATION

To solve for the ramp resistor we can follow the following circuit that represents the actual electric circuit leading to the current sense pin:



$$V_{CS}=V_{sense}\left(\frac{R_{ramp}}{R_{ramp}+R_{current}}\right)+V_{ramp}\left(\frac{R_{current}}{R_{ramp}+R_{current}}\right)$$ using superposition

$$V_{CS}=\left(\frac{R_{ramp}}{R_{ramp}+R_{current}}\right)\left(V_{sense}+V_{ramp}\frac{R_{current}}{R_{ramp}}\right)$$

Now if we divide every term by t_{on}, we can express it by using slopes instead:

$$S_{CS}=\left(\frac{R_{ramp}}{R_{ramp}+R_{current}}\right)\left(S_{sense}+S_{ramp}\frac{R_{current}}{R_{ramp}}\right)\rightarrow$$

$$S_{CS}=k\left(S_{sense}+M_r S_{off}`\right)$$

where $k=R_{ramp}/(R_{ramp}+R_{current})$ and M_r represents the percentage of the off slope we want to inject.

$$S_{ramp}\frac{R_{current}}{R_{ramp}}=M_r S_{off}`$$

Solving for R_{ramp} leads to

$$R_{ramp}=\frac{S_{ramp}}{M_r S_{off}`}R_{current}$$ where Mr correspond to ramp coefficient we need, here 50%.

If we apply numercial values, we get:

$S`_{off}$ = 11.86 mV/us

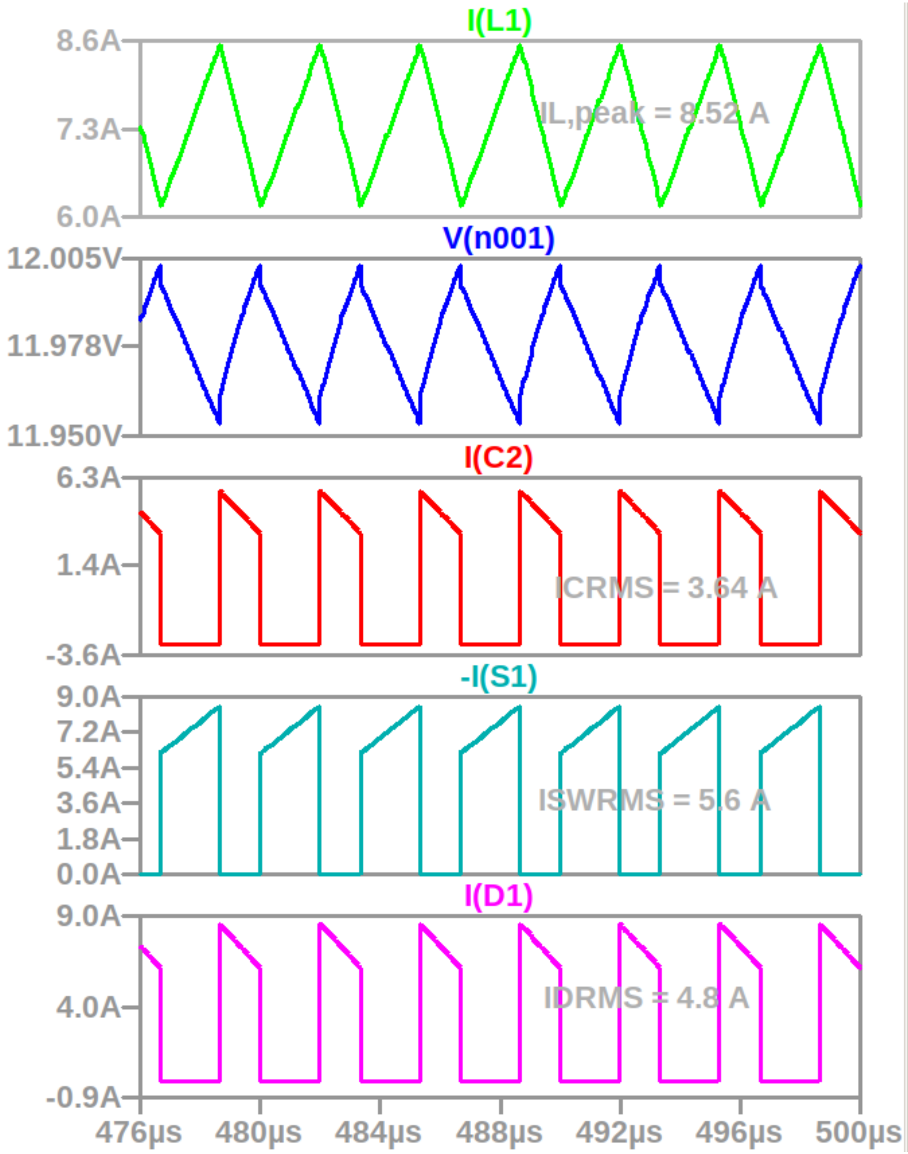
S_{ramp} = 1.2V/3.33us = 360 mV/us

$R_{current}$ = 1 kΩ (arbitralily fixed)

M_r = 50%

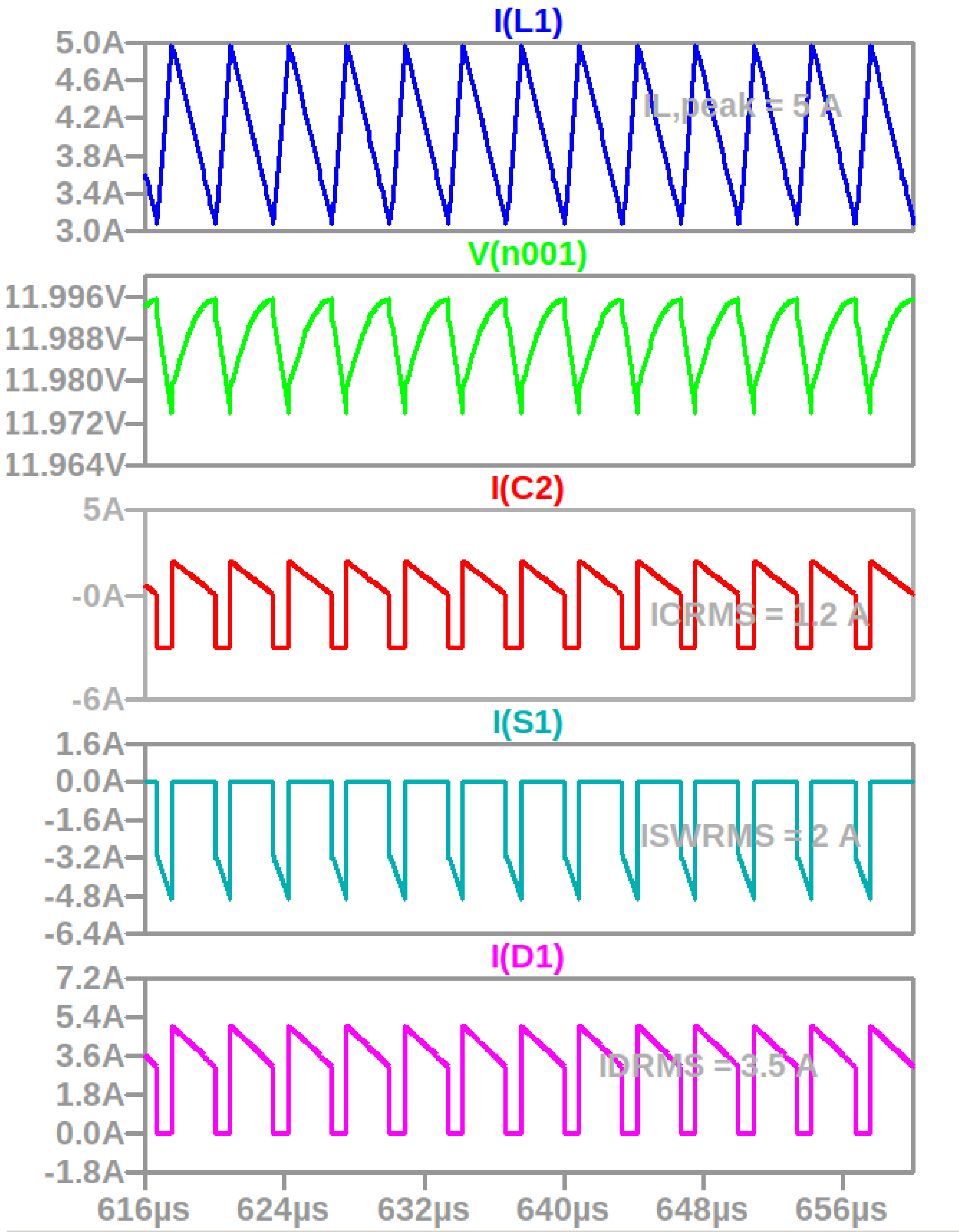
$$R_{ramp}=\frac{360\text{e-}3\,V/us}{\frac{1}{2}11.86\text{e-}3\,V/us}1\text{e}3=60.7\,k\Omega$$

→ simulation conditions at VIN = 5 V (worst case conditions)

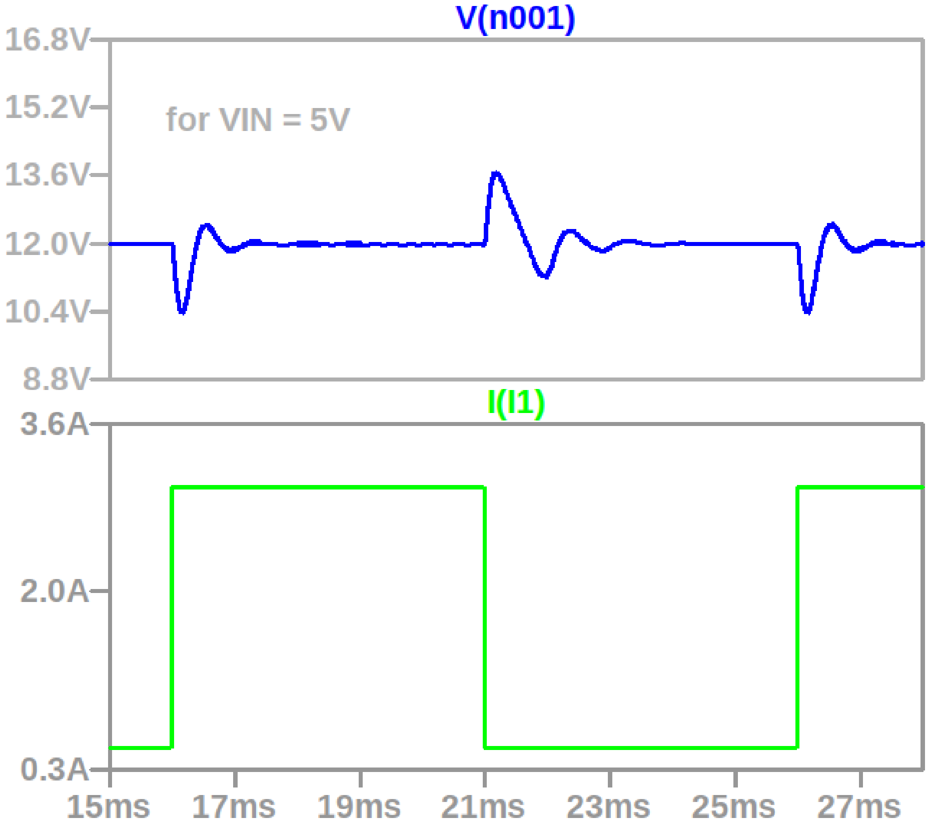


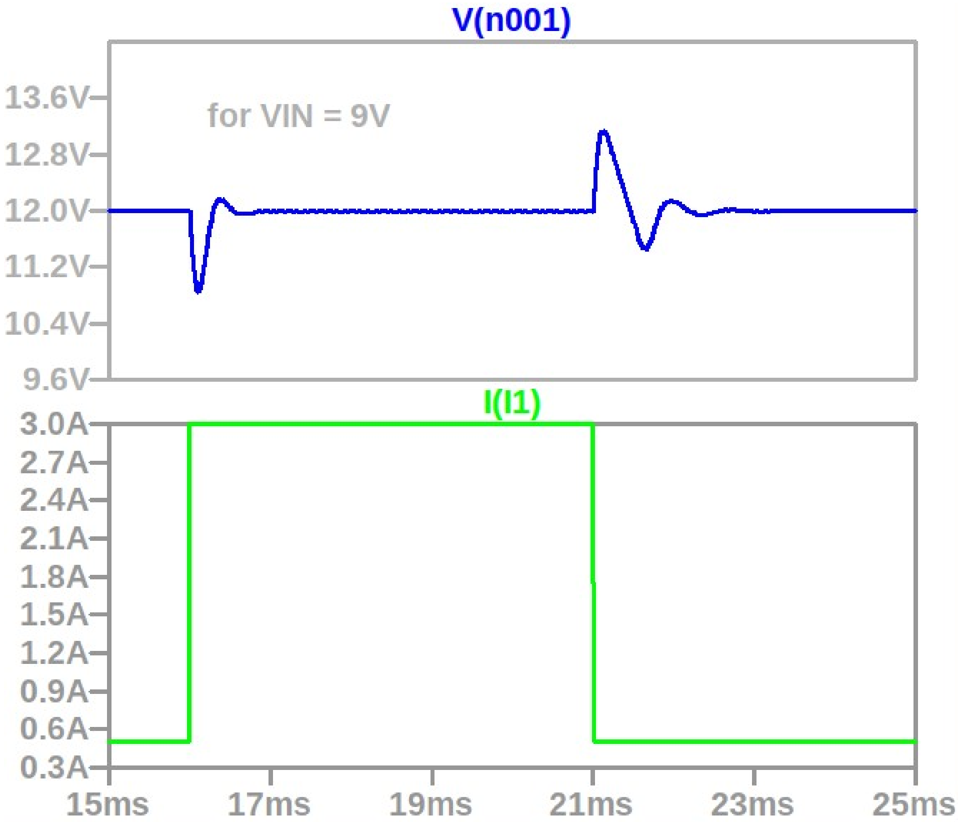
Measurements at low input voltage show a diode average power of 272mW whereas the MOSFET dissipates 315 mW. As usual, these numbers are indicative, and bench measurements must be taken to obtain the final numbers.

→ simulation conditions at VIN = 9 V



The transient repsonse is perfectly stable at both input levels as the following figures show:

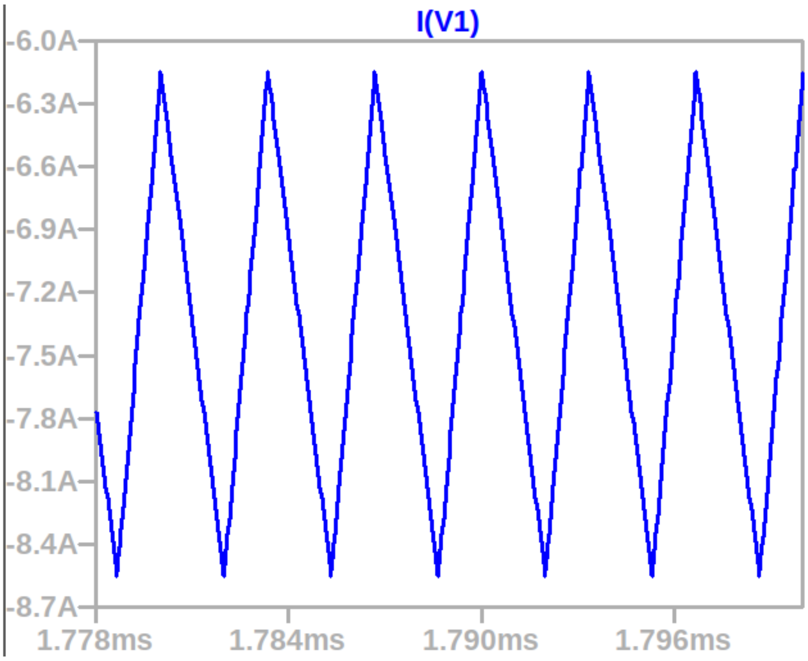


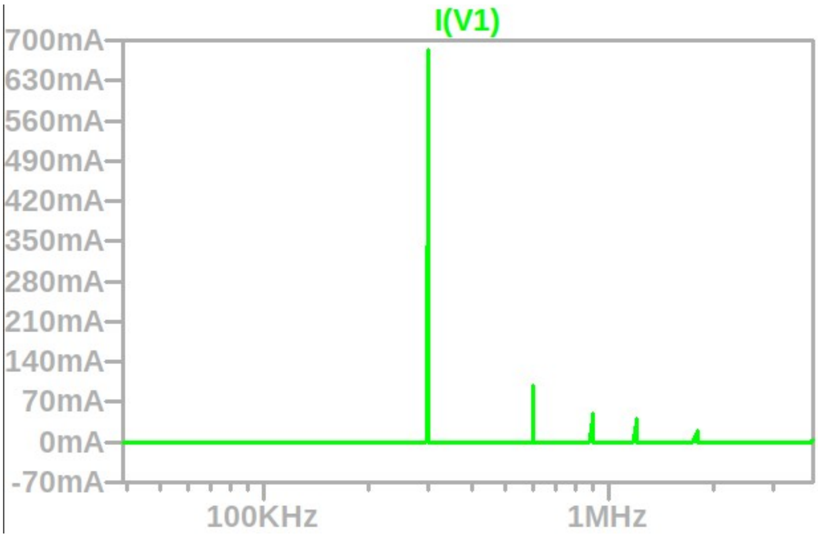


INPUT FILTER

For this design we have in input ac current specification of 1 mA peak to peak with filter installed.

(boost input current signature at VIN 5 VDC without input filter)





→ the peak current at 300 kHz reaches 686 mA

$$I_{peak}(fundamental) = 0.686\,A$$

$$Att_{filter} < \frac{1\,mA}{0.686\,A} < 1.45\text{e-}3 \text{ or } better\ than\ a\ 57\ dB\ attenuation$$

$$f_0 < \sqrt{Att_{filter}} \cdot F_{sw} < 11.5\,kHz$$

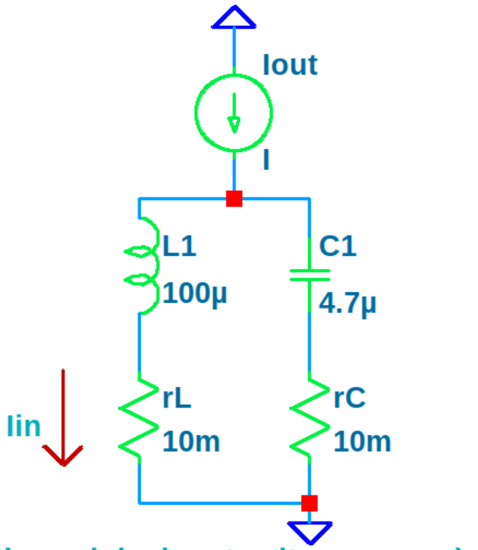
A key selection criteria for selecting capacitors is its rms current folwing through it, so the current flowing through Cin will be

$$I^2_{\text{in,rms}} - I^2_{\text{in,ac}} - I^2_{\text{in,dc}} = 0$$

$$I_{Cin,rms} = I_{\text{in,ac}} = \sqrt{I^2_{\text{L,rms}} - I^2_{\text{in,dc}}} = \sqrt{7.4074^2 - 7.3744^2} = 0.698\,Arms\ where\ I_{\text{L,rms}} = I_{\text{in,rms}}$$

For L = 10 uH we get,

$$C=\frac{1}{4\pi^2 f_o^2 L}=19\,uF$$



Iin (the current going through he input voltage source)

We verify the final attenuation at 300kHz of the filter with the parasitics added: (the transfer function can be derived by state-space matrix algebra or a simple current division)

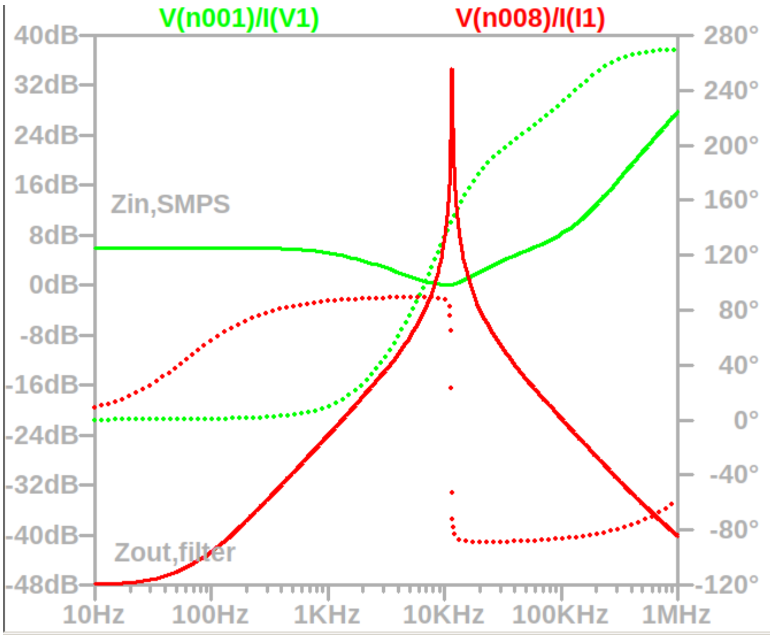
L = 10 uH rL = 3 mΩ C = 19 uF rC = 5 mΩ (note that this values might be iterated)

$$\left\|\frac{I_{in}}{I_{out}}\right\|_{at\,300kHz}=\sqrt{\frac{r_C^2+\frac{1}{(\omega C)^2}}{(r_L+r_C)^2+\frac{1}{(\omega C)^2}-\frac{2L}{C}+(\omega L)^2}}=\text{¿}$$

$$=\sqrt{\frac{0.005^2+\frac{1}{(2\pi\cdot 300\text{e}3\cdot 19\text{e-}6)^2}}{(0.005+0.003)^2+\frac{1}{(2\pi\cdot 300\text{e}3\cdot 19\text{e-}6)^2}-\frac{2\cdot 10\text{e-}6}{19\text{e-}6}+(2\pi\cdot 300\text{e}3\cdot 10\text{e-}6)^2}}=\frac{28.38\text{e-}3}{18.81}=1.5\,mA$$

which is approximately enough.

(both ac converter input impedance and filter output impedance are shown in the following graph)



After we added a damping network to the EMI filter we get the following result:

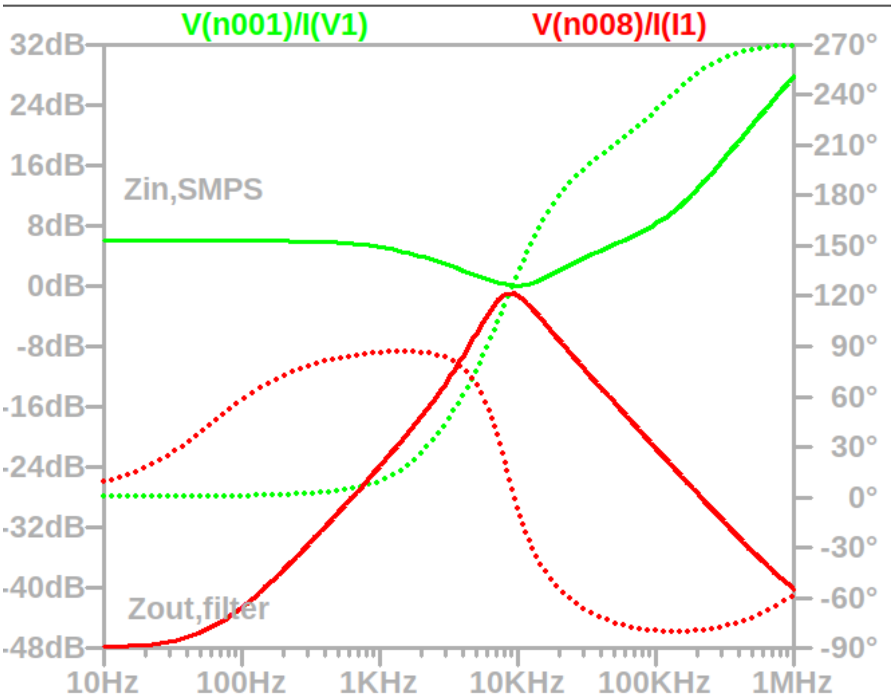
(filter circuit Rdamp ~ sqrt(L/C)) i.e., the characteristic impedance of L3 and C6

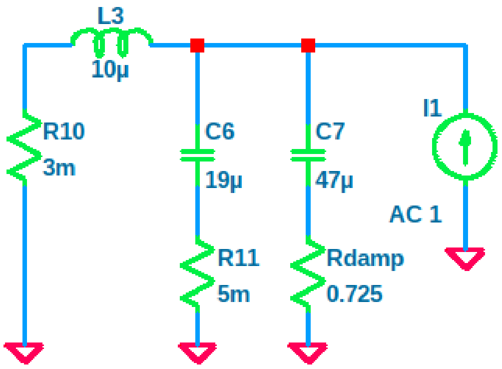
$$R_{damp}=\sqrt{\frac{L}{C}}=\sqrt{\frac{10e\text{-}6}{19e\text{-}6}}=0.725\,\Omega$$

C7 should be chosen to have lower impedance than Rdamp at the approximate resonance pulsating frequency of 1/sqrt(L3*C6)

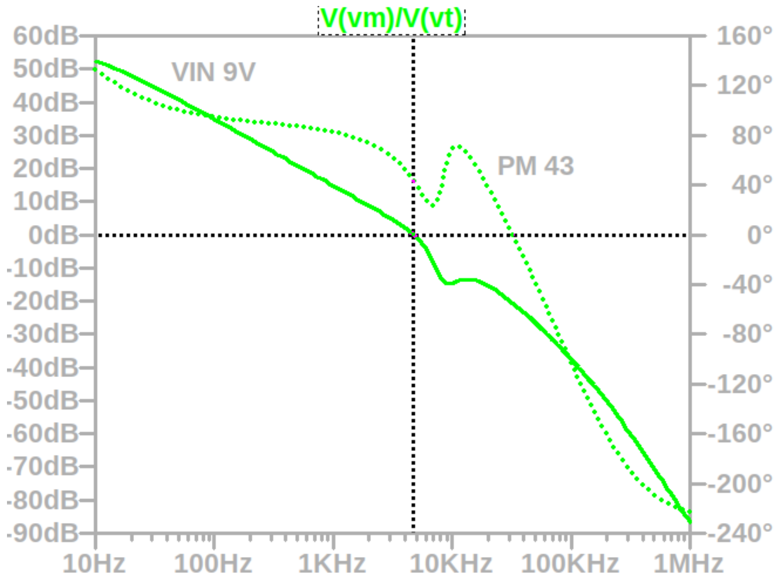
$$X_C=1/(2\pi f_0 C_7)\rightarrow C_7=1/(2\pi f_0 X_C)=1/(2\pi\cdot 11.4e3\cdot 0.2)=69\,e-6\,F\; for\, an\, impedance\, of\; X_C=0.2\,\Omega$$

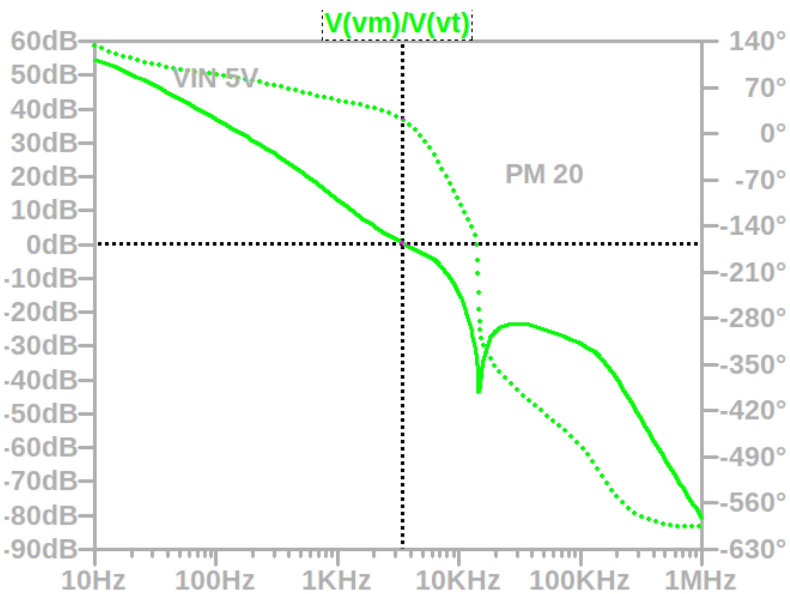
$$where\; f_f=\frac{1}{2\pi\sqrt{(L3*C6)}}=\frac{1}{2\pi\sqrt{(10e\text{-}6\cdot 19e\text{-}6)}}=11.5\,kHz$$





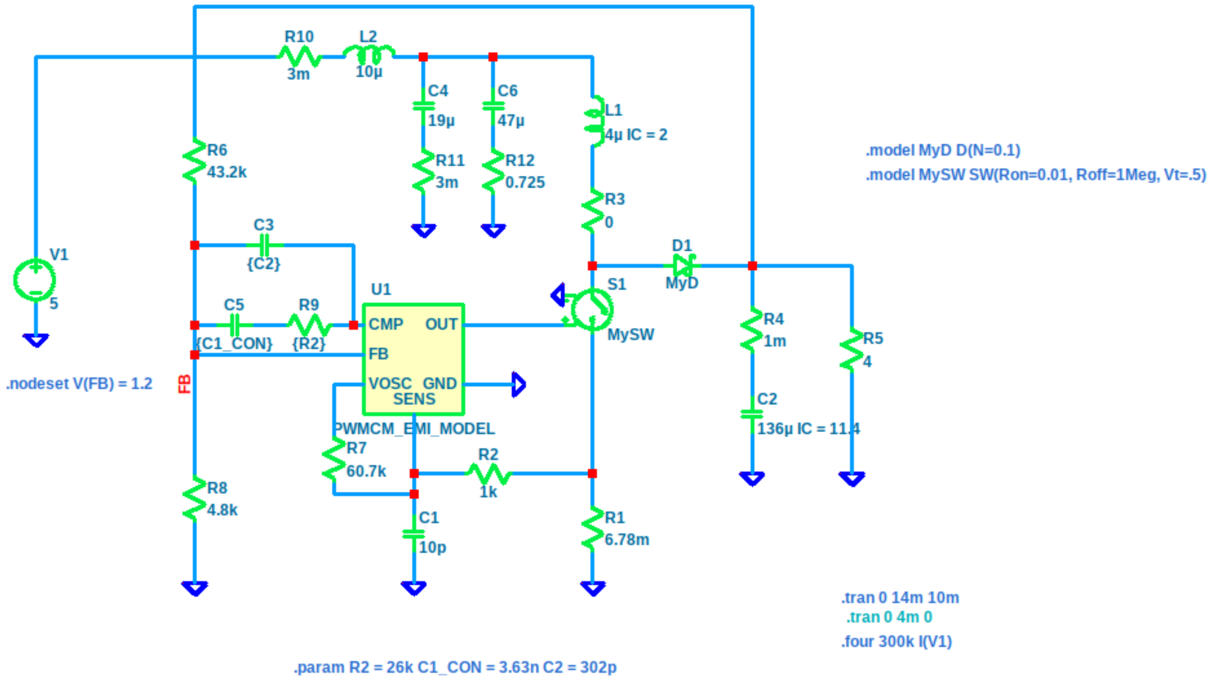
Once the input filter is properly damped we need to verify that the loop-gain phase margin does not suffer from the filter presence:



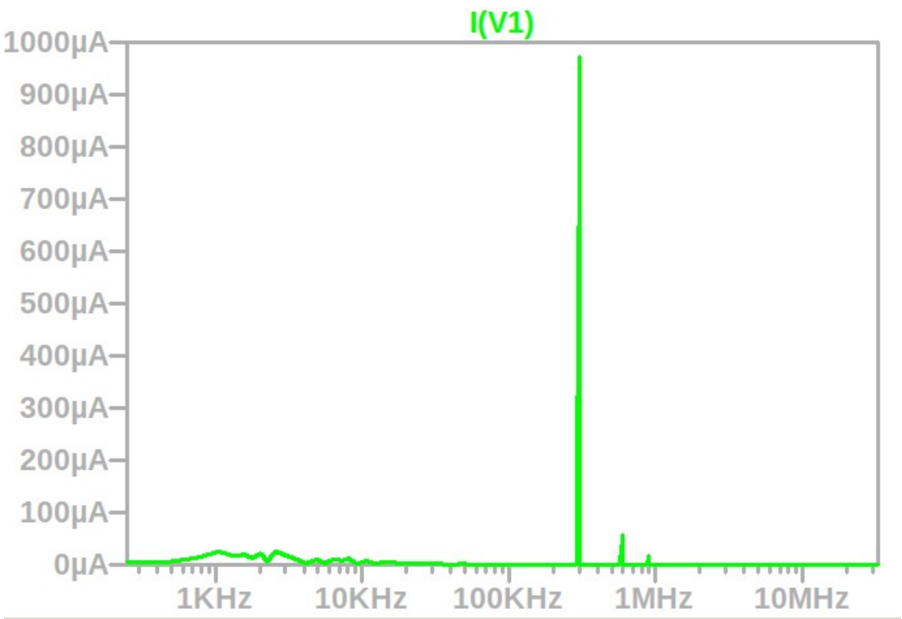


If the transient response on the final prototype is not addequate, a reiteration is needed.

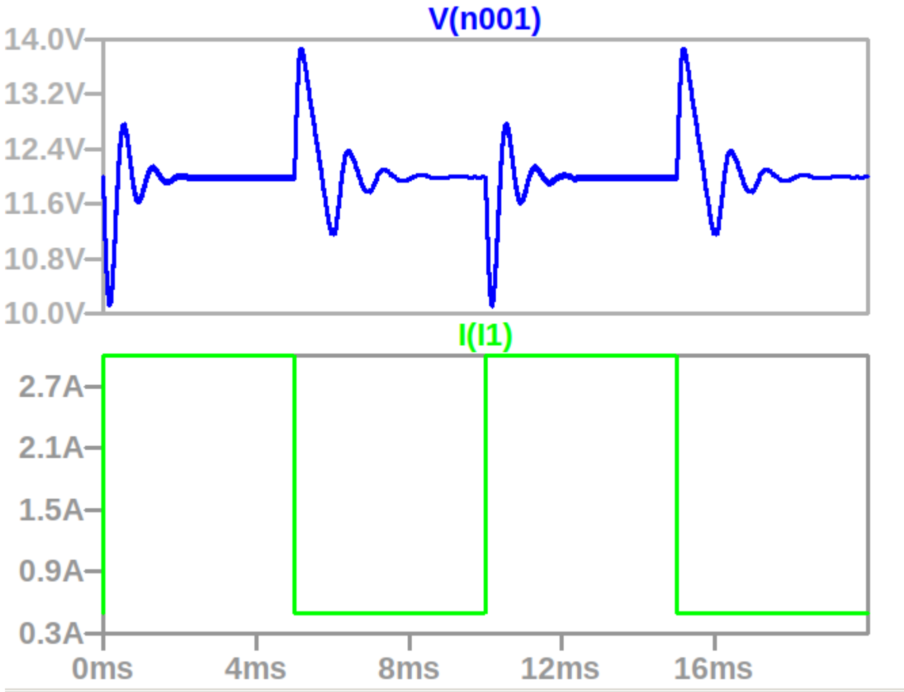
(final circuit)



(input current harmonics amplitudes)



→ $I_{peak(fundamental)}$ is less than 1 mA, but the transient response shows an oscillatory response now with the input filter added, which may not pass the demanded specifications.



→ possible controller selection: TI LM5121