Stabilizing a Converter with NCV12711

AND90138/D

NCV12711 is a high−performance current−mode dc−dc controllers continuously operating from a low 4 V input voltage up to 45 V. The part is well suited to drive flyback converters or non−isolated topologies such as a boost or a SEPIC used in a variety of applications. For stabilizing such a converter, a primary−side circuit is necessary to provide voltage regulation and frequency compensation. The on−board operational amplifier (op−amp) is perfectly suited for this function but it is important to understand how to design the compensation network. This is the object of this application note.

An Op−Amp in the Feedback Loop

The circuit can be wired in a type 2 compensator following the suggestion of Figure 1. A type 2 compensator assembles a pole at the origin, a zero and a high−frequency pole. By adjusting the distance between the zero and the pole, the designer can choose the *phase boost* he needs for his design. By phase boost, we imply the amount of phase lead generated by the compensator to build enough phase margin at the selected crossover frequency f_c [1] once the power stage is compensated. This phase boost can be determined by observing the control−to−output transfer function plot of the considered converter at the selected crossover frequency and assess the phase lag at f_c.

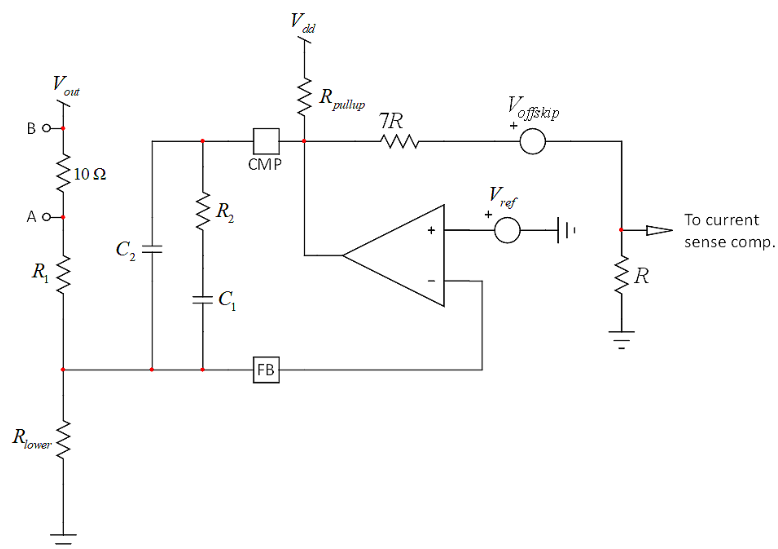

Figure 1. The Operational Amplifier is used for Regulation and Compensation Purposes in a Type 2 Compensator

The transfer function of the compensation path can be expressed as follows in a *low−entropy* form:

$$G(s) = \frac{V_{FB}(s)}{V_{out}(s)} = -G_0 \frac{1 + \frac{\omega_z}{s}}{1 + \frac{s}{\omega_p}} \quad \text{(eq. 1)}$$

If we consider capacitor C_2 much smaller than C_1, we have the following simplified definitions:

$$G_0 = \frac{R_2}{R_1} \quad \text{(eq. 2)}$$

$$\omega_z = \frac{1}{R_2 C_1} \quad \text{(eq. 3)}$$

$$\omega_p \approx \frac{1}{R_2 C_2} \quad \text{(eq. 4)}$$

In this expression, G_0 represents the so−called *mid−band gain* which is the magnitude value of the compensator forcing a 0 dB loop gain $T(s)$ at the selected crossover. We will see in the below lines how to determine the value of this mid−band gain. R_{pullup} is the internal equivalent resistance seen at the COMP pin and is approximately 5 kΩ for the NCV12711. It sets the maximum current the op−amp can source.

The typical asymptotic response of this compensator appears in Figure 2. The pole and the zeros are adjusted to produce the desired phase peak occurring at the crossover frequency f_c. In the figure, they are set apart to provide a 50° phase boost as an example. We can show that this peak occurs at the geometric means of the zero and the pole position and this is where the crossover frequency is usually selected:

$$f_c = \sqrt{f_z f_p} \tag{eq. 5}$$

Several methods exist among which the k factor introduced in the 90's [2] provides a convenient tool.

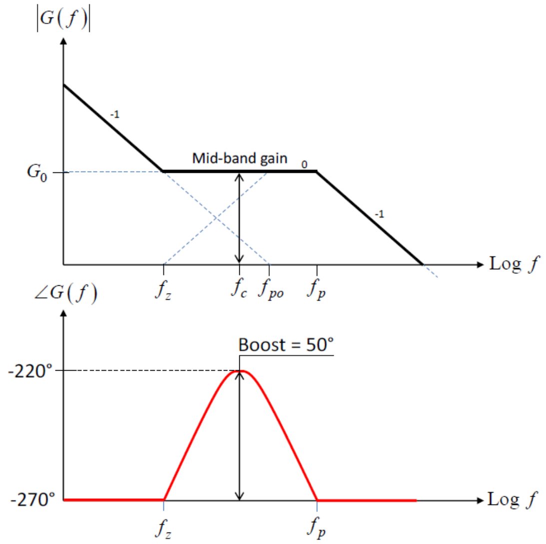

Figure 2. By Adjusting the ω_{po} Pole, it is possible to set the Mid−band Gain to the needed Value

The k factor is determined based on the wanted phase boost at a given crossover f_c. With its value on hand, you can determine where to place the zero and the pole to produce the necessary phase lead. You first determine the mid−band gain G_0 to force the magnitude to cross over the 0 dB axis at f_c by adjusting resistance R_2. With the pole and zero on hand, you carry on and determine C_1 and C_2.

Design Example

The control−to−output transfer function determination represents the starting point to stabilize a converter: if you stimulate the feedback pin, how does the *stimulus* propagate through the converter to generate a *response* across the load? Without the power stage response, there is no point trying to compensate a converter regardless of its topology. The transfer function can be obtained in different ways:

1. *Analytical analysis*: you can use a small−signal model and derive the transfer function by solving equations and plotting the results to extract the information you need.
2. *SPICE simulation*: with an averaged model such as the PWM switch, you can simulate the converter and extract its power stage response. Some simulators such as LTspice offers the possibility to extract the ac response from a switching circuit but the process is extremely time consuming.
3. *Prototype test*: you assemble a prototype and measure the response with a network analyzer
4. *Piecewise linear simulator*: SIMPLIS® is a typical PWL simulator which lets you extract the small−signal response of a switching converter without resorting to an average model.

We will look at the SIMPLIS simulation that appears in Figure 3. It is a non−isolated 12 V converter. No need to simulate the entire controller − especially all protection circuits that do not impact the ac response – as what matters is the control−to−output chain including the feedback pin, the current−sense comparator and the various dividers. Delays can be added if significant compared to the switching period of course.

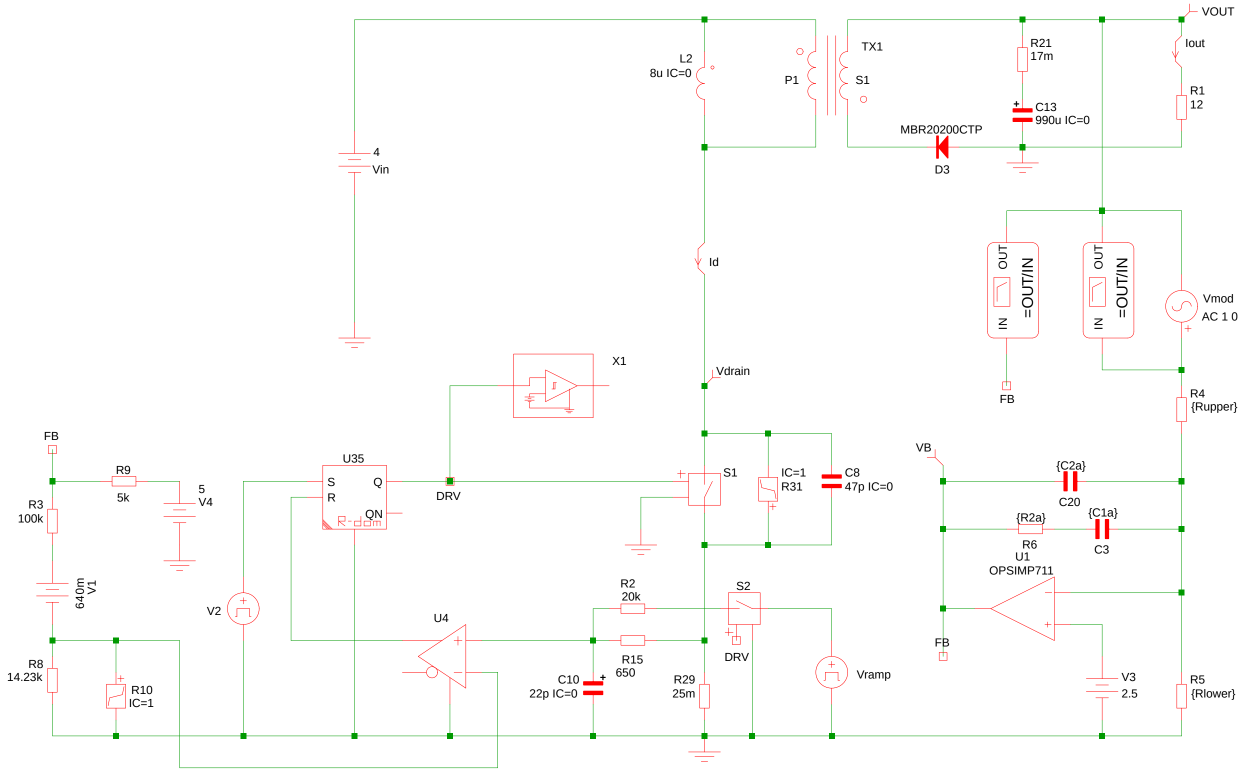

Figure 3. A SIMPLIS Simulation is the Fastest Way to look at Switching Waveforms and Obtain the Power Stage Frequency Response

This circuit works on the demonstration version Elements and will deliver all the interesting waveforms in a few seconds. We are looking at a 4 V converter delivering 12 V and loaded by a 1 A output current. SIMPLIS very precisely determines the steady−state operating point via its periodic operating point analysis or POP. The cycle−by−cycle waveforms are given in Figure 4. Once this is obtained, the ac analysis can take place and results are delivered in Figure 5. However, before we select a possible crossover frequency, we need to check the position of the right−half−plane zero (RHPZ) which limits the maximum possible value. It is determined using the following formula:

$$f_{z_1} = \frac{(1 - D)^2 R_{load}}{2\pi D L_p N^2} = \frac{(1 - 0.6)^2 \times 12}{6.28 \times 0.6 \times 8u \times 2^2} \approx 16\ \text{kHz} \qquad \text{(eq. 6)}$$

N is the transformer turns ratio with the convention 1:N.

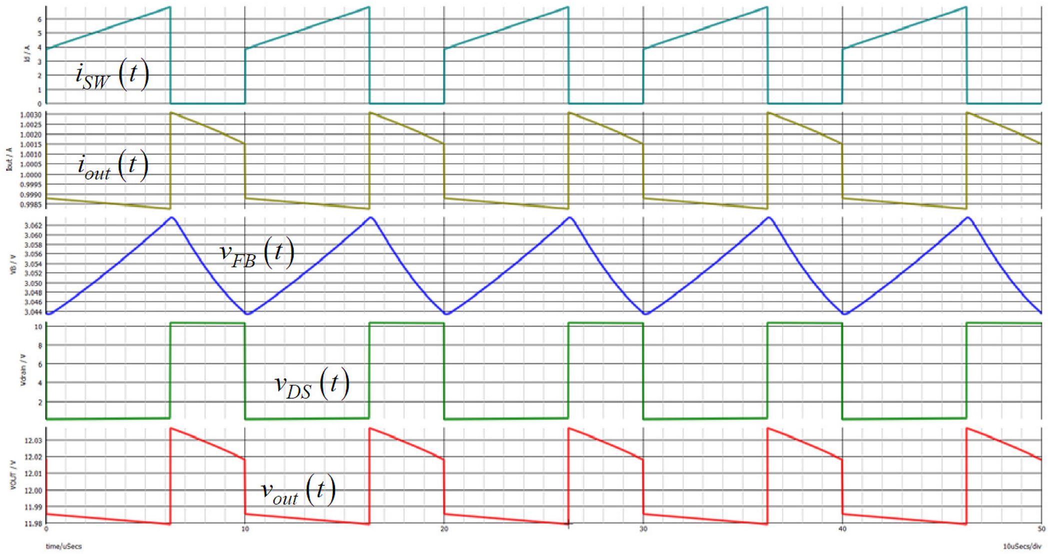

Figure 4. The Switching Waveforms confirms the Correct Operation at a 4 V Input Voltage. These Idealized Waveforms let you quickly Determine the ac Response of the Converter.

The RHP zero unlike a left−half−plane type lags the phase and distorts the phase margin if the crossover is too close to its location. We recommend selecting a crossover frequency at 20−30% of its lowest position. In our case, we have selected a crossover of 1 kHz to cope with this requirement. From the Bode plot, we can extract two interesting parameters: the attenuation at the chosen crossover frequency and the phase at this value. We read a 20 dB attenuation (G_{fc}) at 1 kHz together with a phase lag of 87°. With these values on hand, we can infer how to tailor our type 2 compensator to boost the phase by this amount:

$$\text{Boost} = \text{PM} - \angle H(f_c) - 90° = 70 - (-87°) - 90° = 67° \quad \text{(eq. 7)}$$

In which PM is the phase margin we target or 70° in this example. The k factor can be determined using the below formula:

$$k = \tan\left(\frac{\text{boost}}{2} + 45°\right) = \tan\left(\frac{67}{2} + 45\right) \approx 4.9 \quad \text{(eq. 8)}$$

Following [1], the pole and zero are placed as follows:

$$f_z = \frac{f_c}{k} \approx 203\ \text{Hz} \quad \text{(eq. 9)}$$

And

$$f_p = k \cdot f_c \approx 4.9\ \text{kHz} \quad \text{(eq. 10)}$$

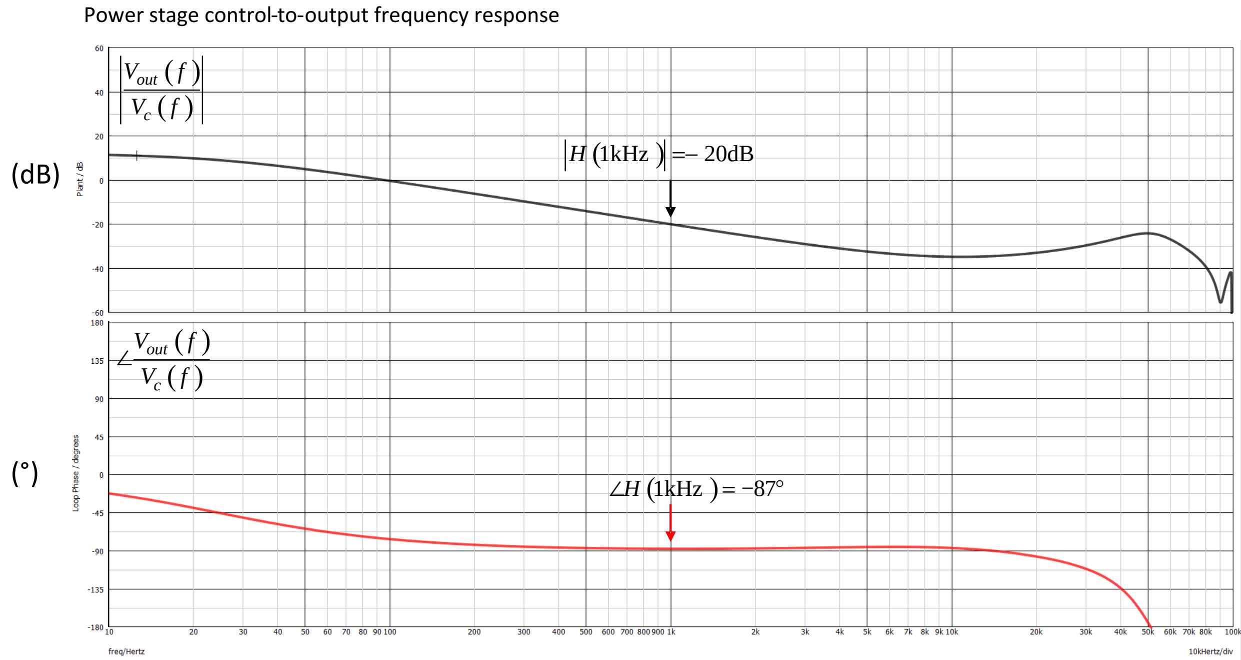

Figure 5. This is the Control−to−Output Transfer Function obtained at the Lowest Input Voltage and the Highest Output Current

Assuming an upper resistor value of 9.5 kΩ (R_1 in Figure 1), you can calculate the value of resistor R_2 based on the compensation gain at f_c:

$$R_2 = R_1 \cdot 10^{-\frac{G_{fc}}{20}} = 9.5k \times 10^{-\frac{20}{20}} = 95\ \mathrm{k\Omega} \qquad \text{(eq. 11)}$$

Considering capacitor C_1 much smaller than C_2 [1], their values can be determined as follows:

$$C_2 = \frac{1}{2\pi\, R_2\, f_p} = \frac{1}{6.28 \times 95k \times 4.9k} = 340\ \mathrm{pF} \qquad \text{(eq. 12)}$$

And

$$C_1 = \frac{1}{2\pi\, R_2\, f_z} = \frac{1}{6.28 \times 95k \times 203} = 8.2\ \mathrm{nF} \qquad \text{(eq. 13)}$$

Using the dedicated Excel™ sheet available from the NCV12711 landing page, you can immediately verify if the compensator response matches your needs. Figure 6 shows a screen capture for the given example. The RHPZ is calculated assuming a CCM operation and the lowest input voltage.

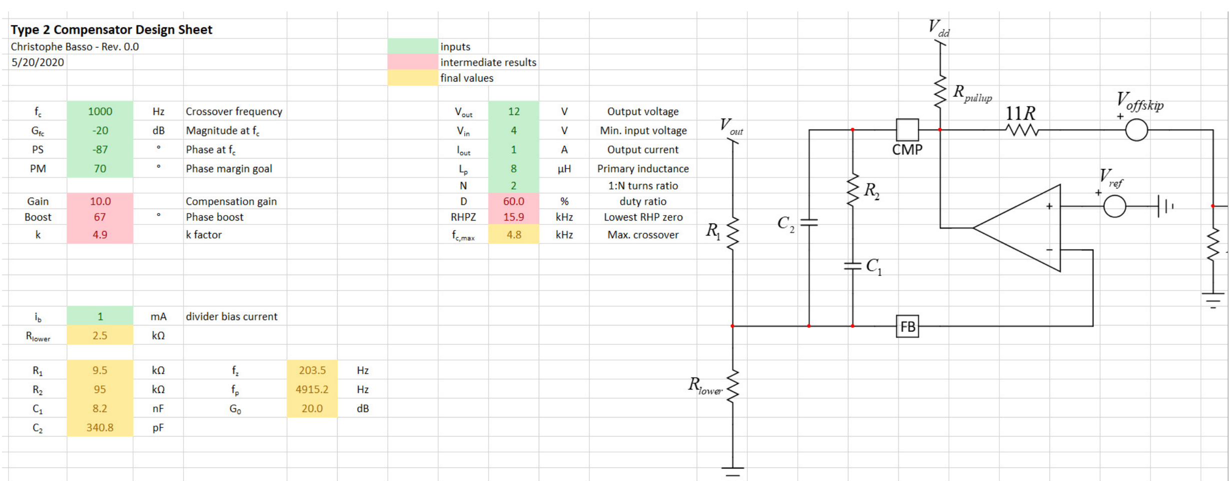

Figure 6. The Sheet Calculates the Components Values around the Op−Amp in a Type 2 Configuration

The magnitude−phase frequency response is immediate as shown in Figure 7 and confirms a response centered at 1 kHz with a phase boost greater than 60°.

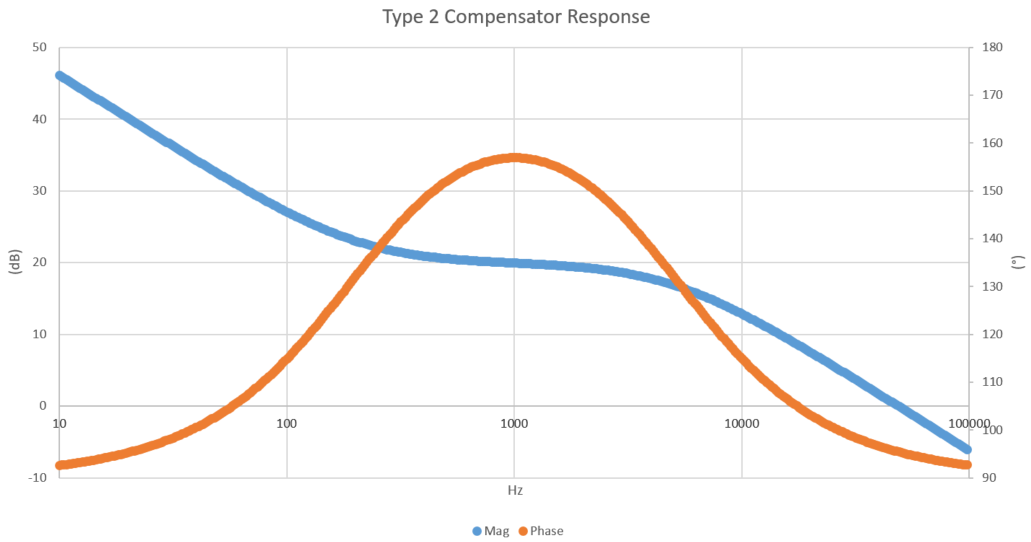

Figure 7. The Type 2 Compensator Produces the Expected Phase Boost at 1 kHz

Please note the presence of a 10 Ω resistance in series with the resistive divider (Figure 1). I recommend its inclusion during the prototype stage. It does not affect the circuit performance at all and it lets you easily connect a frequency−response analyzer (FRA) later on, when the board comes back from production. In high−density designs, it is often uncomfortable to cut traces and insert a series resistance as proposed. When it is already there with two little pins in place, your job is easier.

We can now run a new simulation and look at the open−loop gain to show what crossover and phase margin we obtain. The compensated loop gain appears in Figure 8.

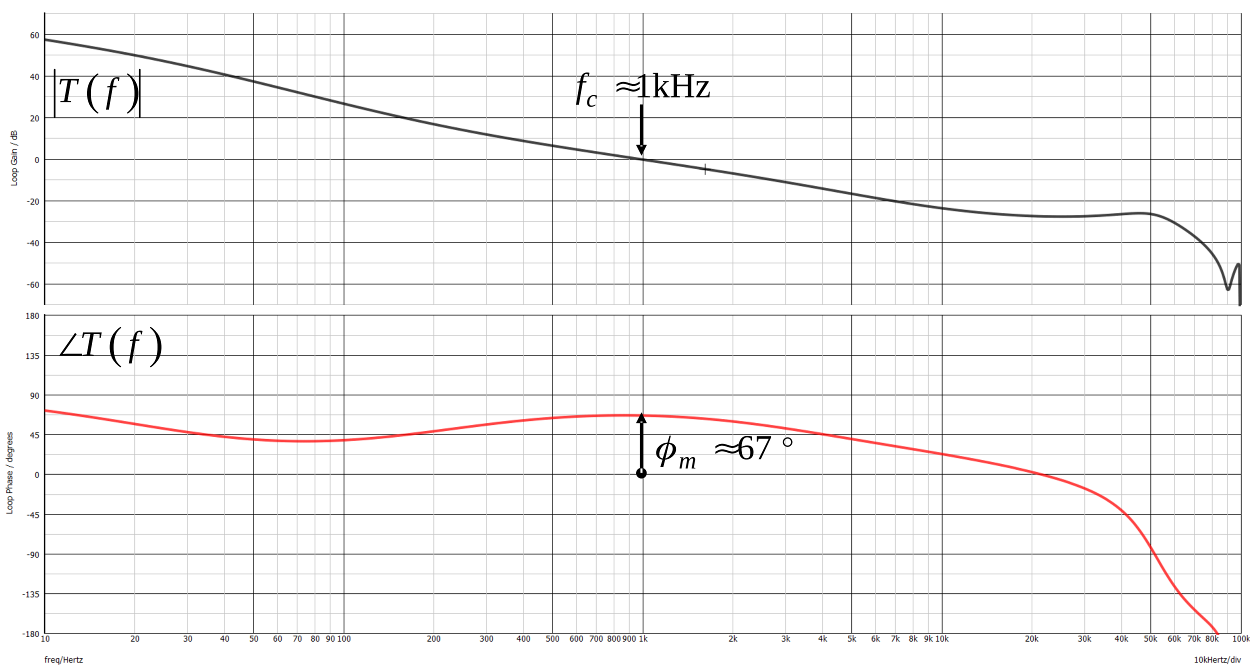

Figure 8. The Compensated Loop Gain shows a Crossover Frequency of around 1 kHz with a Decent Phase Margin below 70° (V_{in} = 4 V)

If we increase the input voltage to 45 V, which is the upper range for this small module, the crossover moves to 1.7 kHz with the same phase margin.

Based on this compensation circuit, we can check the transient response with the recommended component values. This time, we can use the complete model in SIMPLIS which includes the oscillator, the various protections, the soft−start circuitry with frequency sweep and so on. The circuit requires the full version while the previous simplified approach could work with the free demonstration version. The simulation template appears in Figure 9. Simulation results shown in Figure 10 for a 4 V input voltage confirm a smooth start−up sequence without overshoot. The controller begins by increasing the switching frequency at low peak currents and goes full speed once the soft−start period is over. The simulation circuit hosts the components values previously calculated and a zoom in the transient response appears in Figure 11. The deviation for a 0.8 A step in 1 μs is around 100 mV peak which is excellent.

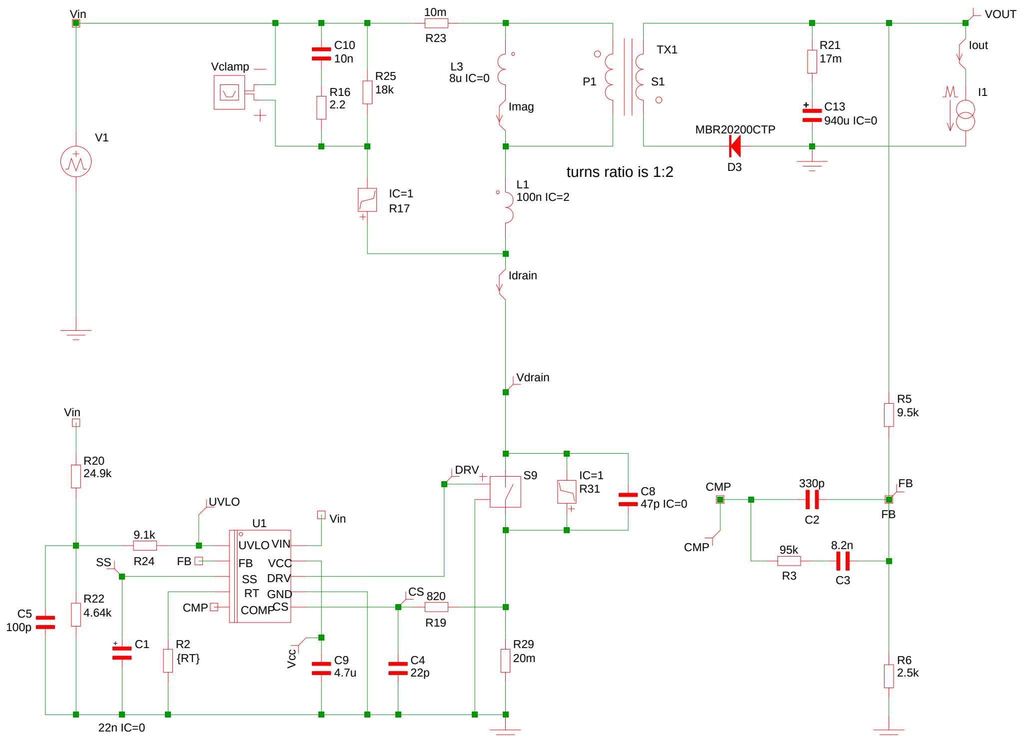

Figure 9. The Complete Simulation Model requires the Full Version to Deliver the Transient Waveforms at the Two Input Line Levels

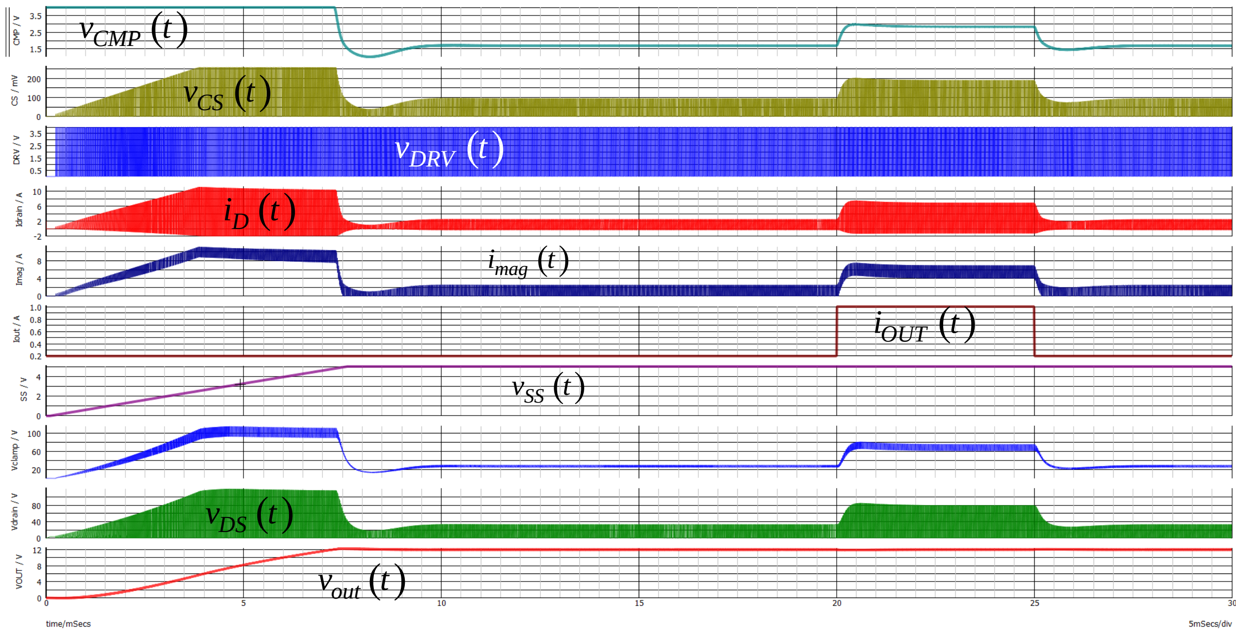

Figure 10. The Simulated Start−up Sequence shows the Frequency Sweep starting at 25 kHz. V_{in} = 4 V, I_{out} = 1 A

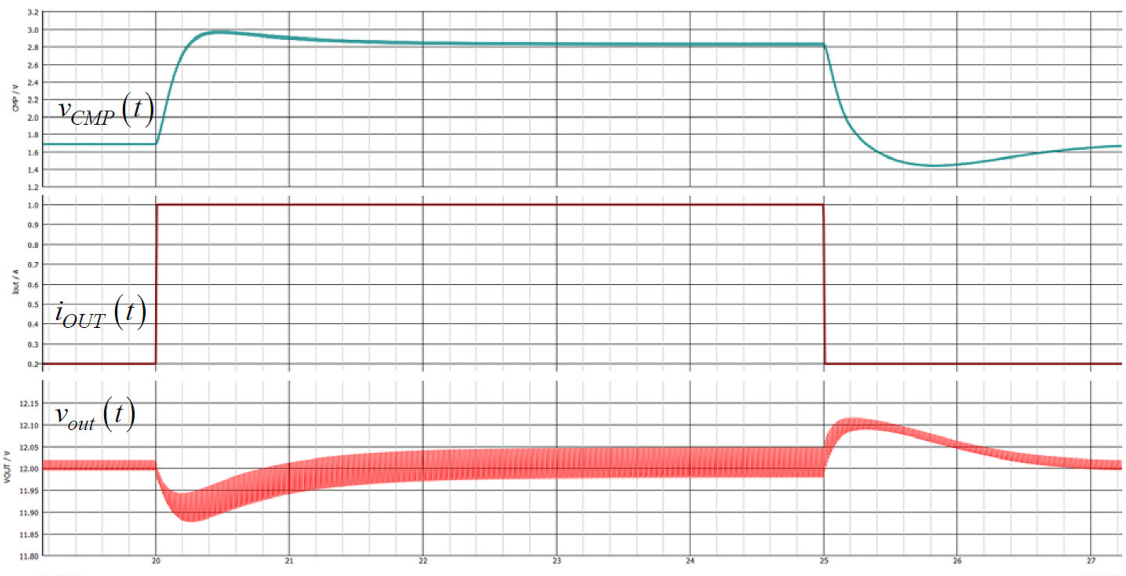

Figure 11. The Transient Response is Stable and shows an Output Voltage Under Control during the Load Steps

A Boost Converter

Capitalizing on the presence of the on−board op−amp, it is possible to implement NCV12711 in a boost converter or any other type of circuit such as SEPIC or Zeta also. The simulation circuit appears in Figure 12. This is a 250 kHz 24 V boost converter operated from a 12 V nominal input and capable to deliver up to 0.5 A when the input voltage is down to 4 V.

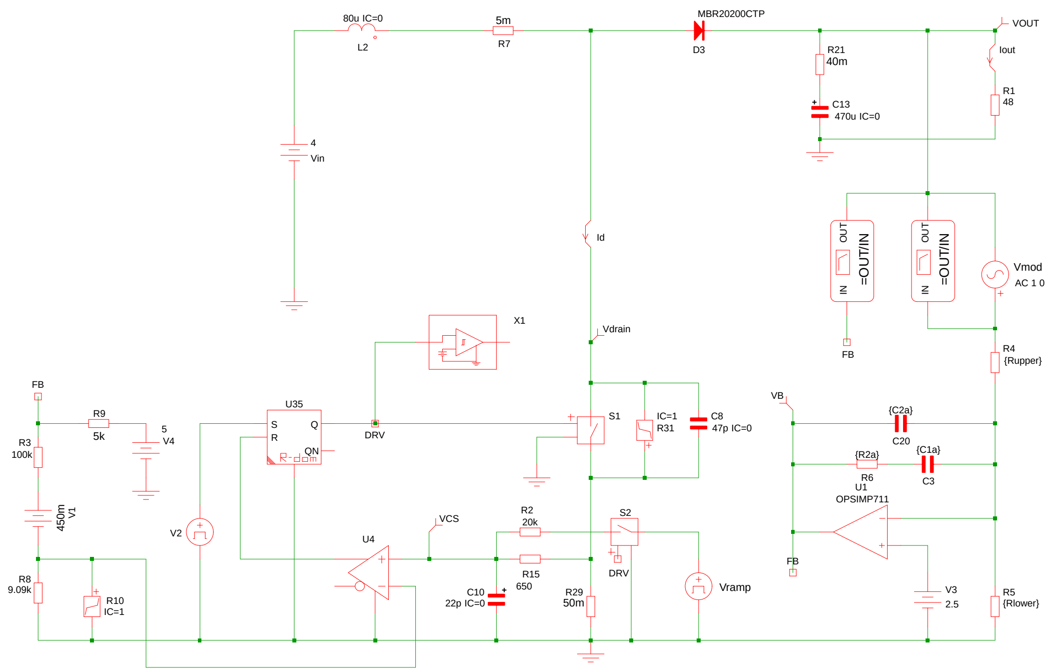

Figure 12. This Boost Converter delivers 24 V/0.5 A from a 4 V Input Voltage Source. The Simplified Schematic will help us Extract the Control−to−Output Transfer Function at the Lowest Input Voltage.

The compensation strategy is similar to what we have seen in the flyback case. We start with the power stage response and it appears in Figure 13.

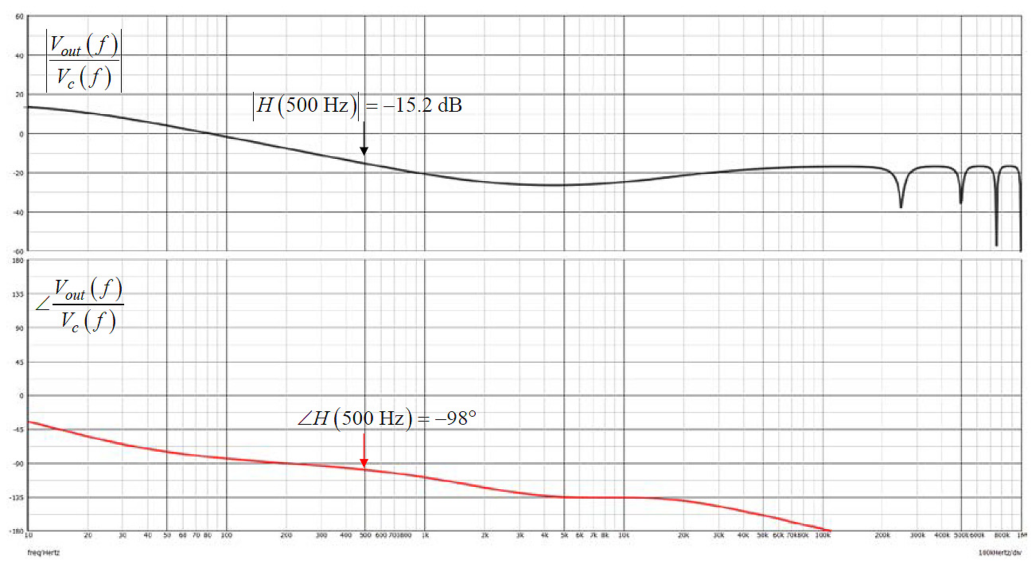

Figure 13. The Crossover Frequency is necessarily Limited considering the RHPZ at 4 V

In a boost converter, the worst−case RHP zero limits the crossover frequency. At a 4 V input voltage with a duty ratio of 84%, the RHPZ is positioned at:

$$f_{z_2} = \frac{(1 - D)^2 R_{load}}{2\pi L} = \frac{(1 - 0.84)^2 \times 48}{6.28 \times 80u} \approx 2.4\ \text{kHz} \qquad \text{(eq. 14)}$$

To maintain adequate phase margin, the crossover frequency cannot exceed 500 Hz (≈ 20% of f_{z2}). The magnitude and phase values at this frequency are extracted from the power stage Bode plot and lead to the following compensation values (Figure 14):

f_c	500	Hz	Crossover frequency		
G_{fc}	-16	dB	Magnitude at f_c		
PS	-100	°	Phase at f_c		
PM	60	°	Phase margin goal		
Gain	6.3		Compensation gain		
Boost	70	°	Phase boost		
k	5.7		k factor		
I_b	1	mA	divider bias current		
R_{lower}	2.5	kΩ			
R_1	21.5	kΩ	f_z	88.2	Hz
R_2	136	kΩ	f_p	2835.6	Hz
C_1	13.3	nF	G_0	16.0	dB
C_2	413.7	pF			

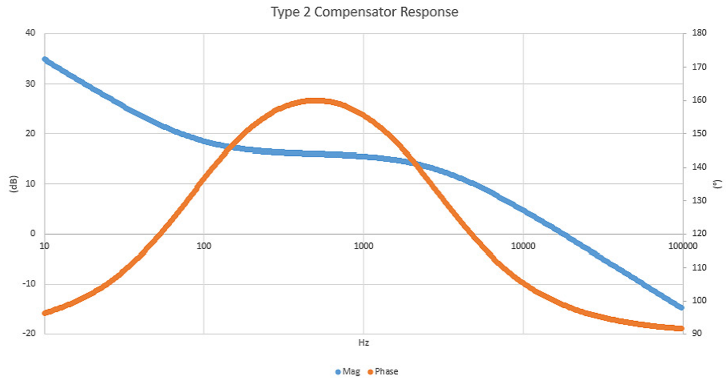

Figure 14. The Op−Amp is Compensated to Produce a Phase Boost of 70°

We can now try these values in the more comprehensive model shown in Figure 15. The frequency is set to 250 kHz by pressing F11 in SIMPLIS.

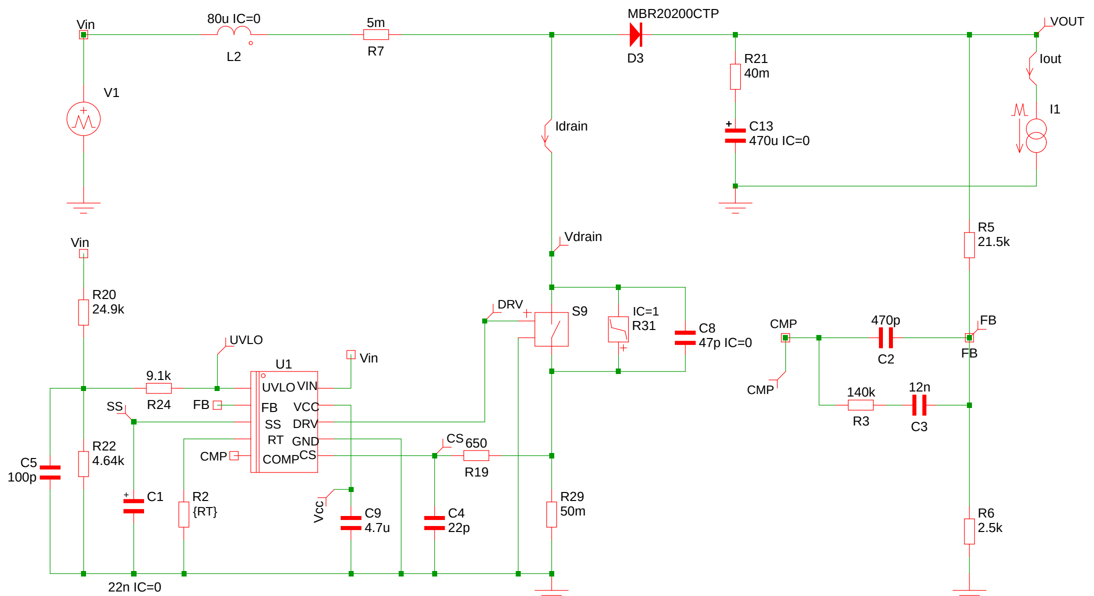

Figure 15. The NCV12711 Model is now at Work to Simulate this Boost Converter

To confirm stability, we have run a transient load step from 100 to 500 mA in 1 µs at two different input voltages, 4 and 15 V. As Figure 16 confirms, the deviation stays within a maximum of 200 mV at the lowest input voltage.

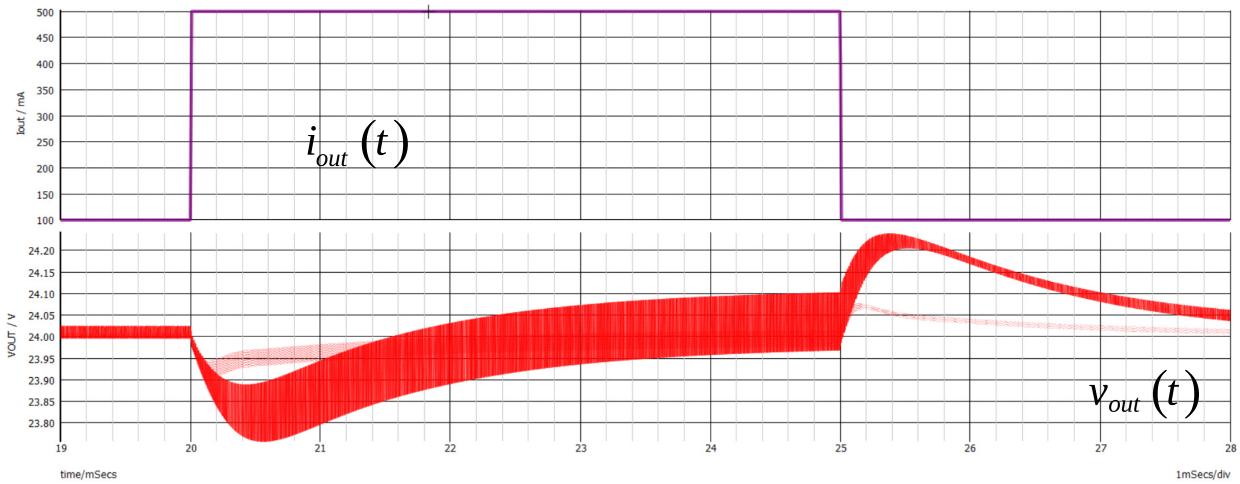

Figure 16. The Boost Converter shows an Excellent Transient Response at Both Input Voltage Extremes

Practical Measurements

It is important to conduct measurements on a prototype to validate the calculations carried with a theoretical model. Owing to the presence of the 10 Ω series resistance on the non−isolated 12 V flyback board, it is easy to clip the probes from the transformer and route signals to the frequency−response analyzer (FRA). The modulation amplitude is automatically adjusted along the frequency axis as the FRA approaches crossover. This is important to avoid saturation phenomena where the system offers less loop gain because the ac modulation is a perturbation fought by the converter. At low frequency, where a high gain exists, you may need a strong amplitude to observe a meaningful signal in the output. On the contrary, as the gain drops in the vicinity and beyond crossover, you will have to reduce the modulation signal considering the reduced rejection capability of the closed−loop converter. If changing the modulation amplitude shows up as a clearly modified Bode plot, then nonlinearity is at work and modulation level or operating point must be accordingly adjusted to bring the converter back into linear mode.

We have slightly changed the compensation network to operate with a 250 μA bias current for the resistive bridge. Resistances R_6 and R_5 in Figure 15 are now 10 kΩ and 38 kΩ respectively. These new values have been entered in our calculation sheet and the obtained values (C_3 = 2.2 nF in series with R_3 = 390 kΩ and in parallel with C_2 = 82 pF) soldered on our demonstration board. The loop gain appears below and shows a 1.5 kHz crossover frequency with a comfortable 61° of phase margin (Figure 17):

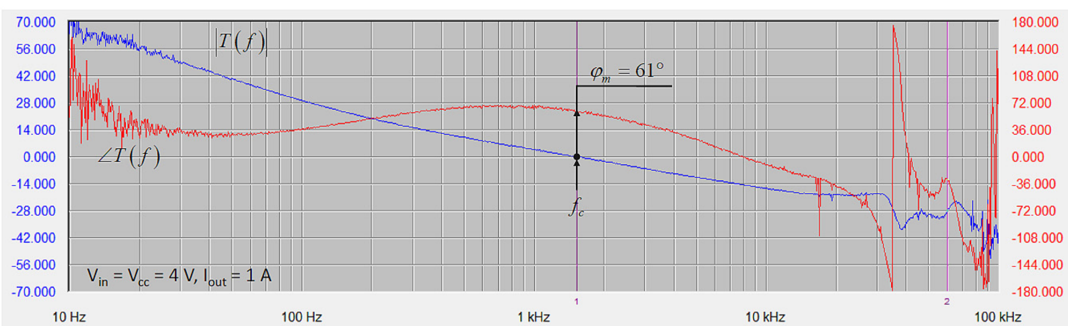

Figure 17. The Loop Gain at Low Line shows a Comfortable Phase Margin

If we now increase the input voltage to 28 V, the loop gain is still very good and shows good phase and gain margins (Figure 18). The crossover reaches 3.1 kHz in this condition:

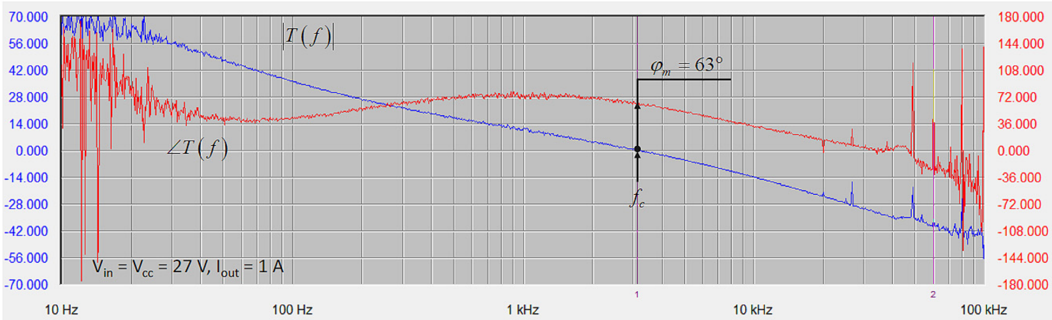

Figure 18. The Loop Gain Does Not Change Much at High Line

Finally, a step−load response will tell us if the overshoot we observe is compatible with our requirements. The response in low− and high−line conditions appears in Figure 19. The response is very stable at both extremes. It is actually possible to estimate what the drop will be considering crossover and output capacitance [1]. Provided the equivalent series resistance (ESR) is small compared to the output capacitor impedance at f_c, the drop can be approximated as:

$$\Delta V_{out} = \frac{\Delta I_{out}}{2\pi f_c C_{out}} = \frac{0.8}{6.28 \times 1500 \times 990u} \approx 86 \text{ mV} \quad \text{(eq. 15)}$$

which is not far from what we measured at low line with the 1.5 kHz crossover frequency.

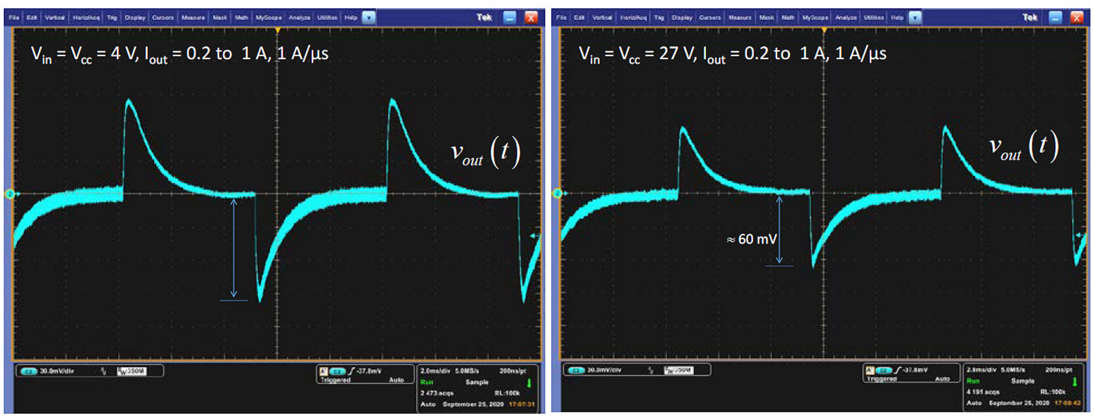

Figure 19. The Transient Response is Excellent at Low (4 V) and High Line (27 V)

Conclusion

This application note describes how to compensate flyback and boost converters built around NCV12711. Using a SIMPLIS model offers a convenient way to verify if the calculated elements meet the design targets such as crossover and phase/gain margins. When you go to the bench, you realize that the values suggested by the automated sheet are very close to the final ones, naturally minimizing the development time towards a final version. This experimental phase on the prototype is mandatory to check how the compensated converter behaves once powered in different conditions. When the model is validated by the hardware, you can run multiple Monte Carlo analyses to check all margins are stable across production spreads.

Reference

[1] Christophe Basso, *Designing Control Loops for Linear and Switching Power Converters*, Artech House, Boston, 2012

[2] D. Venable, *The k−Factor: A New Mathematical Tool for Stability Analysis and Synthesis*, Proceedings of Powercon 10, 1983, pp. 1–12

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