AN2623
Application note

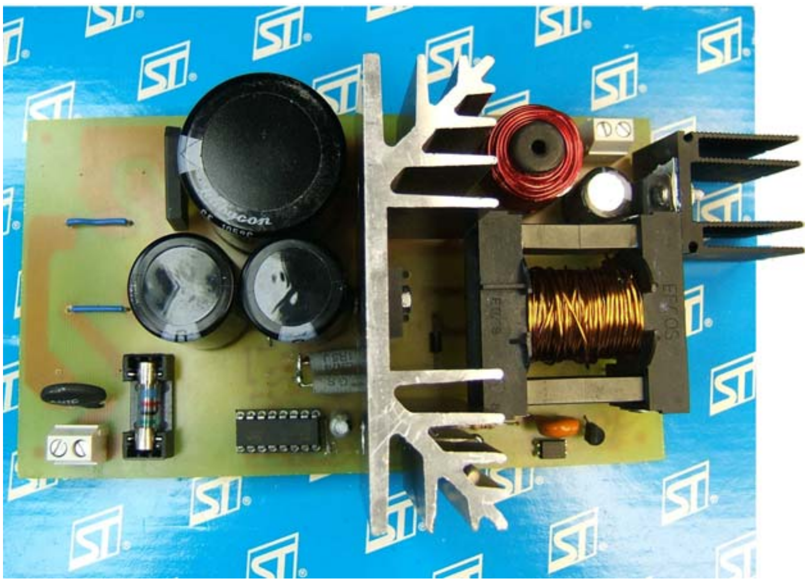
Evaluation board
for off-line forward converter based on L5991

Introduction

This application note gives a practical example of a 160 W, isolated, forward converter using the L5991, high frequency current mode PWM controller. Design procedures for both the power stage and controller are presented.

Generally for this power level the norm ICE61000-3-2 imposes the use of a PFC pre-regulator stage, but some countries do not require compliance to this norm. The forward converter presented here does not have a PFC.

Figure 1. 160 W off-line forward converter, evaluation board



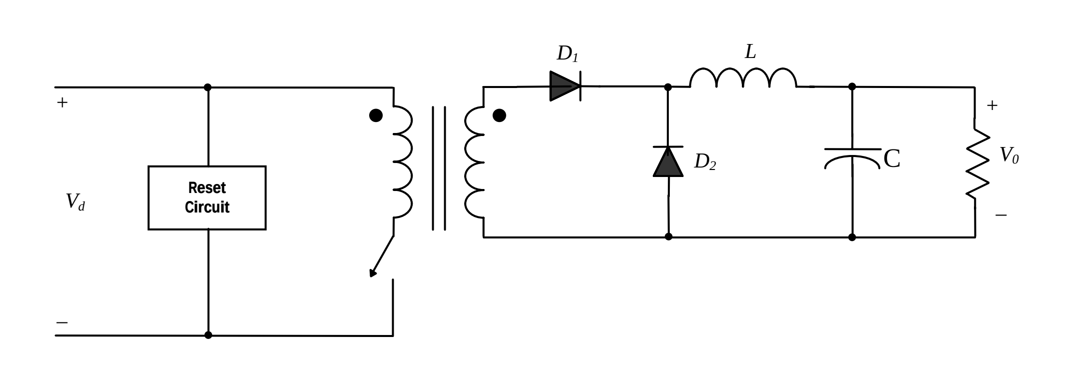
Contents

List of figures

1 Basis of forward topology

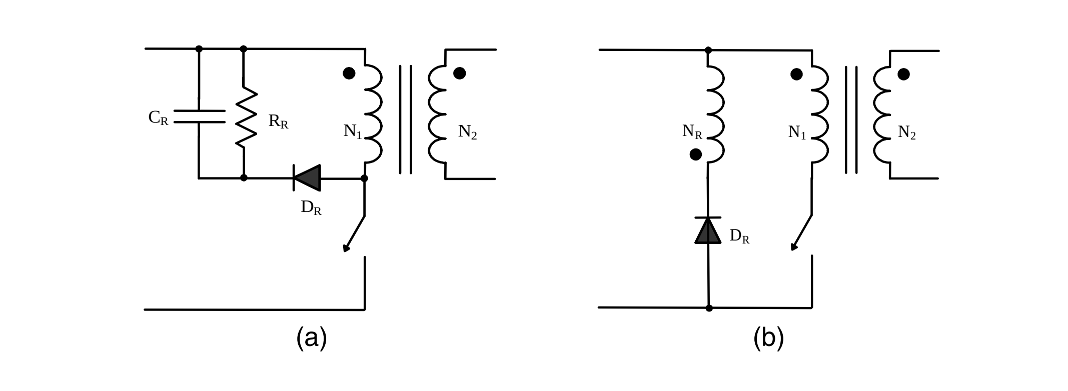
A forward converter is typically used in off-line applications in the 100 W - 300 W power range. A simplified schematic of the forward converter can be seen in *Figure 2*.

Figure 2. Basic forward converter topology



A natural limitation of the forward converter is the need to completely reset the transformer, cycle by cycle, before the next MOSFET switches on. Different circuits are used for this purpose with advantages and drawbacks. The two simplest and most commonly used reset schemes are: the RCD reset circuit and the reset auxiliary winding both shown in *Figure 3* (a-b). In the design presented in this document, the reset winding was used. It is advantageous with respect to efficiency because the energy stored in the magnetizing inductor goes back to the input and is not lost as using an RCD snubber net. The drawback of the reset circuit is that, generally, a higher voltage Power Mosfet is needed. In the present design a 900 V MOSFET was used.

Figure 3. Reset circuits



The primary controller IC used is the L5991. It is based on a standard current mode PWM controller and includes features such as programmable soft start, adjustable duty cycle limitation and a standby function that reduces the switching frequency when the converter is lightly loaded. The standby function, in this case, is not used to prevent the transformer from saturation. The output voltage regulation is obtained through a voltage reference and an error amplifier (TL1431) placed at the secondary side. A charge pump connected to an auxiliary winding guarantees a stable supply at the controller itself.

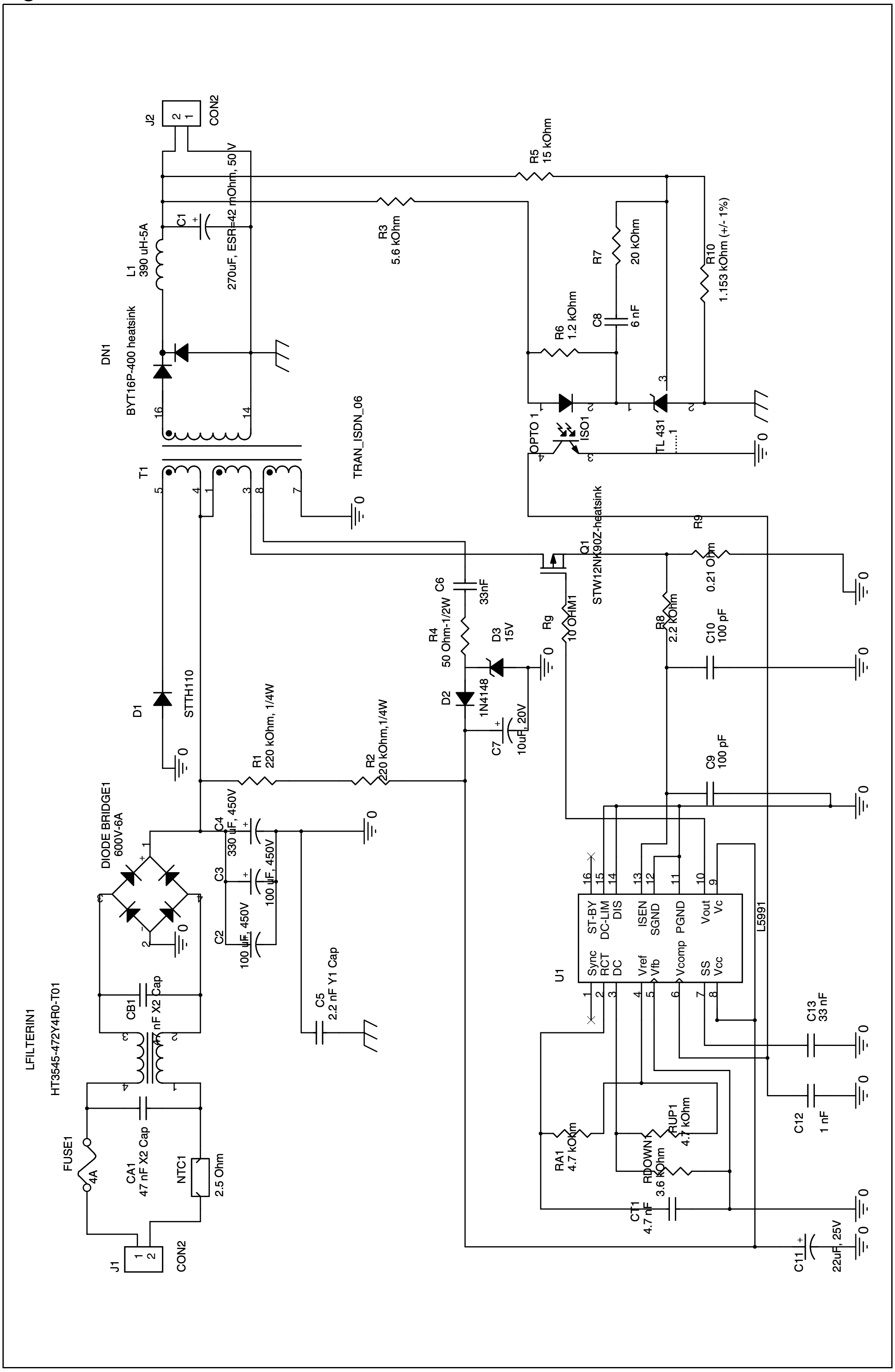
2 Main characteristics

The design procedure is presented in this section and we will refer to the electrical schematic in *Figure 4*. The power supply electrical specifications are shown in *Table 1* below.

Table 1. Input and output parameters

Input parameters		
V_{in}	Input voltage	88 ÷ 290 V_{RMS}
f_{line}	Line frequency	50/60 Hz
Output parameters		
V_{out}	Output voltage	35 V
I_{out}	Output current	4.5 A max continuous, 0.45 A min
P_{out}	Output power	160 W max
	Efficiency at full load	80%
$\Delta V_{out}\%$	Max tolerance on output voltage	3%
$\Delta V_{out\ HF}$	Max output voltage ripple at switching frequency	350 mV
$T_{A\ max}$	Maximum ambient temperature	70 °C

Figure 4. Electrical schematic



3 Design circuit

This section describes the design of the major parts of the circuit.

3.1 Primary controller: L5991

As previously stated, the L5991 is used as the primary controller and its components must first be selected. Refer to the L5991 datasheet for the choice of the two resistors (R_A, R_B) and one capacitor (C_T) which allows setting separately the operating frequency of the oscillator in normal operation (f_{osc}) and in standby mode (f_{sb}). In this application, it was established that the device must work at the unique frequency (in this case $R_B \rightarrow \infty$) of 60 kHz in normal and in standby operation. This frequency is calculated using R_A in the following formula:

Equation 1

$$f_{osc} = \frac{1}{C_T \cdot (0.693 \cdot R_A + K_T)}$$

where K_T=160 Ω and C_T is calculated fixing the discharge oscillator capacitor time T_d=5%T_{sw}

Equation 2

$$T_d = 30 \cdot 10^{-9} + K_t \cdot C_t \Rightarrow C_t = 4.7\,\text{nF},\ R_A = 5.6\,\text{k}\Omega$$

Establishing a D_{max} = 50%, L5991 allows obtaining this last value in two different ways. The method that allows implementing the slope compensation, if needed, was used.

The duty cycle limitation is obtained by applying the following voltage to pin3 :

Equation 3

$$V_3 = 5 - 2^{(2 - D_{max})} \Rightarrow V_3 = 2.17\,\text{V}$$

fixing (refer to *Figure 4*) R_{up}=4.70 kΩ, we can then immediately calculate R_{down}=3.60 kΩ

3.2 Output filter

Admitting a max current ripple on the inductor ΔI_{Lout} equal to 20% of I_{outMAX}, it is necessary to select an inductor value according to *Equation 4*:

Equation 4

$$L_{out} = \frac{V_{2min} - V_{diode} - V_{out}}{\Delta I_{out}} \cdot \frac{D_{max}}{f_{sw}} \Rightarrow L_{out} = 342\ \mu\text{H}$$

The RMS (root mean square) current through the inductor is given by *Equation 5*:

Equation 5

$$I_{RMS-Lout} = \sqrt{I^2_{out} + \frac{\Delta^2_{out}}{12}} \Rightarrow I_{RMS-Lout} = 4.58\ \text{A}$$

The peak current through the inductor is:

Equation 6

$$I_{Peak-Lout} = I_{out} + \Delta I_{Lout} = I_{Peak-Lout} = 5.4\ A$$

According to these results, L_{out} was chosen as the Coil Craft's inductor PCV-1-394-05L whose inductance value is L_{out}=390 µH.

According to the max high frequency voltage ripple (Δ_{VoutHF}=350 mV) from the electrical specifications, the necessary minimum capacitor value (C_1 in the *Figure 4*) and its maximum admitted ESR (Equivalent Series Resistance) are calculated as follows:

Equation 7

$$C_{outmin} = \frac{V_{out}}{\Delta V_{outHF}} \cdot \frac{1}{8 \cdot f^2_{sw}} \cdot \frac{1 - D_{max}}{L_{out}} \Rightarrow C_{outmin} = 4.5\ \mu F$$

Equation 8

$$ESR_{max} = \frac{\Delta V_{outHF}}{\Delta I_{out}} \Rightarrow ESR_{max} = 388\ m\Omega$$

The RMS current through the output capacitor must not exceed the current rate of the selected capacitor and is calculated as:

Equation 9

$$I_{RMS-Cout} = \sqrt{I^2_{RMS-Lout} - I^2_{out}} \Rightarrow I_{RMS-Cout} = 860\ mA$$

According to these requirements a C_{out}=C_1=270 µF (capacitance value) 63 V (Voltage rate) ZL series Rubycon electrolytic capacitor was selected with an ESR of 42 mΩ and max current capability of 1495 mA.

3.3 Output diodes

The maximum reverse voltages across the rectifier diode and the free wheeling diode (D1-D2 in the *Figure 2*) can be calculated as:

Equation 10

$$V_{diodeR} = \frac{V_{1max}}{n} - V_{dropF} \Rightarrow V_{diodeR} = 328V$$

Equation 11

$$V_{diodeF} = \frac{V_{1max}}{n} - V_{dropR} \Rightarrow V_{diodeR}$$

V_{dropF} and V_{dropR} are, respectively, the voltage drop in the freewheeling diode and in the rectifier diode, when they are forward biased, and n=1.25 is the turn ratio between the primary and the secondary winding of the transformer. Considering that the voltage drops in the two diodes are the same, we can conclude from *Equation 10* that $V_{diodeR} = V_{diodeF}$.

The maximum RMS and the average currents through the rectifier diode are calculated as:

Equation 12

$$I_{RMSdiodeR} = I_{out} \cdot \sqrt{D_{max}} \cdot \sqrt{\left(1 + \frac{1}{12} \cdot \left(\frac{\Delta I_{out}}{I_{out}}\right)^2\right)} \Rightarrow I_{RMSdiodeR} = 3.2\text{ A}$$

$$I_{AVGdiodeR} = I_{out} \cdot D_{max} = 2.25\text{ A}$$

and for the free wheeling diode:

Equation 13

$$D_{min} = D_{max} \cdot \frac{V_{dcmin}}{V_{dcmax}} = 11.5\% \Rightarrow I_{RMSdiodeF} = I_{out} \cdot \sqrt{(1 - D_{min})} \cdot \sqrt{\left(1 + \frac{1}{12} \cdot \left(\frac{\Delta I_{out}}{I_{out}}\right)^2\right)} = 4.23\text{A}$$

Equation 14

$$I_{AVGdiodeF} = I_{out} \cdot (1 - D_{min}) = 3.825\text{ A}$$

In *Equation 12*, *13*, and *14* the currents are calculated in Full Load condition considering the worst case for each diode and can be used to calculate the maximum power dissipation for each diode.

To reduce the number of components, the size of the board, and to minimize power losses, the ST double fast recovery rectifier BYT16P-400 was selected.

Although the two diodes inside the same package are always working complementarily, in order to choose the heat sink, the total power losses in the worst case can be calculated as if, instead of two diodes there is only one that flows through the whole current of the inductor:

Equation 15

$$P_{totDiode} = V_t \cdot I_{out} + R_d \cdot I^2{}_{out} \cdot \left(1 + \frac{1}{12} \cdot \left(\frac{\Delta I_{out}}{I_{out}}\right)^2\right) = 5.53\text{ W}$$

Considering T_{AmbMax} = 70 °C, the power losses just calculated, and the maximum junction temperature T_{Jmax} (see diode datasheets) of the selected diode, it is possible to determine the total thermal resistance of the diode:

Equation 16

$$R_{thmax} = \frac{T_{jmax} - T_{AmbMax}}{P_{lossesR}} = 15°\text{ C/W}$$

The R_{thmax} that results is lower than the max junction to ambient thermal resistance $R_{thJ\text{-}A\,max}$ of the select diode, so a heat sink with thermal resistance $R_{thSN} \cong$ 13 °C must be added.

3.4 Power transformer design and MOSFET choice

Ideally in a forward converter, the energy flows forward from the primary side to the secondary side without any storage in the transformer. But the real transformer does not have infinite magnetizing inductance, so during the on-time of the power MOSFET some energy is stored in the magnetic core. The proper magnetic core and the primary winding turn number have to be selected in order to avoid core saturation. The proper magnetic core and primary winding turn number must be selected taking into account that some energy is also dissipated in the magnetic core.

An empirical formula that gives an indication regarding the needed area Product for magnetic core that has to be selected for the application is shown in *Equation 17*:

Equation 17

$$AP_{min} = \left(\frac{11.1 \cdot P_i}{0.141 \cdot \Delta B \cdot f_{sw}}\right) = 1.47\ cm^4$$

where ΔB is maximum flux density swing in Tesla for normal operation and its typical value is within 0.2-0.3T in the case of the forward converter. This value has to be chosen in order to avoid saturation and to limit core losses; we chose 0.2T. The selected core is ETD39 ($AP=2.2cm^4$, $Ae=125\ mm^2$) in N27 material. Considering this kind of core and the transformer's max temperature rise ΔT_{max}=40 °C, the maximum allowed total power loss is:

Equation 18

$$P_{LOSTtrasfTOTAL} = \frac{\Delta T}{R_{thCORE}} = 2.5\ W$$

and the result of the relative max allowed core loss:

Equation 19

$$P_{fe} = \frac{2 \cdot P_{LOSTtrasfTOTAL}}{k_1 + 2} = 1.21\ W$$

(k_1 is the ΔB swing exponent relative at N27 material) so the real value of the maximum flux density swing is:

Equation 20

$$\Delta B = 2 \cdot \left(\frac{P_{LOSTtrasfTOTAL}}{V_e \cdot k_0 \cdot f^{k_2}{}_{sw}}\right)^{\frac{1}{k_1}} = 0.146\ T$$

(k_0 is the loss coefficient of N27 material, K_2 is the frequency exponent of N27 material and V_e is the effective volume of ETD39's core). The minimum primary turns is given by:

Equation 21

$$N_{1min} = \frac{V_{dcmin} \cdot D_{max}}{A_e \cdot f_{sw} \cdot \Delta B} = 42$$

and we chose N_1=42.

The turn ratio *n* between primary and secondary side is defined as:

Equation 22

$$n = \frac{N_1}{N_2} = \frac{V_{dcmin}}{V_{2min}} = \frac{D_{max} \cdot V_{dcmin}}{V_{out} + V_{dropR}} = 1.15$$

where N_1 and N_2 are the number of turns of the primary and secondary side, V_{out} is the output voltage and V_{dropR} is the diode rectifier voltage drop. The secondary turns number is N_2=36. The magnetizing inductance of the primary side is given by:

Equation 23

$$L_m = A_L \cdot N^2{}_1 \cdot 10^{-9} \Rightarrow L_m \cong 3.8\ mH$$

where A_L is the inductance for turn square in $nH/turns^2$.

Considering that the total instantaneous current at the primary side is

Equation 24

$$i_{1tot}(t) = i'_2(t) + i_m(t)$$

where $i'_2(t)$ is the secondary winding current during t_{on} reported at primary side and $i_m(t)$ is the magnetizing current.

The magnetizing current expression is:

Equation 25

$$i_m(t) = \frac{V_{inmin}}{L_m} \cdot t$$

And its peak value is:

Equation 26

$$I_{mpk} = \frac{V_{inmin} \cdot f_{sw}}{L_m \cdot D_{MAX}}$$

The peak value for the $i'_2(t)$ is:

Equation 27

$$I'_{2pk} = \left(I_{out} + \frac{\Delta I_0}{2}\right) \cdot \frac{1}{n}$$

And the value for $i'_2(t)$ at switch-on is:

Equation 28

$$I'_{2min} = \left(I_{out} - \frac{\Delta I_0}{2}\right) \cdot \frac{1}{n}$$

The ripple current at the primary side is:

Equation 29

$$\Delta I_1 = I_{mpk} + I'_{2pk} - I'_{2min}$$

The rms value for the current at the primary side is:

Equation 30

$$I_{1totRMS} = \sqrt{D_{MAX} \cdot \left((I'_{2min})^2 + \Delta I_1 \cdot I'_{2min} + \frac{1}{3} \cdot \Delta I^2_1\right)}$$

In this case neglecting the $i_m(t)$, from (*Equation 24*), it is possible to write following formula:

Equation 31

$$I_{1totRMS} \cong \frac{I_{RMSdiodeR}}{n} = 2.75\ \text{A}$$

Considering (*Equation 18*) and (*Equation 19*), the maximum allowed copper power losses in the windings transformer can be immediately calculated. It is possible to select the diameter for primary and secondary winding; we have chosen d_1=0.25 mm and d_2=0.8 mm.

Considering that for the L5991 the maximum allowed voltage value at the current sense (I_S, pin n°13) is 1 V, it is possible to determine the value of the current sense resistor (R_9 in *Figure 4*):

Equation 32

$$R_9 = \frac{1}{I_{1totpeak}} = 0.23\ \Omega$$

where $I_{1totpeak}$ is the peak value of $i_{1tot}(t)$.

The maximum turn ratio k between primary and reset winding, as known in technical literature, in order to achieve the complete demagnetization of the transformer is the following:

Equation 33

$$k_{max} = \frac{1 - D_{max}}{D_{max}} \Rightarrow k_{max} = 1$$

Choosing

Equation 34

$$k = \frac{N_R}{N_1} = 0.96$$

the necessary number of turns for reset winding is N_R=41.The maximum reverse voltage and average current of reset diode are given by ($I_{AVE-1°magn}$ is the magnetizing average current of the primary side):

Equation 35

$$V_{REV-R} = V_{DCmax} \cdot k + V_{DCmax} = 806\ V$$

Equation 36

$$I_{AVE-R} = \frac{I_{AVE-1°magn}}{k} = 0.11\ A$$

The Bipolar ultrafast diode STTH110 was chosen. Concerning the MOSFET the maximum drain voltage is:

Equation 37

$$V_{drainMax} = (V_{dcMax} - V_{dropReset}) \cdot \left(\frac{1}{k}\right) + V_{dcMax} = 838V$$

with $V_{dropReset}$ as the voltage drop in the reset diode. The max rms drain current is:

Equation 38

$$I_{drainRMS} = I_{tot1RMS}$$

so the Zener-protected SuperMESH Power Mosfet STW12NK90Z was chosen. Calculating the estimated total MOSFET power losses, it is easy to concludethat a substantial heatsink (around $R_{th} \leq 5$ °C/W) is necessary.

3.5 Feedback loop

Since current mode control is employed using the L5991 current mode controller, the power stage of the forward converter exhibits a single output pole due to the output capacitor and load combination, along with a zero due to the ESR of the output capacitor. The goal of the compensator is to achieve a slope of -20 db/decade for the closed loop gain, with a phase margin greater than 45 degrees at the crossover frequency. To achieve good dc regulation, a high low-frequency gain is another requirement for the compensator. For continuous conduction mode operation, the transfer function of the forward converter (power stage) is:

Equation 39

$$G_1(s) = G_{1o} \cdot \frac{\left(1 + \frac{s}{\omega_z}\right)}{\left(1 + \frac{s}{\omega_p}\right)}$$

where (referring to *Figure 4*)

- G_{1o} is the power block gain and results in $G_{1o} = \frac{n \cdot R_0}{3 \cdot R_9}$
- R_0 is the effective total load resistance of the controlled output defined as $R_0 = \frac{V^2_{out}}{P_0}$
- R_9 is the current sense resistance
- *n* is the turn ratio between the primary and secondary side
- $\omega_z = \frac{1}{ESR_{cout} \cdot C_{out}}$
- $\omega_p = \frac{1}{R_0 \cdot C_{out}}$

In order to reach the objective previously stated at the beginning of this section, the feedback compensation network transfer function, using L5991, is obtained as:

Equation 40

$$C(s) = C_0 \cdot \frac{1 + \frac{s}{\omega_{zc}}}{1 + \frac{s}{\omega_{pc}}} \cdot \frac{1}{s}$$

where (referring to *Figure 4*)

- C_0 is the feedback block gain and results in $C_0 = \frac{12 \cdot 10^3 \cdot CTR}{R_5 \cdot C_8 \cdot R_3}$
- CTR is the current transfer ratio of the optocoupler
- R_5 is the upper resistance of the out voltage divider of feedback net
- R_3 is the polarization resistance of the optocoupler
- C_8 and R_7 are the capacitance and the resistance of the TL431's feedback net
- ω_{zc} is the zero of the feedback net $\Rightarrow \omega_{zc} = \frac{1}{R_7 \cdot C_8}$ to compensate ω_p
- ω_{pc} is the pole of the feedback net $\Rightarrow \omega_{pc} = \frac{1}{12 \cdot 10^3 \cdot C_{12}}$ to compensate ω_z
- C_{12} is the capacitor connected at COMP pin of L5991

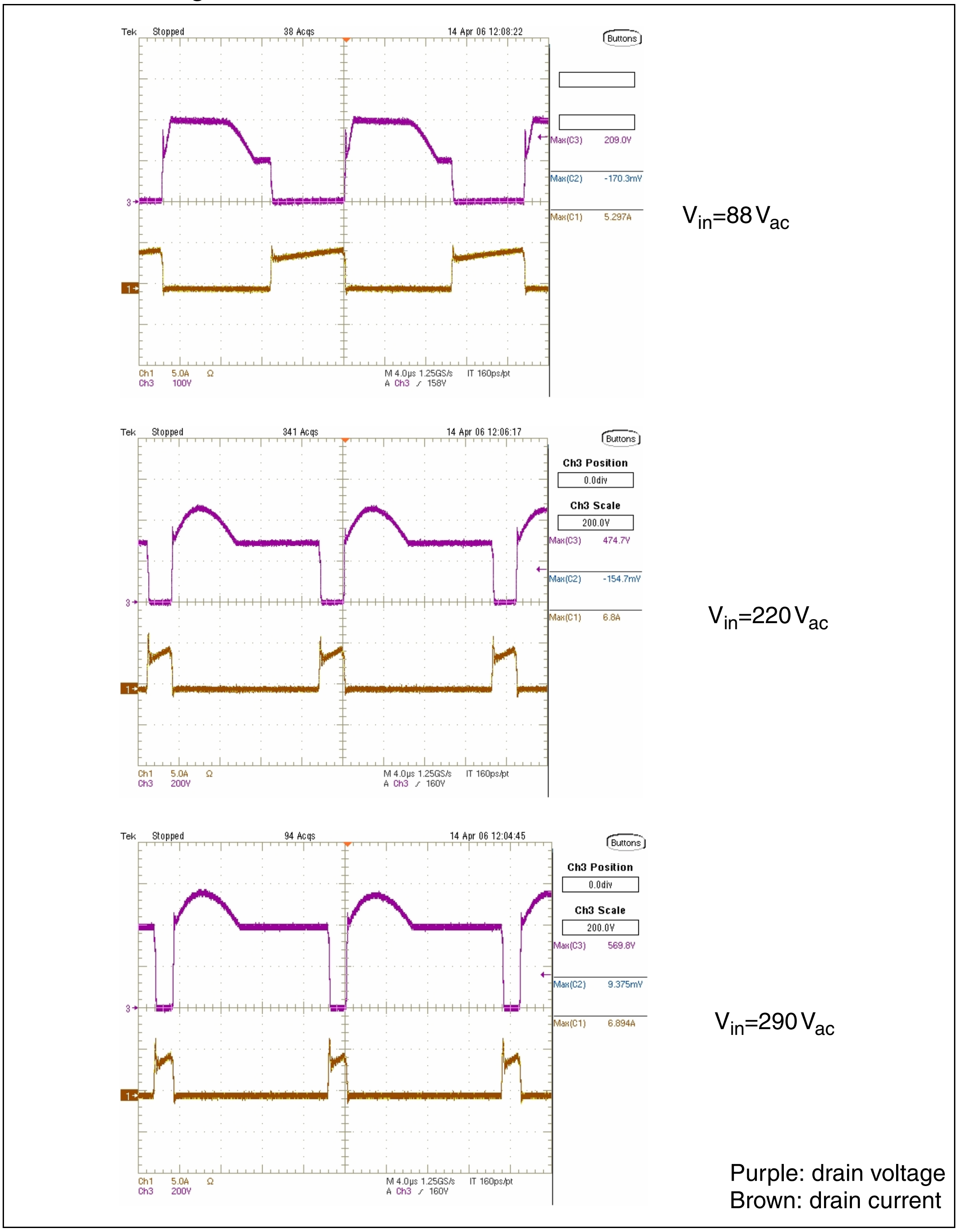
Choosing the crossover frequency $f_c \cong 0.1 \cdot f_{sw} \Rightarrow f_c = 5\ \text{kHz}$ it is possible to place compensator zero f_{zc} around $\frac{f_c}{3} \Rightarrow f_{zc} = 1600\ \text{Hz}$ and the compensator pole f_{pc} above $3 \cdot f_c \Rightarrow f_{pc} = 15\ \text{kHz}$.

Considering R_3=5.6 kΩ, R_5=15 kΩ, CTR=1 Ω were calculated, choose the following values: C_{12}=1 nF, C_8=6 nF, R_7=20 kΩ.

4 Experimental results

The schematic of the tested board is given in *Figure 4*. The graphs in *Figure 5* show the drain voltage and current at the minimum, nominal and maximum input mains voltage during nominal operation at full load.

Figure 5. V_{ds} and I_{ds} of STW12NK90Z in full load condition at different input voltages

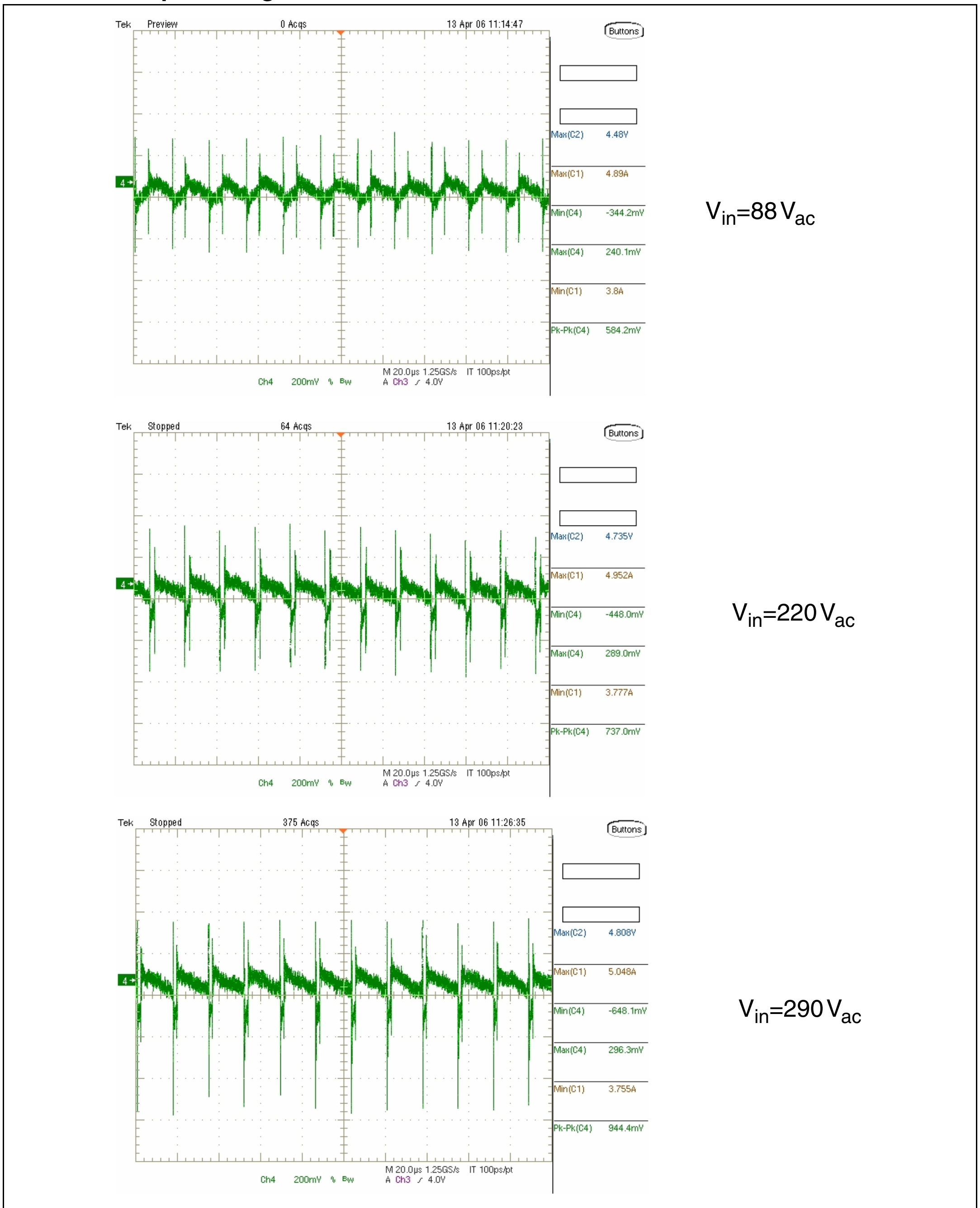


The drain peak voltage (570 V) assures a reliable operation of the STW12NK90Z with a good margin against the maximum B_{VDSS}.

4.1 High frequency ripple of output voltage and load regulation

Figure 6 shows the high frequency ripple of output voltage at minimum, nominal and maximum input voltages.

Figure 6. High frequency ripple of output voltage in full load condition at different input voltages



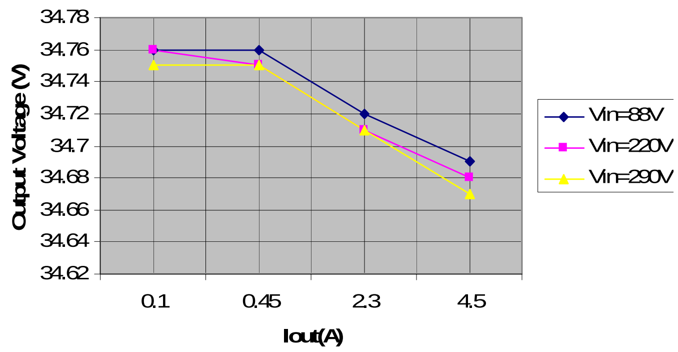
Apart the voltage spike, the voltage ripple of the output (at full load) for every input voltage is given in *Table 2*.

Table 2. Value of high frequency output ripple at full load condition

V_{in}(V)	V_{outHF} (mV)	V_{outHF}%
88	176	0.5
220	240	0.69
290	324	0.9

Figure 7 shows the behavior of the output voltage regulation against the load. It is easy to see from the graph that, changing the load, the output voltage is practically constant.

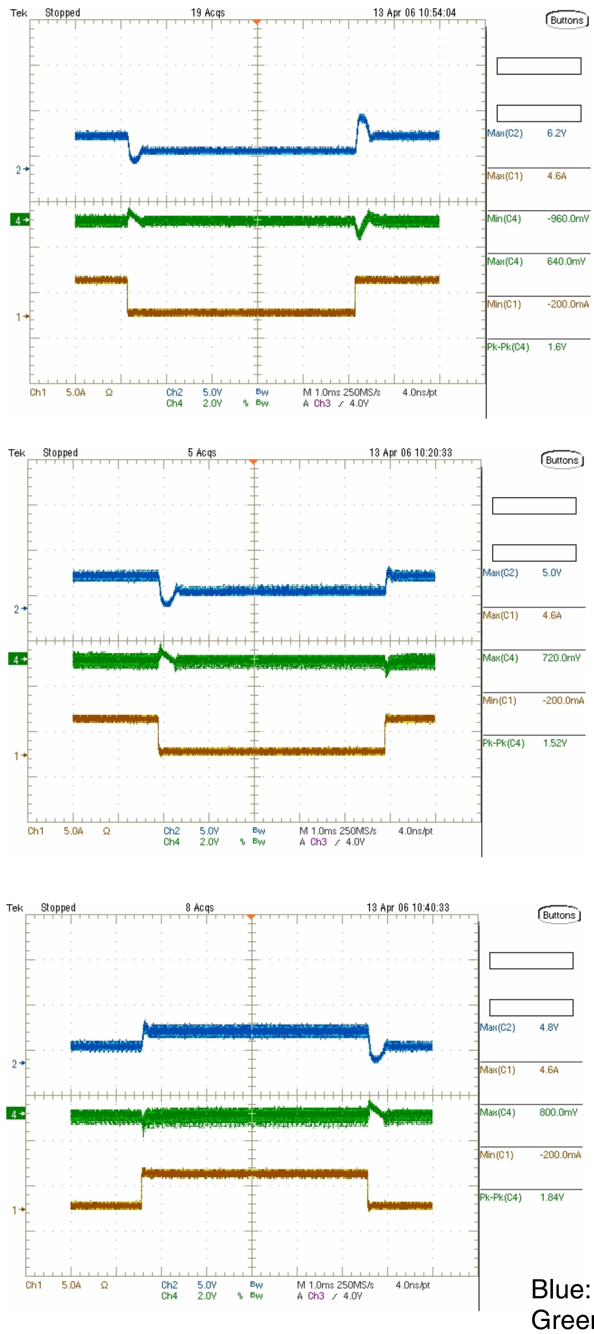
Figure 7. Output voltage behavior against the load and the V_{in}



4.2 Dynamic load test

The graphs in *Figure 8* show the output voltage regulation against a dynamic load variation (between max load and 10% max load).

Figure 8. Behavior of system under dynamic load at different input voltages



V_{in}=88 V_{ac}

V_{in}=220 V_{ac}

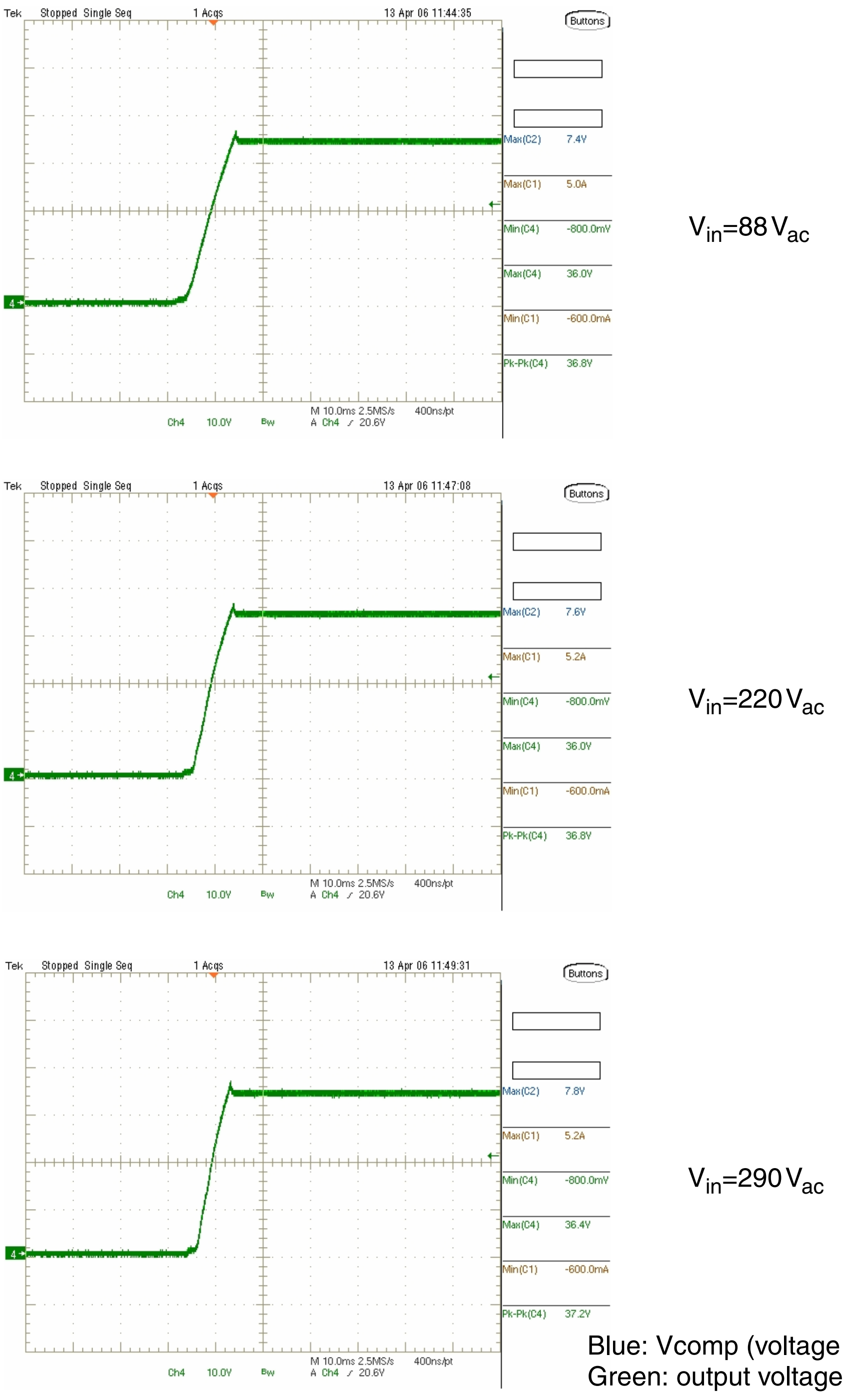
V_{in}=290 V_{ac}

Blue: Vcomp (voltage on pin 6 of L5991)
Green: output voltage
Brown: output current

4.3 Start-up behavior

Figure 9 shows rising slopes at full load of the output voltage at nominal, minimum and maximum input main voltages. As shown in the graphs, the rising times are fairly constant.

Figure 9. Behavior of system under dynamic load at different input voltages

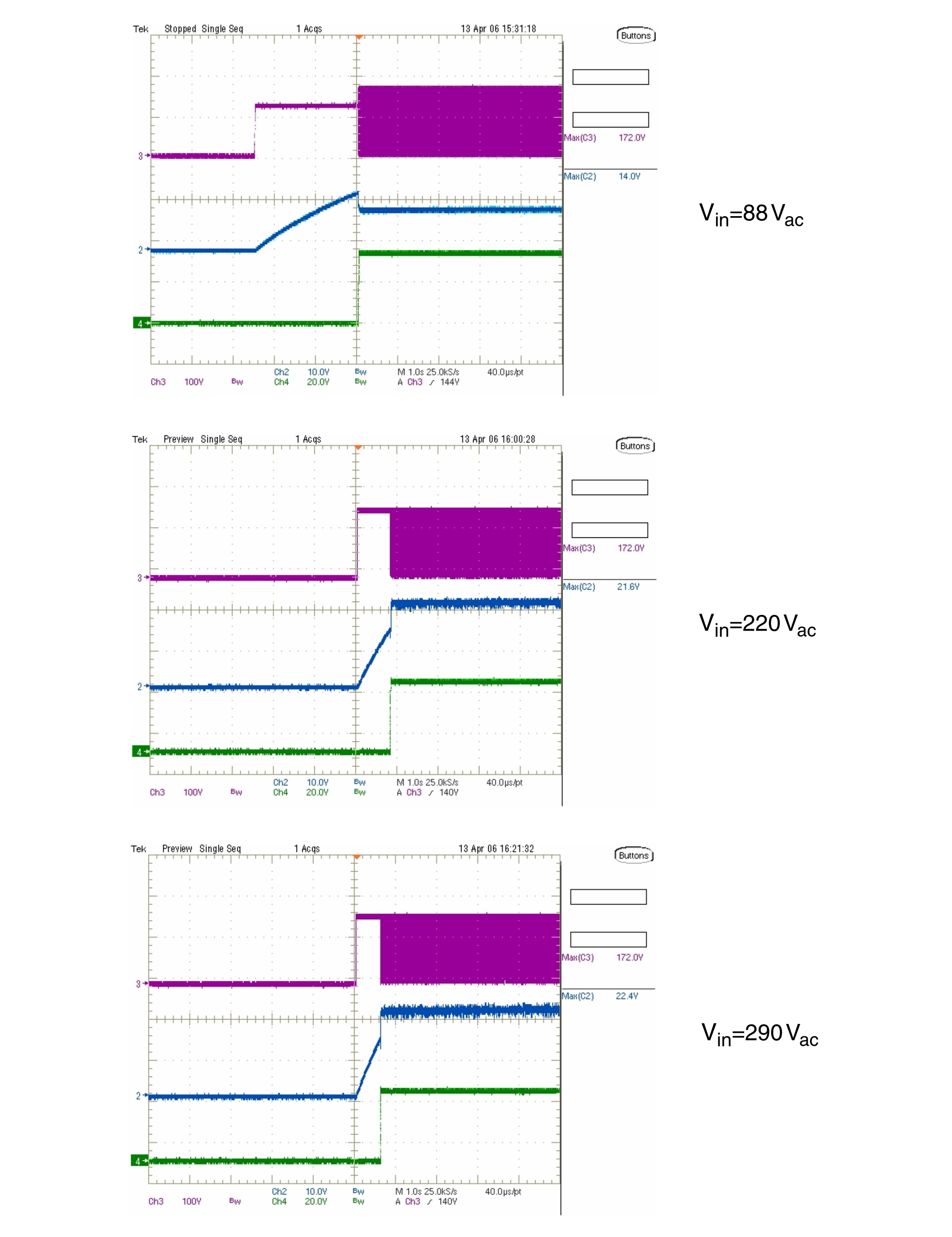


Blue: Vcomp (voltage on pin 6 of L5991)
Green: output voltage
Brown: output current

4.4 Wake-up time

Figure 10 shows the waveforms with wake-up time measures at nominal, minimum and maximum input voltages. Obviously due to the circuit characteristics, the wake-up time is not constant but it is dependent on the input voltage. The measured time at 88 V_{ac}, 220 V_{ac} and 290 V_{ac} are (respectively) 2.48 sec, 780 ms and 580 ms which are rather common values for this kind of power supply.

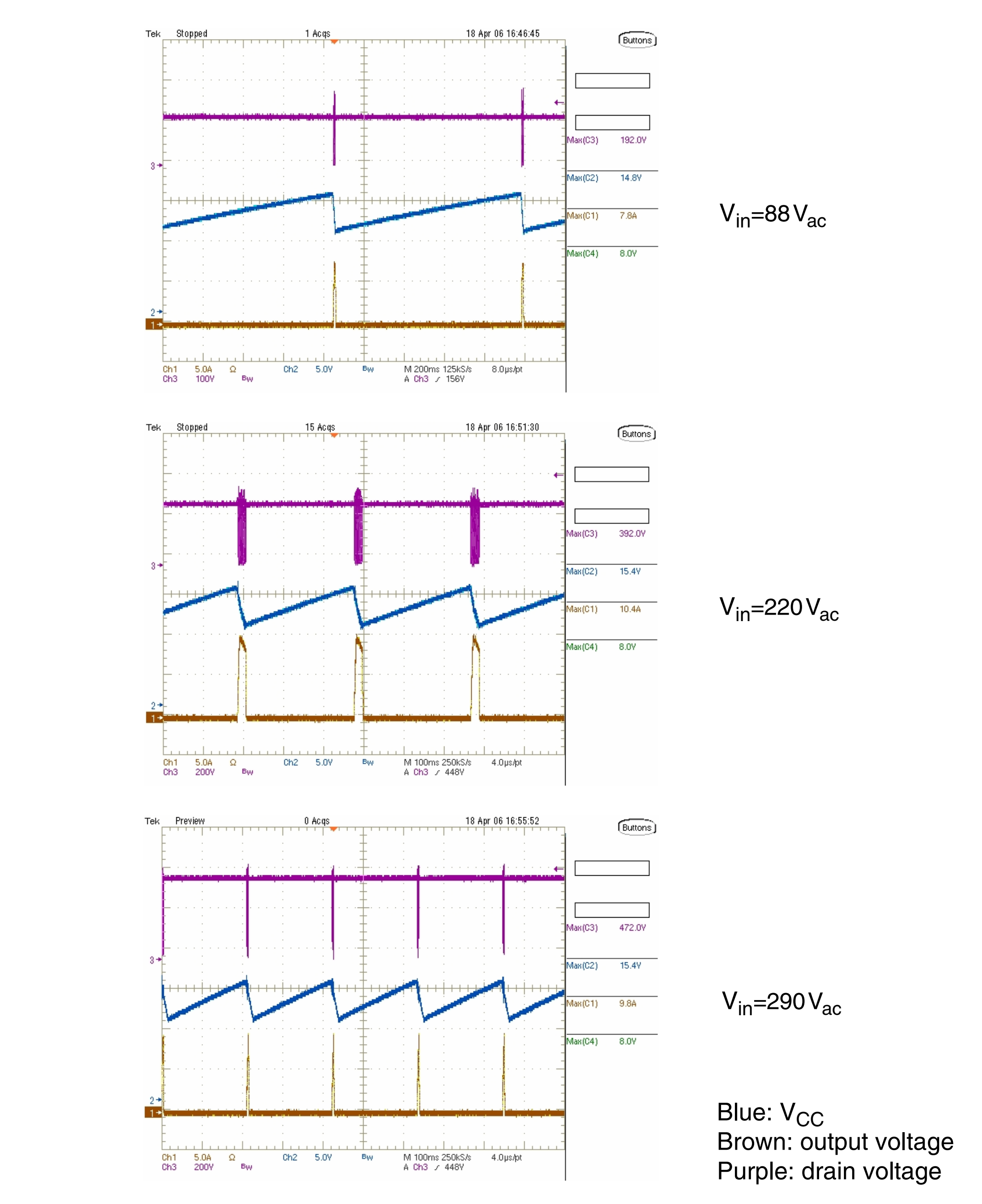
Figure 10. Wake-up time of the system at different input voltages



4.5 Short circuit test

All tests have been done at nominal, maximum and minimum input voltages. For all conditions the drain voltages are always below B_{VDSS}. As clearly indicated in the waveforms, the circuit starts to work in hiccup mode. Because the working time and the idle time are imposed by the charging and discharging time of the auxiliary capacitor C11 (refer to *Figure 4*), they are proportional to the input mains voltage.

Figure 11. Behavior of the system in short circuit condition at different input voltages



As expected the circuit protects itself as well.

4.6 Thermal measurement and global efficiency

One of the most critical parts of the power supply is the MOSFET. As previously seen at the end of *Section 3.4: Power transformer design and MOSFET choice* a heatsink is necessary. To verify the correct thermal behavior of the MOSFET, it was checked at maximum load and maximum input voltage. The device reaches the thermal steady state of 94 °C.

Figure 12 shows the global efficiency in function of the input voltage for two values of load. From the graph, we can conclude that the board has good efficiency. In absolute terms the minimum value is around 80% for V_{in}=290 V and the maximum is around 86% for V_{in}=88 V. The technical requirements for this converter have been respected.

Figure 12. Efficiency of the system

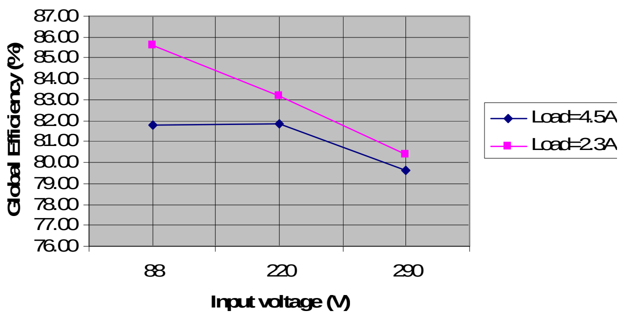


Table 3. Bill of material

Item	Quantity	Reference	Properties
1	1	1	TL 431 STMicroelectronics part
2	2	CB1	47 nF X2 Cap
		CA1	47 nF X2 Cap
3	1	CT1	4.7 nF
4	1	C1	270 µF, ESR=42 mΩ, 50 V, electrolytic capacitor
5	2	C2	100 µF, 450 V, electrolytic capacitor
		C3	100 µF, 450 V, electrolytic capacitor
6	1	C4	330 µF, 450 V, electrolytic capacitor
7	1	C5	2.2 nF, Y1 Cap
8	1	C6	33 nF
9	1	C7	10 µF, 20 V electrolytic capacitor
10	1	C8	6 nF
11	2	C9	100 pF ceramic capacitor
		C10	100 pF ceramic capacitor
12	1	C11	22 µF, 25 V, electrolytic capacitor
13	1	C12	1 nF ceramic capacitor

Table 3. Bill of material (continued)

Item	Quantity	Reference	Properties
14	1	C13	33 nF ceramic capacitor
15	1	Diode bridge1	600 V - 6 A
16	1	DN1	BYT16P-400 STMicroelectronics part + heatsink
17	1	D1	STTH110 STMicroelectronics part
18	1	D2	Diode 1N4148
19	1	D3	Diode Zener Vz=15 V
20	1	FUSE1	4 A
21	1	ISO1	OPTO 1-PC817
22	2	J1	CON2
		J2	CON2
23	1	LFILTERIN1	Common Choke, 15 mA
24	1	L1	Inductor, 390 µH-5A
25	1	NTC1	2.5 Ω
26	1	Q1	STW12NK90Z - STMicroelectronics part + heatsink
27	2	RUP1	4.7 kΩ
		RA1	4.7 kΩ
28	1	RDOWN1	3.6 kΩ
29	1	R1	220 kΩ, 1/4 W
30	1	R2	220 kΩ,1/4 W
31	1	R3	5.6 kΩ
32	1	R4	50 Ω-1/2 W
33	1	R5	15 kΩ
34	1	R6	1.2 kΩ
35	1	R7	20 kΩ
36	1	R8	2.2 kΩ
37	1	R9	0.21 Ω
38	1	R10	1.153 kΩ
39	1	T1	Transformer
40	1	U1	L5991 STMicroelectronics part
41	1	Rg	10 Ω

5 Revision history

Table 4. Document revision history

Date	Revision	Changes
29-Oct-2007	1	Initial release