AND8257/D

Implementing a Medium Power AC-DC Converter with the NCP1395

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APPLICATION NOTE

INTRODUCTION

This document describes all the necessary design steps that need to be evaluated when designing the NCP1395 controller in an LLC resonant converter topology. A 240 W AC−DC converter has been selected for the typical application.

The design requirements for our 240 W AC−DC converter example are as follows:

Requirement	Min	Max	Unit
Input Voltage	90	265	Vac
Output Voltage	−	24	Vdc
Output Power	0	240	W
Operating Frequency	65	125	kHz
Efficiency Under Full Load	90	−	%
No Load Consumption	−	1000	mW

LLC series resonant converter topology has been selected to meet efficiency requirements. The NCP1395 resonant mode controller is a very attractive solution for such designs because it offers the following features.

Brownout Protection Input

The divided down input voltage of the converter is permanently monitored by the Brownout pin (pin name). If the voltage on the bulk capacitor falls outside of the desired operating range, the controller drive output will be shut off. This feature is necessary for an LLC topology because it is usually optimized to operate over a narrow range of bulk voltages.

Immediately Fault Input

This input can be used as a shutdown input in some applications (LCD television SMPS, etc.). It can also be used to induce skip mode operation of the LLC converter, during the light load conditions. Standby consumption of whole supply can thus be significantly decreased.

Timer Based Fault Protection

The converter stops operation after a programmed delay when this input is activated. This protection can be implemented as a cumulative or integrating characteristic. Thus under transient load conditions the converter output will not be turned off, unless the extreme load condition exceeds the timeout.

Internal Transconductance Amplifier

This internal transconductance amplifier can be used to create effective overload protection. As the result the power supply can be operated in either CV or CC mode. This feature is very useful for the battery chargers applications.

Common Collector Optocoupler Connection

The open collector output allows multiple inputs to the feedback pin, for example overcurrent sensing circuit, overtemperature sensor, etc. The additional input can pull up the feedback voltage level and take over the voltage feedback loop.

Please refer to the NCP1395A/B data sheet for a detailed description of all the functions.

Demo Board Connection Description

The schematic for the 240 W demo board is shown in Figure 1. The demo board contains three blocks: a PFC front stage (which is necessary for the required power level and to restrict the bulk voltage operating range of the downstream resonant converter), an LLC converter, and an auxiliary buck converter which provides bias power for PFC and LLC controllers.

The NCP1653A PFC controller is used to control the PFC front stage. Capacitors C_1−C_5 together with BALUNs L_1, L_2 and varistor VDR_1 forms the EMI filter which suppresses noise conducted to the mains.

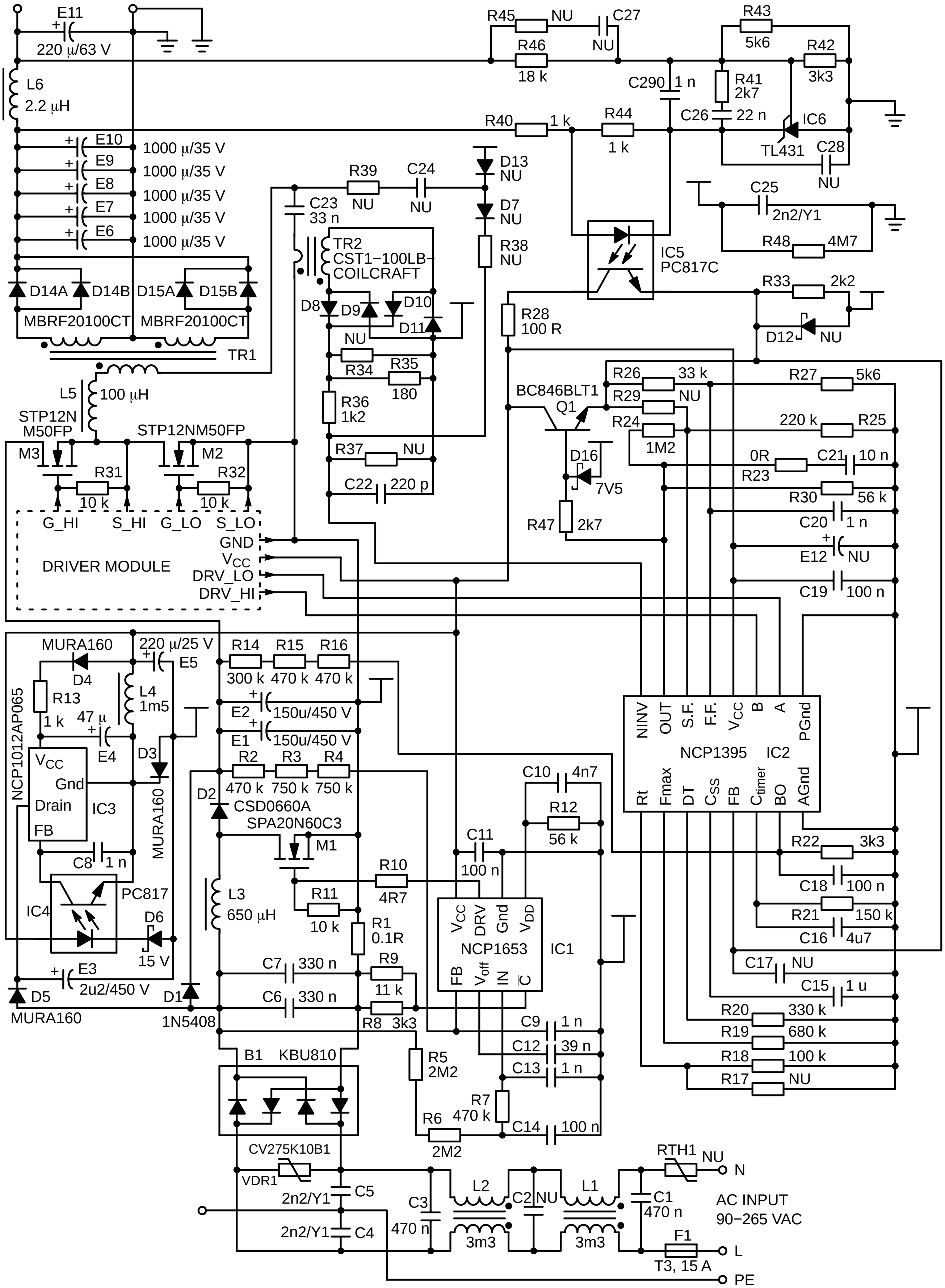

Figure 1. Schematic of the NCP1395 Demo Board

A bridge rectifier is used to rectify the input AC line voltage. Capacitors C_6 and C_7 filter the high frequency ripple current, which is generated by the PFC stage. In this application a Classical PFC boost topology is used. The PFC power stage is formed by: inductor L_3, MOSFET switch M_1, SiC diode D_2, bulk capacitors E_1, E_2 and inrush current bypassing diode D_1. The current in the PFC stage is monitored by current sense network R_1, $R_8//R_9$. The output voltage of the PFC stage is regulated to a nominal 400 Vdc via feedback loop components $R_2−R_4$ and C_9. Part of the rectified input voltage is taken by the divider $R_5−R_7$ and capacitors C_{13}, C_{14} to create over power protection of the PFC switch. Capacitor C_{12} filters the control voltage and sets the PFC feedback loop bandwidth. Devices connected to Pin 5 of the NCP1653A set the operation mode of the PFC stage to CCM and also dictate over power protection level. Please refer to the application note AND8184/D for detailed explanation how to design a PFC using the NCP1653A controller.

An ON Semiconductor NCP1012 monolithic switcher is used for the auxiliary buck converter. This converter provides a stable Vcc to assures proper operation of the PFC and LCC controller under all operating conditions and fault conditions, such as when a short−circuit is applied to the output of the LLC converter.

The NCP1012 is connected as a high side switch. Diode D_3 is the freewheeling diode. The input power for the buck converter is supplied from the rectified mains via diode D_5 and electrolytic capacitor E_3. Feedback is done via diode D_6, optocoupler IC_4 and capacitor C_8. Supply voltage of the NCP1012 is maintained on the capacitor E_4 using diode D_4, resistor R_{13} and internal DSS architecture. The internal dynamic self supply block is inactive during steady state operation decreasing the power consumption of the buck converter. The output voltage of the buck converter is regulated to 16 V allowing some margin above the PFC controller V_{ccON} level, which is 15 V maximum. Please refer to application note AND8191/D for other information regarding NCP10xx products.

As it was previously mentioned the NCP1395 resonant mode SMPS controller is used to control the LLC converter power stage, provide output voltage regulation, and fault protection.

The power stage of the LLC converter is formed by bulk capacitors E_1, E_2, MOSFETs M_2, M_3, resonant inductor L_5, transformer TR_1 and resonant capacitor C_{23}. A center tapped winding is used on the secondary side to increase efficiency of the converter. Electrolytic capacitors $E_6−E_{11}$ together with inductor L_6 serve as an output filter.

The output voltage is set at 24 Vdc using a TL431 (IC6) for feedback. The resistive divider formed by R_{46}, R_{42} and R_{43} provide a sample of the output voltage to the reference pin of the TL431 (2.5 V). The control loop feedback compensation is done with the series combination of capacitor C_{26} and resistor $R_{41.}$ The biasing current for the IC_6 is provided by the resistor R_{44}. The optocoupler IC current is set by the series resistor R_{40}.

The current through the feedback optocoupler is translated to a voltage on the primary side by the resistor $R_{33.}$ The voltage is then applied to the NCP1395 feedback pin. A Zener diode, D12, clamps the maximum feedback voltage and resistor R_{28} limits the current through D12.

The output power level at which the controller enters skip mode is set by the voltage divider R_{26}, R_{27}. The fast fault input is filtered by the capacitor C_{20}. Resistor R_{30} sets the voltage gain on the output of the operational transconductance amplifier, capacitor C_{21} is used for the current feedback loop compensation. The voltage on the OTA output is clamped to 7.5 V maximum with the resistor R_{47} and Zener diode D_{16}. The clamp is necessary because a higher voltage could cause the controller to enter skip mode during startup, or during overload. The output current level for which the skip mode takes place (during the overload conditions) is set by the divider R_{24}, R_{25}.

The primary current for the LLC power stage is sensed by current transformer TR_2 along with diodes $D_8−D_{11}$, resistors R_{35}, $R_{36,}$ and capacitor C_{22}. An alternative to the current transformer is to sense the primary current with the sensing circuit formed by the resistors R_{38}, R_{39} diodes D_7, D_{13} and capacitor C_{24}. This alternative is included in the demo board layout.

The minimum operating frequency of the converter is set by the resistor R_{18}, and the maximum operating frequency is set by resistor R_{19}. The dead time between the outputs A and B is set by the resistor R_{20}. The soft−start duration is set by capacitor C_{15}, and the timer duration is set by capacitor C_{16} together with resistor R_{21}.

The Brown Out circuit monitors the bulk capacitor voltage with the resistive divider set by R_{14}, R_{15}, R_{16}, $R_{22,}$ and capacitor C18. When the bulk capacitor voltage drops outside the desired operating range of the LCC converter, the output drives are turned off.

A decoupling capacitor, $C_{19,}$ is used between the ground and V_{cc} pin of the controller to improve the noise immunity.

The switches for the Half Bridge (M_2 and $M_{3)}$ are driven from a High Side Driver module which is mounted vertically on the converter's main board. This arrangement allows the designer to test the entire driver topology quickly and easily. Two versions of the driver are available as demo board accessories. Schematics for both versions are shown in Figures 2 and 3. One version uses the NCP5181 − integrated high voltage driver and is tailored for consumer applications where the price is important.

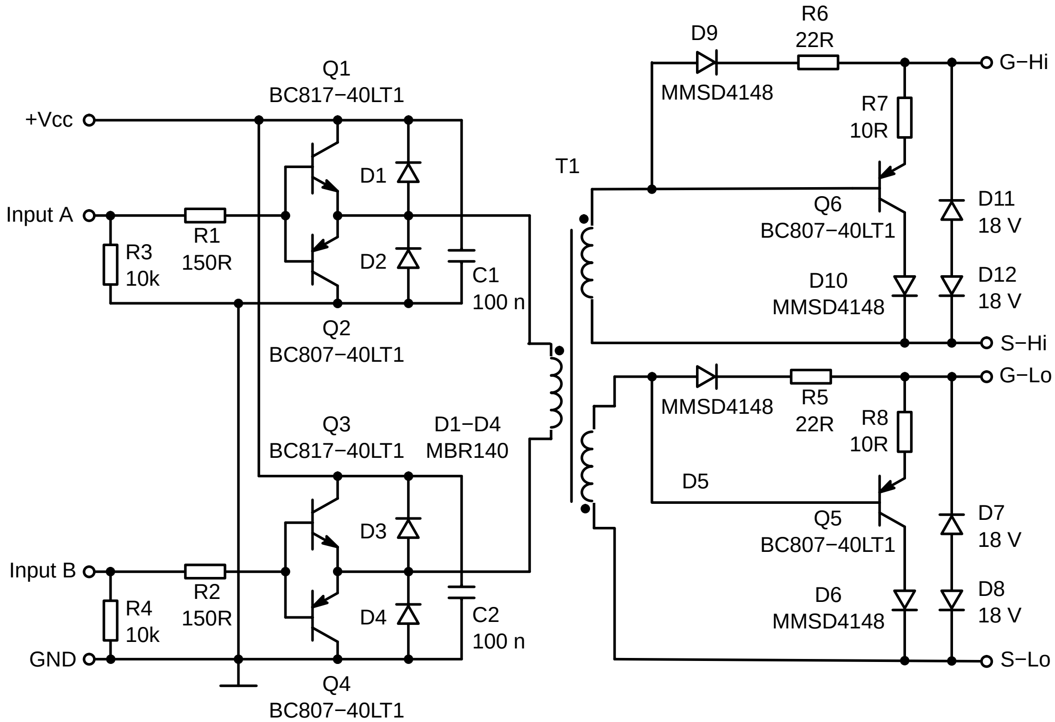

Figure 2. Connection of the MOSFET Driver with Transformer

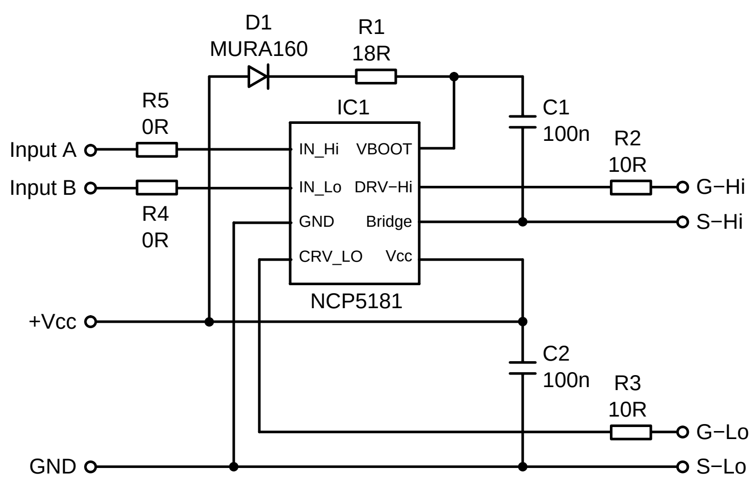

Figure 3. Connection of the MOSFET Driver with NCP5181 − Integrated Driver

Design steps for resonant tank components values are described in details below.

LLC Converter Stage Design

An LLC resonant converter is an attractive topology in comparison to a traditional half bridge for the following reasons:

- An LCC converter is capable of ZVS while operating over the entire set of anticipated load conditions.
  With ZVS the switches are turned on when its drain voltages are zero, the result is nearly zero turn−on losses and a reduction in the EMI signature. The classic half bridge topology, which uses hard switching, can have significant turn−on switching losses and this can increase the EMI signature.
- Low turn−off current: switches are turned off under low current and thus the turn−-off losses are also lowered in comparison to a classical half bridge topology.
- Zero current turn−off of the secondary diodes: when the converter is operating under full load condition, the output rectifiers are turned of under zero current which results in an reduced EMI signature.
- Same component count as half bridge topology: The component count is nearly the same between an LLC converter and a classical HB configuration.

Disadvantages of the LLC converter lie in these features:

- Higher peak and RMS currents in the primary and secondary windings in comparison to the classical HB topology. Thus this topology isn’t very attractive for very high output current levels.
- Narrow input voltage operating range: LLC converter has to be optimized for narrow input voltage range, if one wants to take full advantage of the topologies benefits.
- Changeable operating frequency: operating frequency of the LLC converter has to vary to keep the output regulated.

LLC Resonant Converter Operation Description

The LLC resonant converter power stage is shown in Figure 4.

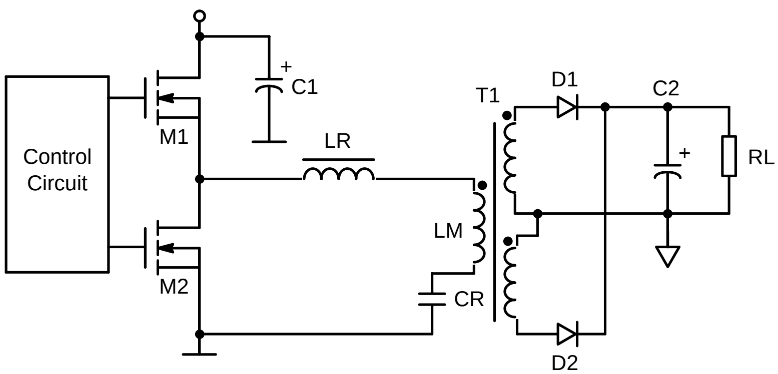

Figure 4. Power Stage of the LLC Resonant Converter

One can observe three resonant components in this topology: L_R–resonant inductance, L_M–magnetizing inductance of the transformer and C_R− resonant capacitor. We can define two different resonant frequencies using Thompson’s Equations 1 and 2:

$$f_{r1} = \frac{1}{2 \cdot \pi \cdot \sqrt{Lr \cdot Cr}} \qquad \text{(eq. 1)}$$

$$f_{r2} = \frac{1}{2 \cdot \pi \cdot \sqrt{(Lr + Lm) \cdot Cr}} \qquad \text{(eq. 2)}$$

This topology behaves like a frequency dependent divider which is shown in the Figure 5 schematic.

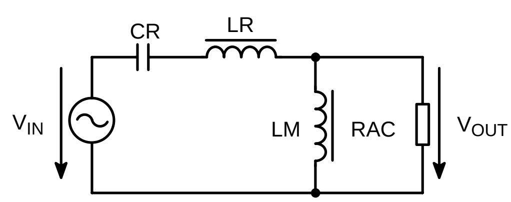

Figure 5. Substitutive Schematic of the LLC Resonant Converter

One can find the gain transfer function of the divider if the fundamental analysis is used [4], [5]. The main idea of this analysis is that only the fundamental frequency is passed through the resonant tank. As a result of this simplification, the real loading resistance needs to be converted to equivalent loading resistance R_{ac} using Equation 3.

$$R_{ac} = \frac{8 \cdot R_L}{\pi^2 \cdot n^2 \cdot \eta} \qquad \text{(eq. 3)}$$

Where:

R_L is the real loading resistance

n is the transformer turns ratio

η is expected efficiency

Using an equivalent load resistance (R_{ac}) we can calculate the gain transfer characteristic of the LLC converter and obtain the characteristic for any given value of load resistance, resonant tank components, and quality factor of the resonant circuit. The fastest way to do it is to use SPICE simulator. The result of such simulation can be seen in Figure 6.

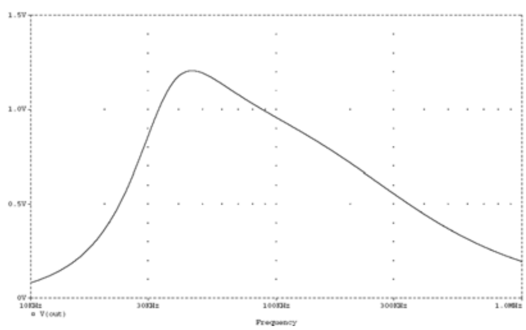

Figure 6. Typical Gain Characteristic of the LLC Resonant Converter

The load will change in the real application so it is necessary to plot the characteristic for several different load conditions in one graph. We can use parametric analysis to produce Figure 7.

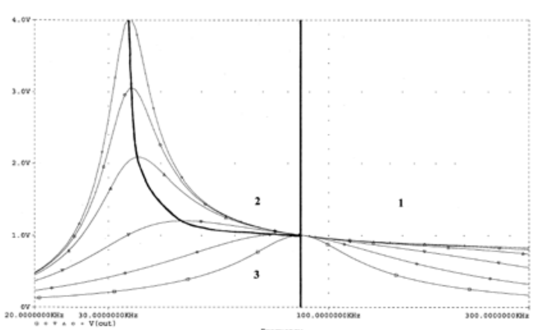

Figure 7. Couple of Gain Characteristic for Different Load Conditions

Three operating areas can be identified in these characteristics:

1. In this area (above the resonant frequency f_{r1}) the converter works as series resonant converter. The transformer magnetizing inductance never participates on the resonance because it is clamped by the converter output – one of the secondary diode is conducting for entire switching period. ZVS condition is naturally assured in this operation area for the entire load.
2. In this area the LLC converter works like a multi resonant converter. Let us go through one switching cycle (please refer to the timing diagram depicted in Figure 8).

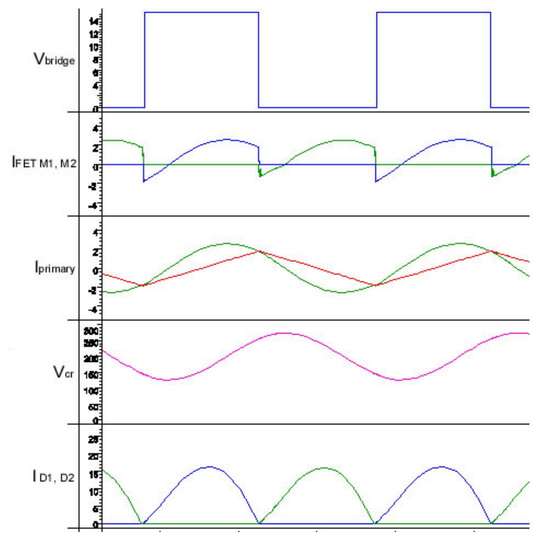

Figure 8. Typical waveforms of the LLC converter operating in the 2nd area of the gain characteristic.

One of the MOSFET switches is turned on. The resonant inductance Lr resonates together with the resonant capacitor C_r while the magnetizing inductance is clamped by the converter output – one of the secondary diode is conducting. When the resonant current decreases to a low value (the same as the magnetizing current) the output diode stop conducting and magnetizing inductance comes to play. The resonant circuit has thus been reconfigured to L_r+L_m – C_r. The resonant frequency f_{r2} is low in comparison to the resonant frequency f_{r1}. Thus the primary current slowly increases. This current stores energy in the magnetic components − mainly in the magnetizing inductance. Now the switching cycle is forced (by the driver) to be finished. The energy stored in the magnetic components causes ZVS for the opposite MOSFET switch using its body diode.

3. In this area the converter entering to the zero current switching mode. This could happen when converter is overloaded. Some protection circuit (passive or active) has to be implemented to prevent entering this area.

The big advantage of the LLC series resonant converter is the fact that the output can be regulated over the entire load range, while the change of the operating frequency is not as high as for other types of resonant converters.

Design of the resonant tank components i.e. L_r, L_m, and C_r is always compromise between maximum load changes, maximum acceptable operating frequency excursion, value of the circulating energy in the resonant circuit and short circuit characteristic. Behavior under short circuit is not an issue when some overcurrent protection is used – like with the NCP1395 controller.

The best way to operate an LLC resonant converter from the efficiency and EMI point of view is to let it work directly at the resonant frequency f_{r1}. Under these conditions the switching losses are minimized, circulating energy in the resonant tank is also low and secondary diodes are turned off under zero current so there are nearly no reverse recovery losses. This optimal operating point can be reached only for one given input voltage and load resistance value. Thus in the practice the LLC converter is usually designed using these operating conditions i.e. under series resonant frequency f_{r1} for full load and nominal bulk voltage (which is given by the PFC stage). When the load is decreased the operating frequency is increased by the feedback loop to keep the output voltage regulated. On the other hand converter has also to cope with the bulk voltage drops, due to transient loading (PFC regulation loop is very slow). Also hold up time requirements come into play. The converter will also operate under resonant frequency f_{r1} in these special cases. The minimum operating bulk voltage of the LLC converter can be effectively limited using the NCP1395 Brown Out input pin.

One important thing that has to be taken into account during the LLC designing is the fact that the manufacturing tolerances of the inductors and capacitors are pretty high for standard production. If one wants to have good repeatability for resonant converter design then higher accuracy of these components has to be specified or the LLC converter has to be designed with higher margins.

Transformer and Resonant Tank Components Design

Based on the mentioned recommendations we can start with the LLC resonant tank components design.

Input variables:

a) Input voltage range i.e. PFC output voltage: 350–420 VDC, 400 VDC nominal.

b) Output voltage: 24 VDC.

c) Max output current: 10 A continuous, overcurrent protection with 115% threshold.

d) Minimum output current: 0 A – skip mode has to be implemented to assure low power consumption under no load.

e) Operating frequency limits: from 65 kHz to 125 kHz.

f) Converter should work under series resonant frequency (f_{r1}) for 400 VDC nominal input voltage and full loaded output. This frequency should be below 100 kHz.

Based on the input requirement f) we can calculate turns ratio of the LLC transformer. Voltages on the resonant capacitor C_r and resonant inductor L_r are the same values but opposite orientation when converter works at resonant frequency f_{r1}. Thus the gain of the converter is given only by the transformer turns ratio value under this operating condition (assume that the leakage inductance of the transformer is low in comparison with magnetizing inductance). We can calculate needed turns ratio value using Equation 4:

$$n = \frac{N_p}{N_s} = \frac{V_{in}}{2 \cdot (V_{out} + V_f)} = \frac{400}{2 \cdot (24 + 0.8)} \approx 8 \quad \text{(eq. 4)}$$

Where:

N_p is the primary turns count

N_s is the secondary turns count

V_{in} is the input voltage (PFC output)

V_{out} is the wanted output voltage

V_f is the secondary diode voltage drop

ETD29 core has been selected for the transformer construction. The primary turns count can be calculated using Equation 5:

$$N_p = \frac{V_{in}}{8 \cdot \Delta B_{max} \cdot f_{swmin} \cdot A_e} = \frac{400}{8 \cdot 0.25 \cdot 65 \cdot 10^3 \cdot 76 \cdot 10^{-6}} \approx 40 \quad \text{(eq. 5)}$$

where:

ΔB_{max} is the maximum flux density excursion

f_{swmin} is the minimum operating frequency of the converter

A_e is the core effective area

The minimum switching frequency will be reached only in special cases − overload and bulk voltage dropouts. The flux density excursion will be always lower for normal steady state operation mode – hysteresis losses.

The secondary turns count can be calculated using Equation 6:

$$N_s = \frac{N_p}{n} = \frac{40}{8} = 5 \quad \text{(eq. 6)}$$

Needed copper area of the primary and secondary windings can be calculated based on the RMS current values.

Now we can calculate resonant components values. There are many variables that can be chosen by the designer, however, poor choices can result in converter efficiency degradation or too wide operating frequency range. It is necessary to take the following into account, during the design:

1. Quality factor of the resonant tank Q (and also its characteristic impedance) will significantly affect the gain characteristic and thus also the operating frequency range. If the Q of the resonant circuit is high, the gain characteristic will be narrow and the operating frequency range will be low. However, if the Q is too high the characteristic impedance will be low and converter operation under overload will be degraded.
2. The $L_m/L_r = k$ ratio will also significantly influence shape of the gain characteristics and thus the ZVS region borders [5], [6].

Based on the previous simulations and calculations Q of 3 and $k = L_m/L_r = 6$ ratio have been chosen.

Now we can calculate characteristic impedance Z_0 of the resonant circuit using Equation 7:

$$Z_0 = \frac{n^2 \cdot R_L}{Q} = \frac{8^2 \cdot 2.4}{3} = 51.2 \approx 51\ \Omega \qquad \text{(eq. 7)}$$

The resonant capacitor value can be calculated from Equation 8. Let us select the nominal operating frequency fop = fr1 = 85 kHz for full load and nominal bulk voltage:

$$C_r = \frac{1}{2 \cdot \pi \cdot f_r \cdot Z_0} = \frac{1}{2 \cdot 3.14 \cdot 85 \cdot 10^3 \cdot 51} = 36.7\ \text{nF} \qquad \text{(eq. 8)}$$

An E6 standard value capacitor of 33 nF has been chosen. The characteristic impedance will then change to 56.7 Ω and Q will be 2.71.

Now we can easily calculate the resonant inductance value, which is given by Equation 9:

$$L_r = Z_0{}^2 \cdot C_r = 56.7^2 \cdot 33 \cdot 10^{-9} = 106\ \mu\text{H} \qquad \text{(eq. 9)}$$

Use standard E6 value of $L_r = 100\ \mu$H.

The magnetizing inductance is given by the selected k ratio (Equation 10):

$$L_m = k \cdot L_r = 6 \cdot 100 \cdot 10^{-6} = 600\ \mu\text{H} \qquad \text{(eq. 10)}$$

Now we have nearly all needed for the simulation and obtain gain characteristic. However, for accurate results the leakage inductance of the transformer has to be taken into account. This inductance will affect on the resonance and it will also change gain of the converter somewhat. We have selected the ETD29 transformer core with classical winding technique so we can assume that the leakage inductance will be around 1% of the magnetizing inductance i.e. $L_{lk} = 6.0\ \mu$H. Let us now summarize real values of the resonant tank components:

$L_{r'} = L_r + L_{lk} = 106\ \mu$H

$L_m = 600\ \mu$H

$C_r = 33$ nF

$f_{op} = f_{r1} = 85.1$ kHz

Now we can start simulation of the resonant converter to obtain gain characteristic for full loaded converter. Simulation schematic, which includes the leakage inductance of the transformer, is depicted in Figure 9.

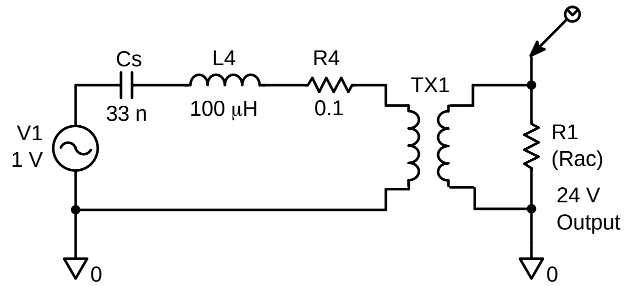

Figure 9. Simulation Schematic of the Proposed Converter

The gain values we are looking for can be calculated from Equations 11 through 13:

$$G_{min} = \frac{2 \cdot (V_{out} + V_f)}{V_{in\ max}} = \frac{2 \cdot (24 + 0.8)}{420} = 0.118 \qquad \text{(eq. 11)}$$

$$G_{nom} = \frac{2 \cdot (V_{out} + V_f)}{V_{innom}} = \frac{2 \cdot (24 + 0.8)}{400} = 0.124 \qquad \text{(eq. 12)}$$

$$G_{max} = \frac{2 \cdot (V_{out} + V_f)}{V_{in\ min}} = \frac{2 \cdot (24 + 0.8)}{350} = 0.142 \qquad \text{(eq. 13)}$$

Operating frequencies for these gains can be read from the simulated gain characteristic which is depicted in Figure 10.

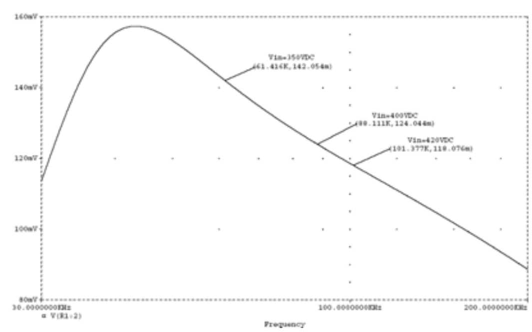

Figure 10. Simulated Gain Characteristic of Proposed LLC Converter – Full Load Conditions

As can be seen the operating frequency range is 61 kHz to 101 kHz, for a fully loaded converter output and with the input voltage range. The simulated operating frequency for the nominal input voltage and full load current is 88 kHz, which is very near to the calculated resonant frequency f_{r1}. One can see that there is enough gain margin to keep the converter in the ZVS region (Region 2 in Figure 7) even during a slight overload on the output.

When the load resistance is increased the gain of the converter will increase too, because of the Q changes. Moreover, the drops in the transformer copper and secondary diodes are lower for lower loading current. For these reasons the feedback loop will try to increase the operating frequency to lower the converter gain. If maximum operating frequency limit is set too low, the feedback loop will not be able to regulate the output. This is exactly what we need to assure skip mode for light load. Under such conditions the feedback pin voltage goes up and triggers the fast fault input via resistor divider R_{26}, R_{27}.

Simulation can again be used to find the maximum needed frequency for light load regulation. If the maximum operating frequency is limited to 125 kHz skip mode will be automatically implemented.

Let us summarize the simulations and design results:

1. Converter should operate at $f_{op} \approx fr_1 = 88$ kHz for full load and nominal bulk voltage.
2. Operating frequency range is 61 kHz to 101 kHz, at full load, over the input voltage range.
3. Operating frequency goes above fr1 frequency for light loads.
4. For very light loads, the converter will reach the maximum frequency limit and the feedback will activate fast fault input. Skip mode will thus take place.

Overcurrent Protection Circuitry

The Q of the resonant tank can drop to a very low value during overload conditions. To overcome incorrect operation in the ZCS region, the primary current has to be limited by the overcurrent protection circuit. This circuit has been already mentioned. The onboard OTA takes over and pulls up the feedback pin via transistor Q1 when input current goes over the desired maximum value.

Results Summarization

Operating frequency of the real demo board prototype is 83 kHz for full load and 395 VDC input voltage which is very close to the theoretical results.

Output current level for which the skip mode takes place has been set to 400 mA using resistor divider R26, R27. Skip level is given by the feedback voltage and thus it is very sensitive to the resonant component values and output voltage setup accuracy!

Measured efficiency for different input voltages and load conditions can be seen in Figures 11 and 12.

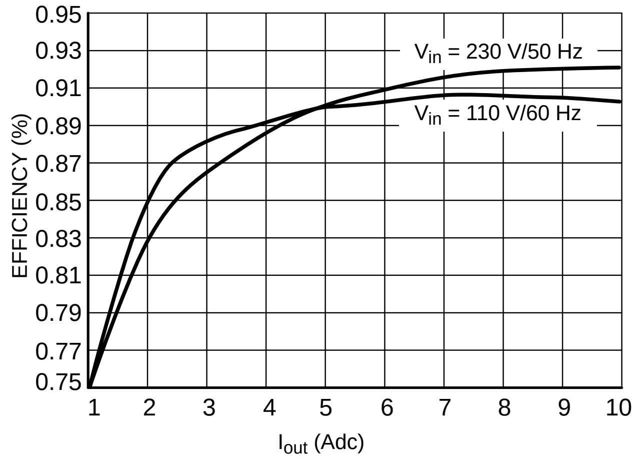

Figure 11. Efficiency of the Designed Converter versus Output Current

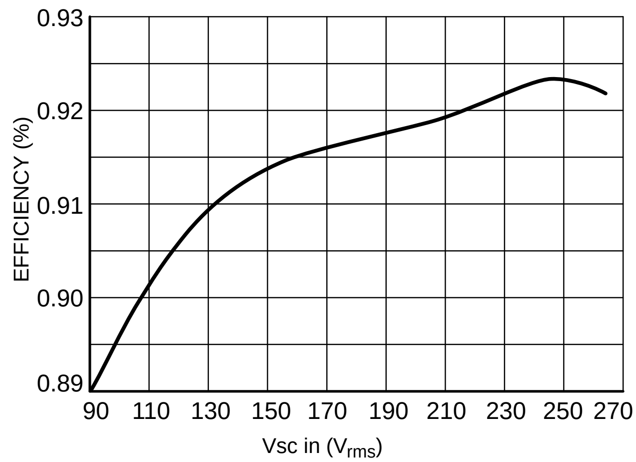

Figure 12. Efficiency of the Example Converter versus Input Voltage for Full Loaded Output

Loading characteristic of the prototype can be seen in Figure 13.

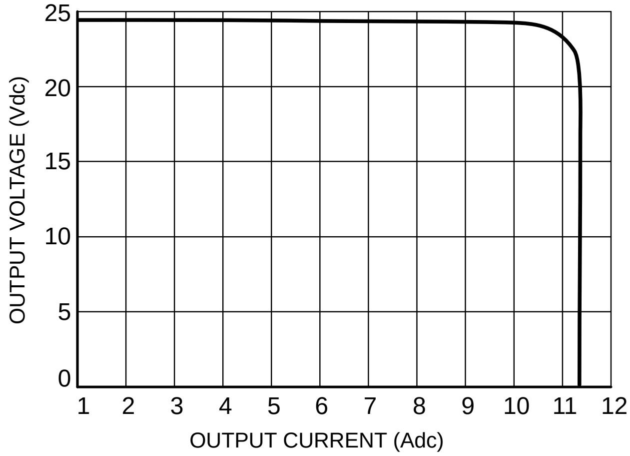

Figure 13. Loading Characteristic of the Proposed Converter

One can see that the CV operation is possible up to 10 A load current. Then the overcurrent circuit starts to limit output power and finally the hiccup mode takes place when output current goes over 11.4 A.

Several snapshots taken from the prototype can be seen in Figures 14 through 22.

Standby consumption is below 1.0 W for both input voltage levels, i.e. 230 VAC and 110 VAC.

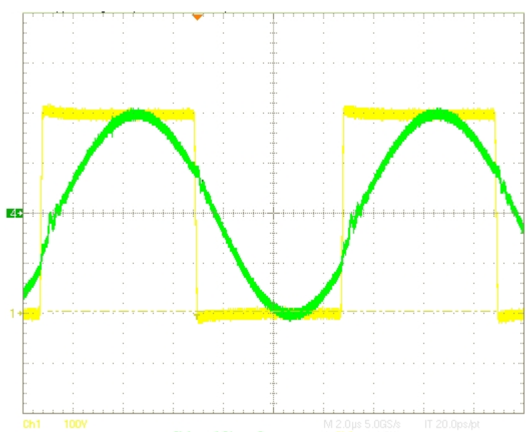

Figure 14. Primary Current and Waveforms for Full Loaded Converter

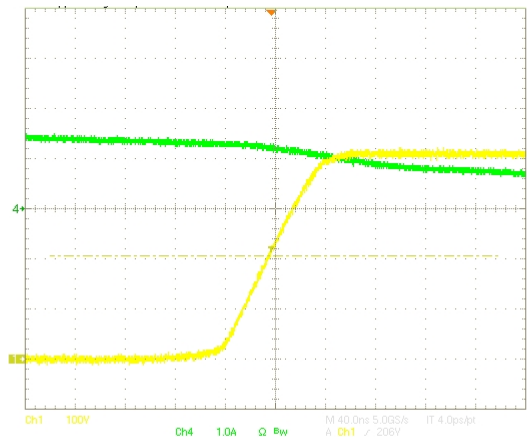

Figure 15. Detail of the ZVS Condition – Rising Edge

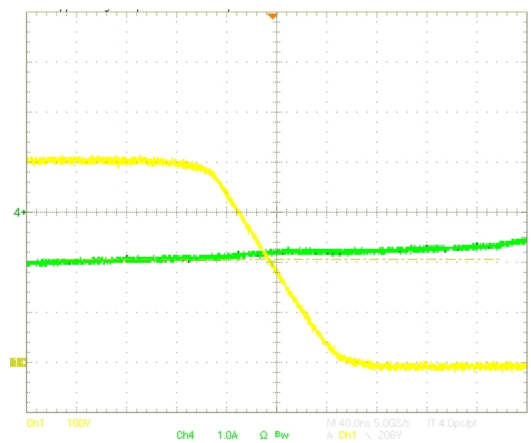

Figure 16. Detail of the ZVS Condition – Falling Edge

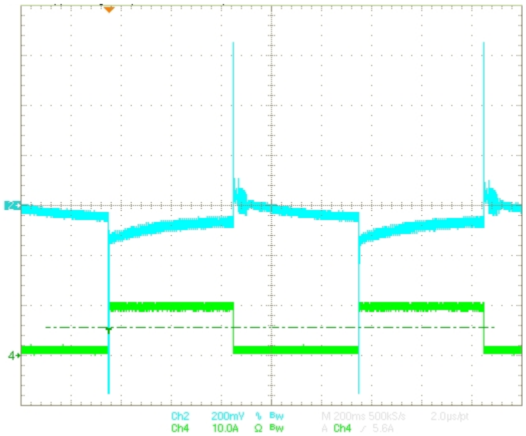

Figure 17. Load Regulation for 230 V Input Voltage

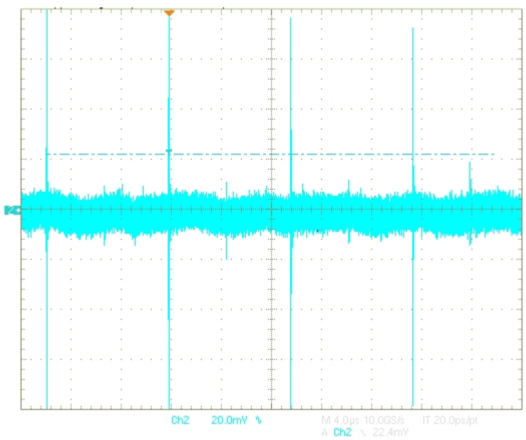

Figure 18. Output Ripple Under Full Load

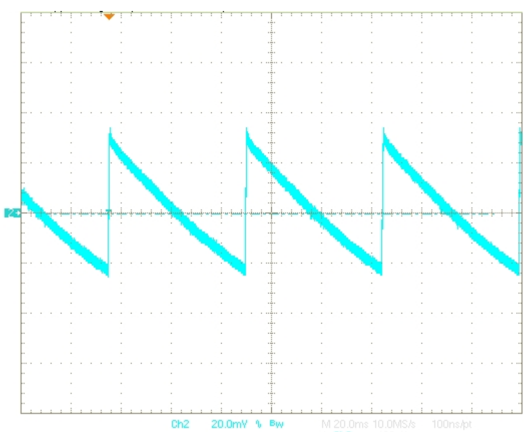

Figure 19. Output Ripple During the Skip Mode

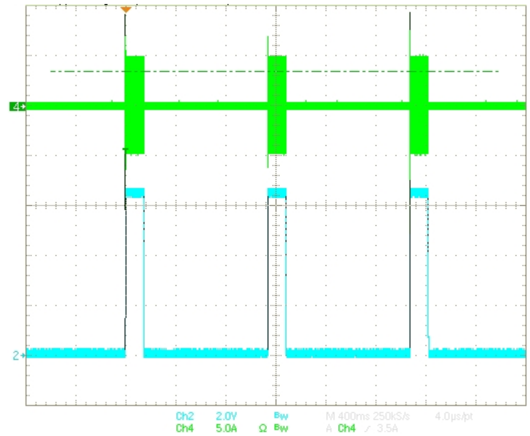

Figure 20. Operating Under Short Circuit

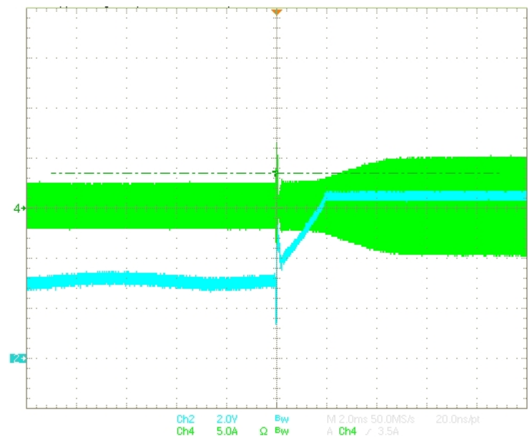

Figure 21. Full Load to Short Circuit Transition

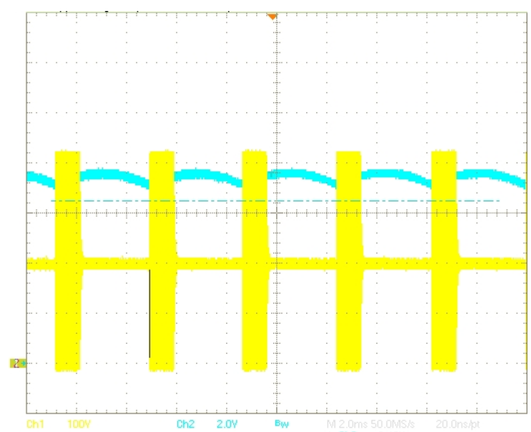

Figure 22. Operation in the Skip Mode

Layout Consideration

Leakage inductance on the primary side is not very critical for LLC converters compared to other topologies, because it will only slightly modify the resonant frequency. However it is well to keep the areas of each power loop as small as possible due to radiated EMI noise. A two−sided PCB with one side a ground plane helps (see Figures 23 through 26).

Special care has to be taken with Pins 1, 2 and 3 of the NCP1395 controller because these are high impedance pins. Ensure that these pins are not near high voltages and high dV/dt or use some ground shielding.

CAUTION

This board is intended only for demonstration and evaluation purpose. Board is designed for free air operation. Temperature damage of its components can occur if the board will be placed under the cover without forced air cooling.

Literature

1. NCP1395A/B data sheet.
2. Application note AND8184/D.
3. Application note AND8191/D.
4. Application note AND8255.
5. Bo Yang − Topology Investigation for Front End DC−DC Power Conversion for Distributed Power System.
6. Milan Jovanovich – Resonant converters training brochure.
7. M. B. Borage, S. R. Tiwari and S. Kotaiah − Design Optimization for an LCL – Type Series Resonant Converter.

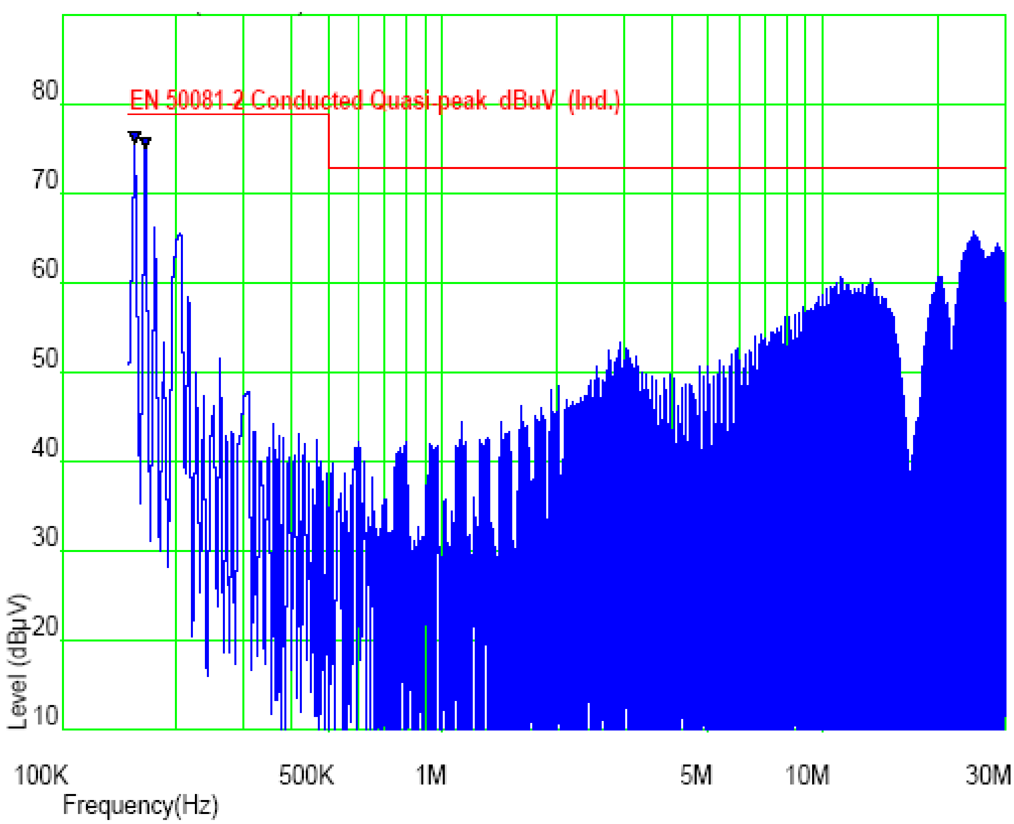

Figure 23. Conducted EMI Signature of the Board

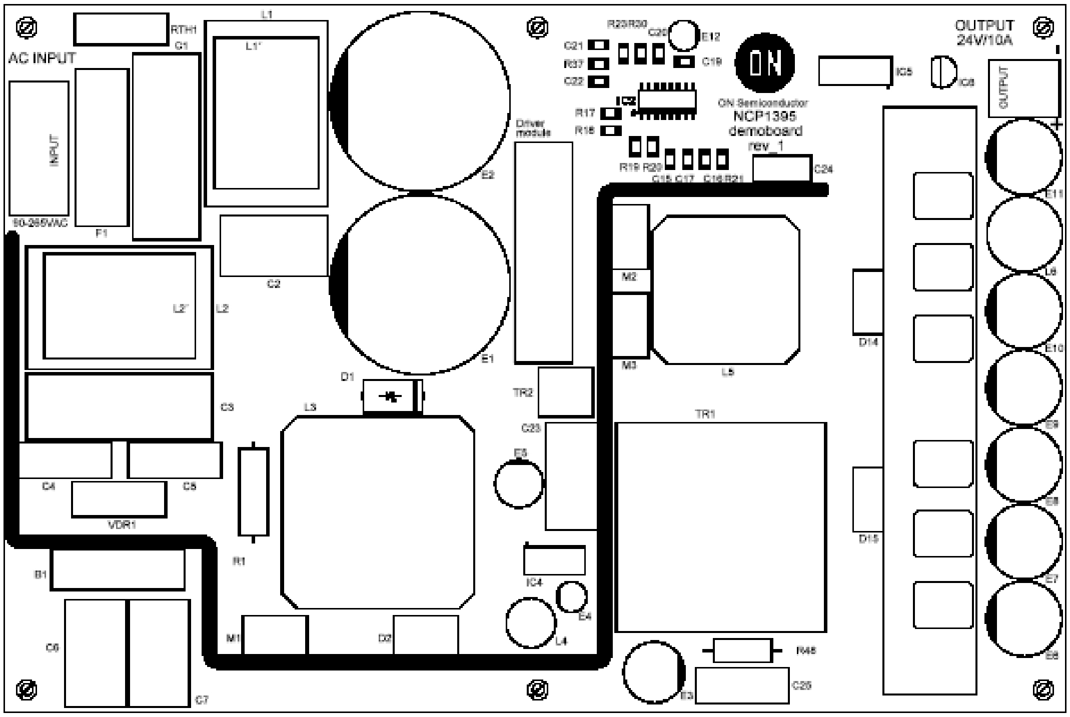

Figure 24. Component Placement on the Top Side (Top View)

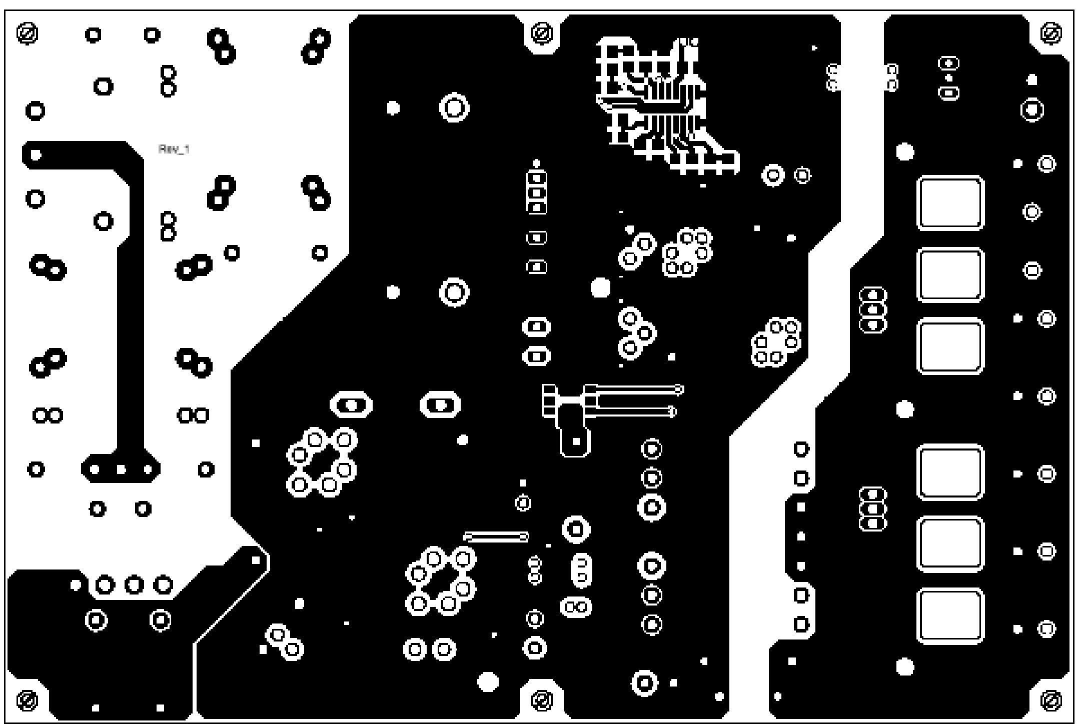

Figure 25. Top Side (Top View)

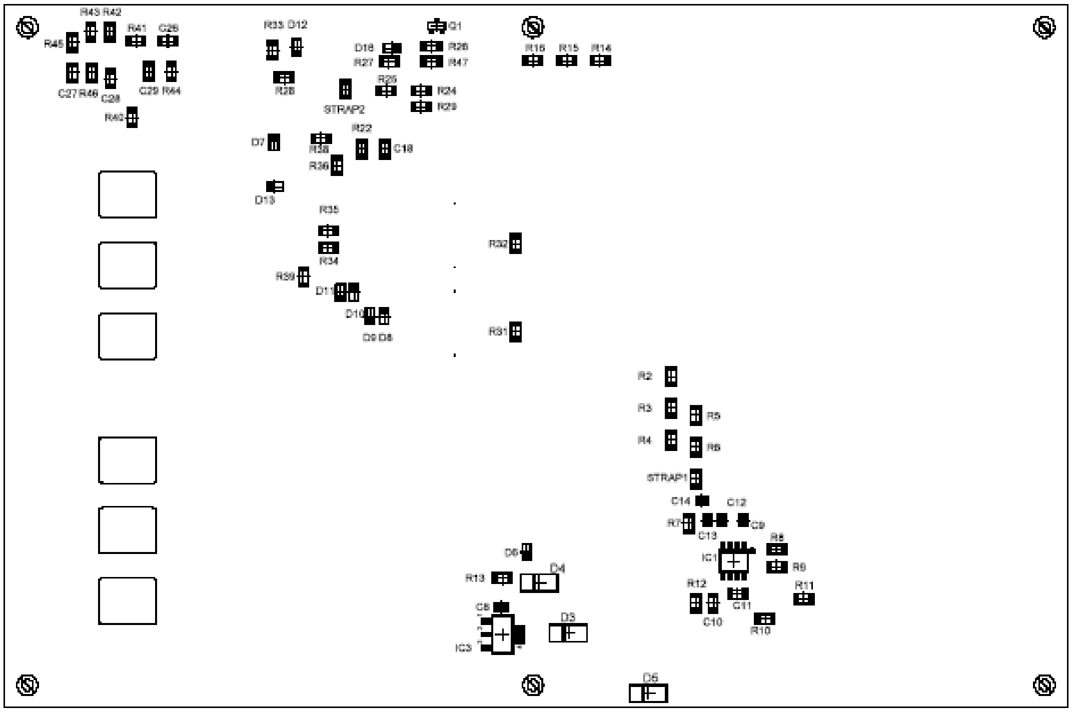

Figure 26. Component Placement on the Bottom Side (Bottom View)

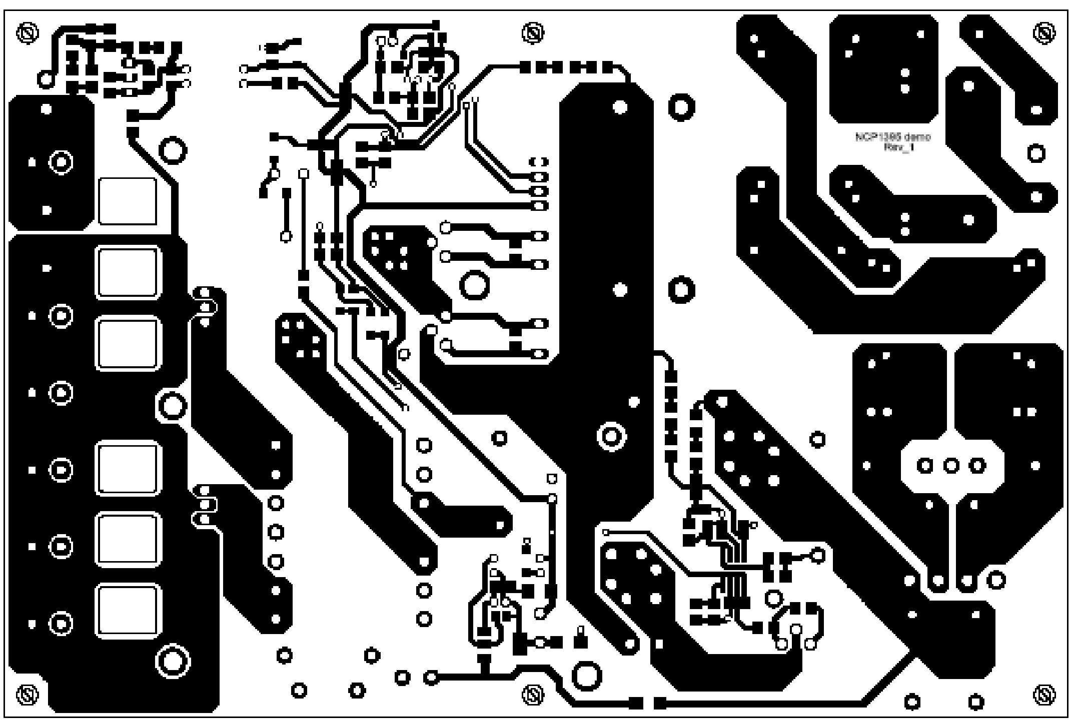

Figure 27. Bottom Side (Bottom View)

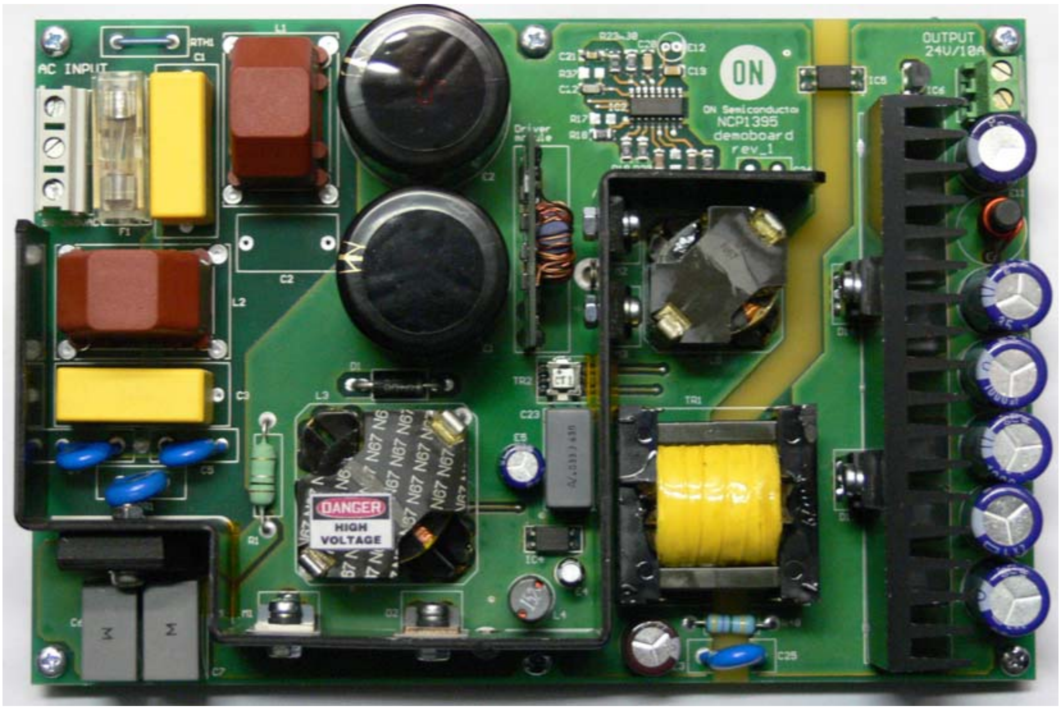

Figure 28. Photo of the Designed Prototype (Real Dimensions are 183 x 122 mm)

NCP1395 Demo Board Parts List

Designator	Description	Value	Manufacturer Part Number
B1	8.0 A Bridge Rectifier	KBU810	KBU810
C1,C3	EMI Suppression Capacitors (MKP)	470n	B81130B1474M
C2		NU	
C4, C5	Safety Ceramic Disc Capacitor	2n2	KZH 2200PF M 2E3 500 Y1/X1 B1
C6,C7	Polyester Chip Capacitor	330n	R46X−0,33UF 15 275V M M1 00
C8, C9, C13	SMD Capacitor	1nF	VJ0806Y102KXBAx
C10	SMD Capacitor	4n7	VJ1206Y472KXBAx
C12	SMD Capacitor	39n	VJ0806Y393KXBAx
C14	SMD Capacitor	100n	VJ1206Y104KXBAx
C15	SMD Capacitor	1u	VJ1206Y105KXBAx
C16	SMD Capacitor	4u7	VJ1206Y475KXBAx
C17		NU	
C11, C18, C19	SMD Capacitor	100n	VJ1206Y104KXBAx
C20, C29	SMD Capacitor	1n	VJ1206Y102KXBAx
C21	SMD Capacitor	10n	VJ1206Y103KXBAx
C22	SMD Capacitor	220p	VJ1206Y221KXBAx
C23	Polypropylene Capacitor	33n	R73−0,033UF 15 630V J 00 AA
C24		NU	
C25	Capacitor, Y1 Class	2n2/Y1	WKP222
C26	SMD Capacitor	22n	VJ1206Y223KXBAx
C27		NU	
C28		NU	
D1	3.0 A 1000 V Standard Recovery Rectifier	1N5408	1N5408
D2	600 V; 4.0 A; Zero Recovery Rectifier	CSD04060A	CSD04060A
D3, D4, D5	Surface Mount Ultrafast Power Rectifier	MURA160	MURA160
D6	Zener Diode 500 mW 15 V	MMSZ15T1	MMSZ15T1
D7		NU	
D8, D9, D10, D11	Small Signal Switch Diode 100 V	MMSD4148T1	MMSD4148T1
D12		NU	
D13		NU	
D14A,B; D15A,B	20 A 100 V Schottky Rectifier	MBRF20100CT	MBRF20100CT
D16	Zener Diode 500 mW 7.0, 5.0 V	MMSZ7V5T1	MMSZ7V5T1
E1,E2	Radial Lead Electrolytic Capacitor	150u/450V	K05 105°C 1500 μF/ 450V 0514 13592
E3	Radial Lead Electrolytic Capacitor	2u2/450V	CERA−2,2/450 10x12,5 KMG
E4	Radial Lead Electrolytic Capacitor	47u/25V	CERA−47/25 5x11 CD268
E5	Radial Lead Electrolytic Capacitor	220u/25V	CERA−220/25 8x12 LXZ
E6, E7, E8, E9, E10	Radial Lead Electrolytic Capacitor	1000u/35V	CERA−1000/35 12,5x25 LXZ
E11	Radial Lead Electrolytic Capacitor	220u/63V	CERA−220/63 10x16 A KMG
E12		NU	

NCP1395 Demo Board Parts List

Designator	Description	Value	Manufacturer Part Number
F1	Fuse	T3,15A	T3,15A
IC1	Compact Fixed−Frequency Current−Mode PFC Controller	NCP1653DR2	NCP1653DR2
IC2	Resonant Mode SMPS Controller	NCP1395A	NCP1395A
IC3	Self−Supplied Monolithic Switcher for Low Standby−Power Offline SMPS	NCP1012AP065	NCP1012AP065
IC4, IC5	Optocoupler	PC817P	PC817P
IC6	Adjustable Shunt Regulator 2.5−36 V/ 1.0−100 mA	TL431BILP	TL431BILP
L1, L2	Common Mode Inductor	2m7	PMEC103/V 2m7
L3	Inductor	650uH	IND−PFC−260W−v1
L4	Inductor	1m5	RFB0810−152L
L5	Inductor	100uH	IND−LLC−v1
L6	Inductor	2.2uH	IND−FLT−2u2−10A
M1	N−Channel 600 V 0.140 Ω−20 A TO−220	SPA20N60C3	SPA20N60C3
M2, M3	N−Channel 550 V @ Tjmax−0.30 Ω − 12 A TO−220FP	STP12NM50FP	STP12NM50FP
Q1	General Purpose Transistors NPN Silicon	BC846BLT1	BC846BLT1
R1	Axial Lead Resistor 3.0 W	0.1R	3W
R2, R7, R15, R16	SMD Resistor	470k	CRCW1206
R3, R4	SMD Resistor	750k	CRCW1206
R5, R6	SMD Resistor	2M2	CRCW1206
R8, R22, R42	SMD Resistor	3k3	CRCW1206
R9	SMD Resistor	11k	CRCW1206
R10	SMD Resistor	4R7	CRCW1206
R12	SMD Resistor	56k	CRCW1206
R13, R40, R44	SMD Resistor	1k	CRCW1206
R14	SMD Resistor	300k	CRCW1206
R17		NU	
R18	SMD Resistor	100k	CRCW1206
R19	SMD Resistor	680k	CRCW1206
R20	SMD Resistor	330k	CRCW1206
R21	SMD Resistor	150k	CRCW1206
R23, STRAP 1, 2	SMD Resistor	0R	CRCW1206
R24	SMD Resistor	1M2	CRCW1206
R25	SMD Resistor	220k	CRCW1206
R26	SMD Resistor	33k	CRCW1206
R27	SMD Resistor	5k6	CRCW1206
R28	SMD Resistor	100R	CRCW1206
R29		NU	
R30	SMD Resistor	56k	CRCW1206
R11, R31, R32	SMD Resistor	10k	CRCW1206

NCP1395 Demo Board Parts List

Designator	Description	Value	Manufacturer Part Number
R33	SMD Resistor	2k2	CRCW1206
R34		NU	CRCW1206
R35	SMD Resistor	180R	CRCW1206
R36	SMD Resistor	1k2	CRCW1206
R37		NU	
R38		NU	
R39		NU	
R41, R47	SMD Resistor	2k7	CRCW1206
R43	SMD Resistor	5k6	CRCW1206
R45		NU	
R46	SMD Resistor	18k	CRCW1206
R48	High Ohmic, High Voltage Resistor	4M7	0.5W VR37 Series
RTH1		NU	
VDR1	Voltage Dependent Resistor	CV 275 K 10 B 1	CV 275 K 10 B 1
TR1	Transformer	TRAFO ETD29	TR−LLC−v3−24V
TR2	CS Transformer	CST1−100LB − CS TR.	CST1−100LB − CS Tr.
H1	PCB Connector	NA	SVPS CEE7.5/3 Grey
H2	PCB Connector	NA	SVPS MV 252/5.08 Green
HS1	Heat Sink	NA	CHL01−BLK
HS2	Heat Sink	NA	SK 505 30 SA
Driver Module with NCP5181			
R1	SMD Resistor	18R	CRCW1206
R2, R3	SMD Resistor	10R	CRCW1206
R4, R5	SMD Resistor	0R	CRCW1206
C1, C2	SMD Capacitor	100n	VJ1206Y104KXBAx
D1	Surface Mount Ultrafast Power Rectifier	MURA160	MURA160
IC1	Integrated Half Bridge Driver	NCP5181	NCP5181
Driver Module with Transformer			
R1, R2	SMD Resistor	150R	CRCW1206
R3, R4	SMD Resistor	10k	CRCW1206
R5, R6	SMD Resistor	22R	CRCW1206
R7, R8	SMD Resistor	8R2	CRCW1206
STRAP	SMD Resistor	0R	CRCW1206
C1, C2	SMD Capacitor	100n	VJ1206Y104KXBAx
D1, D2, D3, D4	0.5 A 40 V Schottky Rectifier	MBR0540T1	MBR0540T1
D5, D6, D9, D10	Small Signal Switch Diode 100 V	MMSD4148T1	MMSD4148T1
D7, D8, D11, D12	Zener Diode 500 mW 18 V	MMSZ18T1	MMSZ18T1

NCP1395 Demo Board Parts List

Driver Module with Transformer			
Q1, Q3	General Purpose Transistors NPN Silicon	BC817−40LT1	BC817−40LT1
Q2, Q4, Q5, Q6	General Purpose Transistors PNP Silicon	BC807−40LT1	BC807−40LT1
TR1	Transformer	NA	TR−DRW−01

Please see the NCP1395A/B product folder on www.onsemi.com for PCB Gerber files and other collateral information regarding this demo board.

PUBLICATION ORDERING INFORMATION

LITERATURE FULFILLMENT:
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