Master's Thesis

Implementation and verification of a dimmable power LED driver

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Abstract

The notion of this thesis is rooted in the concept of a smart street light system using LEDs. The purpose was to use a reference design from Texas Instruments, containing a power supply and a dimmable current controlled LED driver with an output of 350 mA. The reference design driver was then redesigned with adjustable output currents of 350, 700 and 1050 mA. The reference design has been examined to get a deep knowledge about the circuit and what changes that had to be made and were. By calculate new values for resistors controlling the three currents and circuit protection the criteria could be achieved. The new values were verified and optimized with simulations in TINA–TI. To use the same circuit for different currents, the design solution was to use jumpers to switch between resistors to get the desired current. An LED lamp at 80 W was constructed for circuit testing and mounted on a heat sink to withstand the high temperature. The characteristics for the circuit such as efficiency for variety input voltage to the driver and different load size were investigated. A layout for a printed circuit board was made in EAGLE and outsourced for manufacturing and the components was mounted for testing.

The task was succeeded in the simulations for the different currents and the dimming with the new calculated values for components but due hardware problem could the PCB not be fully tested.

Acknowledgments

First of all we would like to thank our advisors Bertil Larsson at EIT and Serdar Köse at Greinon Engineering AB for all support during this thesis work. We would also like to give a special thanks to Bertil Lindvall and Martin Nilsson for wise comments and helping hands when needed. Thanks to our corridor neighbors Christoffer Cederberg and André Ericsson for good inputs and advices. Big greetings to Sara Nilsson, thank you for the component donation. Thanks to Johan Persson and Anders Eriksson for keeping us company during the deserted months in the summer. Last but not least, we would like to express our gratitude to our roommates David Etienne and Julian Rath for the time spent together at 4111.

Thea & Oscar

Table of Contents

1 Introduction

This thesis work has been done in collaboration with Greinon Engineering AB as a part of their project "*Intelligent Street Lighting*".

1.1 Greinon Engineering AB

Greinon was founded in 2012 and their profile is to "*develop intelligent engineering solutions for resources optimization and cleaner environment*". They have expertise in wireless communication, network design and implementation, software and hardware development and resources optimization.

One of the developments by Greinon is an intelligent street light system. The idea is to save energy by dimming the street lights depending on the need of light. By using sensors, the street lights can be able to sense the presence of humans/bicycles/vehicles and light up when needed. Each pole can communicate and send information to each other and to a centralized server to be monitored in real time. Another advantage is that this system can detect broken lights, which a normal street light system can not. In today's community when a street light breaks, the system relies on the residents in the city to report where there is a broken street light to the city council. The new system is estimated to save up to 80 % in energy consumption. [1]

1.2 Purpose

The purpose of this thesis is to investigate the opportunity of making a current controlled LED driver consisting of one output that can be switched between the output currents of 350, 700 and 1050 mA. A power supply with three separated drivers [2], each of 350 mA, will be used as a reference design. The driver will be modified to be able to handle the three different output currents from the same driver. Before any construction is made, the new design will be implemented and verified in simulation software.

Requirements:

1. The LED driver should be able to handle a 20 – 80 V LED source.
2. The LED driver should be able to provide outputs of 350, 700 and 1050 mA.
3. The LED driver should be able to dim the LED source with PWM.
4. The PCB schematics and layout shall be prepared in EAGLE CAD.

1.3 Restrictions

Simulations will only be performed on the driver circuit and not the power supply. The reason for this is that no changes are supposed to be made on the power supply. By this reason there is no point of spending time building and simulating it. The output voltage from the power supply, V_{DC} and V_{AUX}, will be replaced with voltage generators to supply the driver with the correct voltage.

1.4 Investigation areas of the different currents

The following areas will be investigated during this thesis to get a deeper understanding of the circuit.

- **Appropriate PWM interval for the driver**
  The appropriate PWM frequency for dimming between 10 – 90 % for 350, 700 and 1050 mA will be calculated and verified through simulations.

- **Variety in the load**
  The driver will be tested with 0, 25, 50, 75 and 100 % load to see how this affects the output current.

- **Variety in the supply voltage**
  The impact of variations of the supply voltage to the driver will be measured.

- **Study the efficiency**
  The efficiency of the driver will be studied during different circumstances and compared to given test results from TI.

- **Study the ripple**
  The ripple of the output current will studied to make sure that it is kept at an acceptable level for each output current. If the ripple gets too high, suggestions of improvements will be given.

1.5 Software

Two software have been used in this thesis work, TINA–TI and EAGLE.

1.5.1 TINA–TI

TINA–TI is a software used for circuit design and simulations and is based on a SPICE engine. The original program TINA was made by DesignSoft inc. and TINA–TI is a development of this program specially made for Texas Instruments (TI) customers as a result of a collaboration between these two companies. TINA–TI is a freeware and can be downloaded from TI's homepage in a few different languages.

The main reason TINA–TI was chosen for this thesis work was due to the fact that some Integrated Circuits (IC) needed from TI, such as the TPS40210, already had models for this software and it was hard to get proper simulation models for the components somewhere else. TINA–TI does not have all the functions and is a bit limited compared to more powerful software from DesignSoft. Though it has no limitations when it comes to number of nodes and parts in the circuits, it was more than enough to accomplish the simulations in this thesis. [3]

1.5.2 EAGLE CAD

EAGLE CAD is a software, developed by Cadsoft Computers, made for schematic and Printed Circuit Boards (PCB) layout and is available for Windows, Macintosh and Linux systems. There is also a free version of EAGLE.

The software has an extensive library of components, including collections from many manufacturers. Manufacturers often do their own EAGLE library and make them available online. Back annotation is possible from the PCB design to the schematics and traces drawn in the schematics will automatically be shown as thin connections between its components in the PCB design for easy trace layout. Auto routing is also available but not used in this thesis. [4]

1.6 Equipment

Equipment used during bread board and PCB testing.

Power Supplies:

- Instek GPC-3060, 3A 5V DC fixed, analog DC Power Supply, 0 – 30 V Dual tracking.
- Powerbox 3000, 3A 5V DC fixed, 0 – 20 V 3A, 0 – 40 V 1.25 A.

Multimeters:

- Fluke 75
- Fluke 10
- UNI-T UT33D

Oscilloscope:

- MSO–X 3014A, 100 MHz, 2 GSa/s

Transformer:

- Berco – Rotary Regavolt 42C, Variable Transformer, Max Input 230 V, Output 0 – 270V

Isolation transformer:

- AB Erik Sundberg, Gb 185 XB, 1 phase, 50 Hz, 2,5 A

1.7 Notifications

- For the IC the pins are named differently at variant sources. In this thesis the names are used from the reference design. Information from other sources may have different pin names but the pin number is always the same.

- In the PCB's reference design, 7 components, C_{100}, C_{102}, D_{100}, D_{101}, D_{102}, R_{100} and R_{101}, are missing from the schematic. A theory could be that they are adjustable for required voltage V_{DD} and V_{DC} and thereby been omitted. The missing components have been placed to fit in the PCB in best suitable way.

- The reference circuit has only been tested by Texas Instruments through simulations and never physically.

2 Theory

By replacing the streetlight with LED lamps many benefits can be achieved like longer lifetime and the advantage of dimming the light and still maintain the same light quality. The diodes in an LED are current sensitivity and to prevent variety in the current, the LED driver can advantageously control the current and there by prevent an broken LED.

2.1 Street light background

The streetlight used today in Sweden and around the world, a majority uses High Pressure Sodium (HPS) lamps. By exchanging these to LED lamps there are several benefits to achieve.

The estimated lifetime of a HPS lamp is 12 – 25 000 hours compared to an LED lamp that has an estimated lifetime of 50 000 hours and sometimes as high as 100 000. The high lifetime contributes to less material and maintenance cost. LED lamps are easy to dim using Pulse Width Modulation (PWM) and with this technique the street light can be dimmed between 1 – 100 % without affecting the quality of the light. LEDs are also not affected by being turned on and off multiple times like an ordinary light bulb. Using PWM dimming may even extend the estimated lifetime due to lower working temperature in the LEDs.

Another advantage is the light that LED lamps emit. The light is perceived much more natural than the light emitted from HPS. This can be expressed in something called Color Rendering Index (CRI) where the maximum value is 100 and correspond to the color in broad daylight. HPS has a value around 25 while an LED 85 – 90. LED also has a higher color temperature, 3 200 – 6 400 K compared to HPS 2000 – 3500 K. These two parameters will result in a clear white light from the LED instead of the yellow light from HPS that is used today, Fig. 1. [5]

Fig. 1. The light emitted from HPS (left) compared to LED (right). [6]

2.2 Control an LED driver

There are some smaller light applications, as a simple flashlight, that can be run without a LED driver and instead use internal resistor in the battery to regulate the current. That is a cheap solution but the disadvantage is that the resistor produces heat when it decreases the current and as the battery discharges the current and the light from the diode will become weaker. The semiconductor junction in an LED is quite sensitive and requires a specific current and does not operate properly if the current is too high or too low. If the current through the LED is too low it will result in a week light performance and for a too high current the consequences could be an overheated and broken LED. An LED driver controls either the voltage, if the LED lamp's properties are well known, or the current. The current controller is more reliable due the fact the diode's current sensitivity and the risk for too high or low current is eliminated.

2.3 Dimming function

The dimming function for an LED lamp can be controlled with both Pulse Width Modulation (PWM) and by controlling the current. Most of the LEDs on the market have a wavelength and a threshold voltage that varies with different currents and will not only change the strength of the light but also the color. Therefore it is preferred to control dimming with PWM since it has a minimal change of the light's color.

The frequency of the PWM has to be high enough to avoid visible flicker, a minimum of 200 Hz is recommended [23], but it is also important to take the switching frequency of the driver into consideration. If the switching frequency is lower than the PWM frequency, or nearly equal, the dimming function will not work properly. It is desirable that the circuit switches multiple times in each pulse to get a steady output. Another aspect is the rise time of the output current. To get an effective and accurate dimming, the relation between the rise time and duty cycle has to be significant. The longer the output stays at a steady state, compared to the rise time, the better the dimming will become. These requirements set the range in which the PWM frequency can be chosen.

3 Reference design

The reference circuit works between an input range of 100 – 277 VAC and consists of two parts, the power supply and the driver part.

3.1 Power supply

The power supply, Fig. 2, will down transform the DC voltage on the secondary side of the transformer, T_1, and with help from voltage regulators the voltage will be kept on a stable level. It also includes overvoltage protection, leading edge blanking circuit, isolated feedback and the controlling IC, UCC28810.

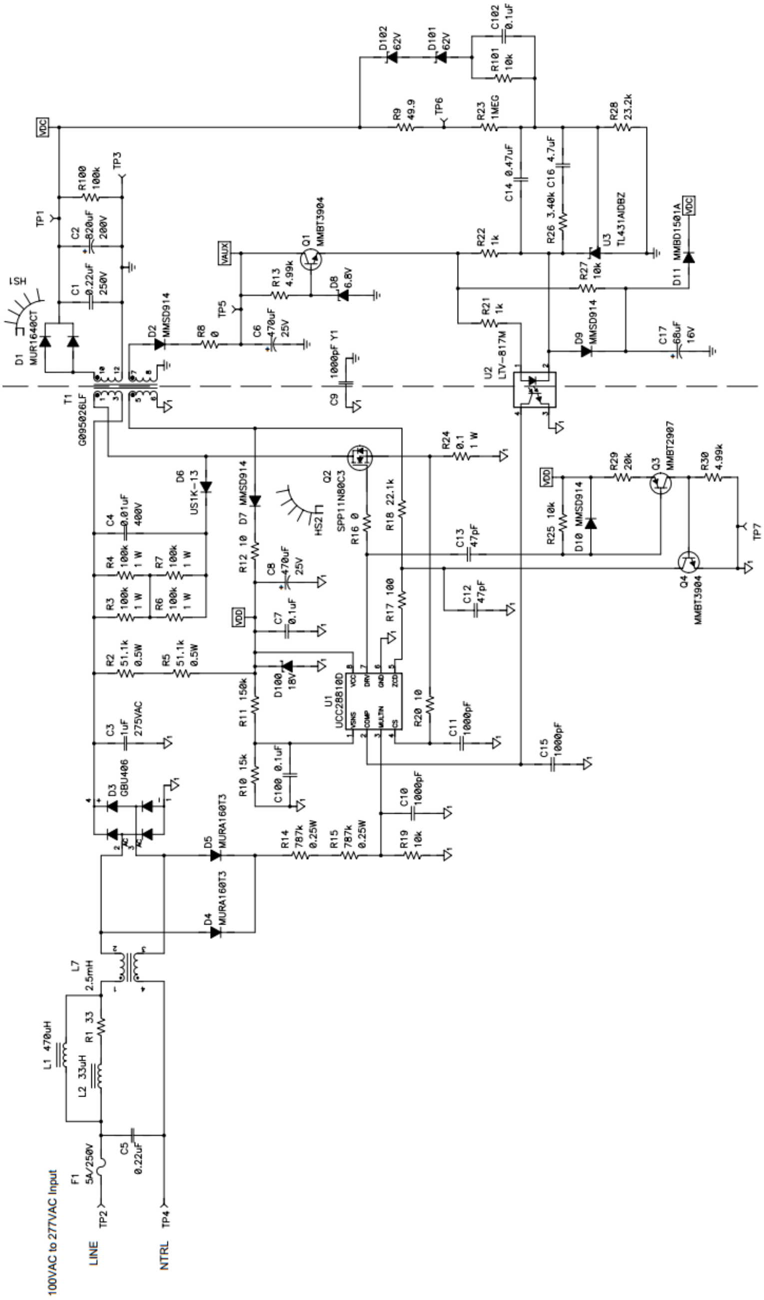


Fig. 2. The reference circuit of the power supply. [7]

3.1.1 Filter and rectifiers

L_1 and R_1 are there to damp oscillations caused by the flyback converter's on and off switching, Fig. 3. [8]

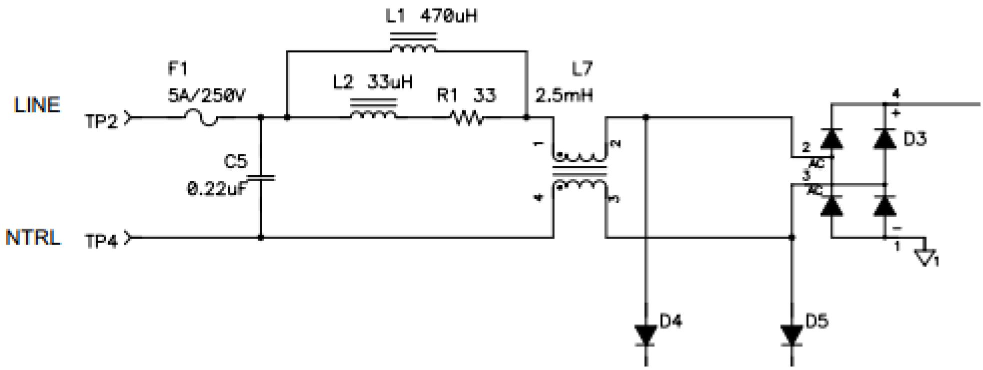


Fig. 3. Part of the schematic of the filter and rectifiers. [7]

3.1.2 Snubber

When the MOSFET, Q_2, is switching off, leak inductance may appear as voltage spikes. To avoid that these high voltage spikes destroys the transistor a snubber is added as an over voltage protection and it takes care of the leak inductance. The resistors take care of some leak inductance by transforming the current into heat. The capacitor charges up by the left overs and discharges the next time the MOSFET switches on, Fig. 4.

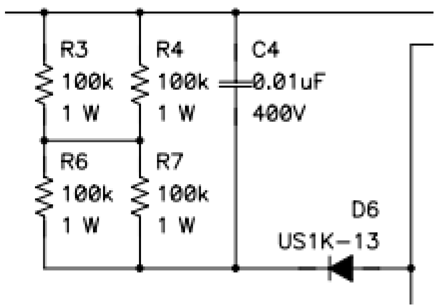


Fig. 4. Part of the schematic of the snubber. [7]

3.1.3 Optoisolator

Since the flyback transformer isolates the primary side from the secondary side an optoisolator is needed to give an isolated current feedback to the IC, Fig. 5.

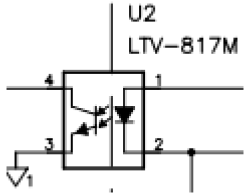


Fig. 5. Part of the schematic of the optoisolator. [7]

3.1.4 Voltage regulators for V_{DC}

U_3 is a part of the feedback loop which regulates the output voltage, Fig. 6. This is a three terminal adjustable shunt regulator that works as a zener diode but is regulated by a reference voltage of 2.5 V. If the voltage over the reference pin exceeds above 2.5 V the shunt regulator will be reverse biased and start to conduct. When this happens currents starts to flow through the diode inside the optoisolator, U_2, which lowers the voltage on the primary side. This will be sensed by the COMP pin on UCC28810 and stops the switching until V_{DC} has fallen to an accepted level. [9]

The two zener diodes, D_{101} and D_{102}, is an overvoltage protection active above 124 V, Fig. 6.

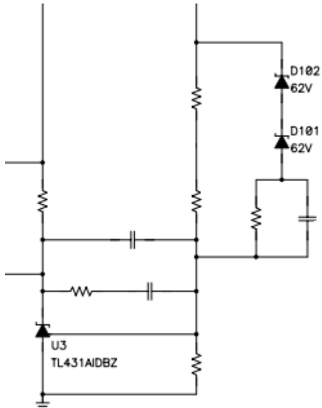


Fig. 6. Part of the schematic with U_3, D_{101} and D_{102}. [7]

3.1.5 The IC in power supply, UCC28810

The 8 pin UCC28810 [8], Fig. 7, is a lightning power control and can control a flyback, buck or boost converter operating in a critical conduction mode. The device includes features such as current sense comparator, overvoltage and open feedback protection. Block diagram for the IC can be found in Appendix C.1.

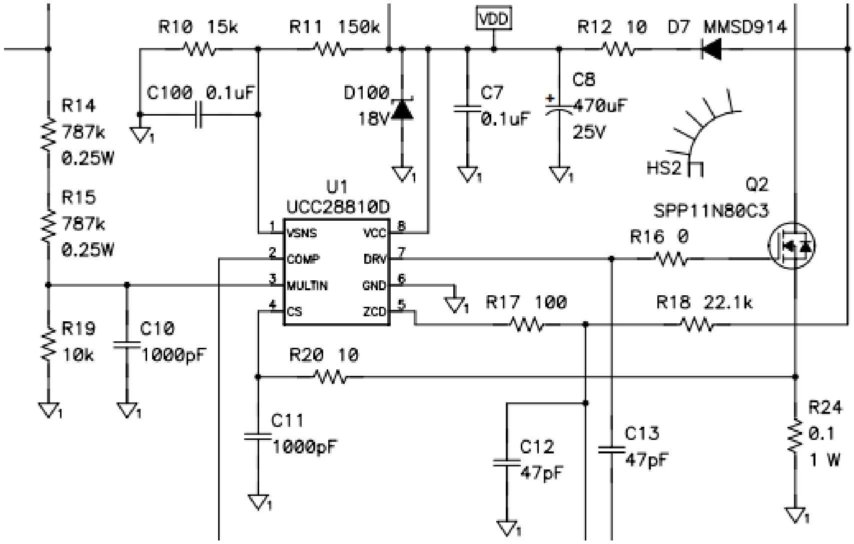


Fig. 7. The IC UCC28810 and its pin connections. [7]

Voltage Sense (VSNS), 1, input

Works as an enable/disable input. There is an internal threshold voltage of 0.67 V which has to be exceeded for the IC to start. This pin also works as an overvoltage protection. When the input voltage goes above 7.5 % of its nominal value of 2.5 V, overvoltage protection is activated and stops the gate drive from switching.

Comparator (COMP), 2, output

This pin is connected to the optoisolator and when the zero energy or overvoltage protection is detected the voltage falls below 2.5 V. When the voltages drops even more, to less than 2.3 V the zero energy detect comparator is activated and prevents the gate drive from switching.

MULTIN, 3, input
An input for the current regulator. It senses the instantaneous supply voltage, V_{DD}, through an external voltage divider. Recommended operation range is between 0 – 3.8 V.

Current Sense (CS), 4, input
Senses the current through the transistor Q_2 and turns the IC off if the current is too high.

Zero Current Detection (ZCD), 5, input
The bias winding from the transformer is used to give feedback to the ZCD pin. When the input voltage exceeds the threshold voltage, 1.7 V, will the gate drive be ready and trig when the voltage decreased below 1.4 V and the bias winding current is 0 A. If the trig signal is absent for 400 µs will a restart timer set gate driver high.

Ground (GDN), 6
Reference ground for the device.

Gate Drive (DRV), 7, output
The gate driver pin controls the MOSFET, Q_2, with PWM and thereby controls the switching of the flyback transformer.

VCC, 8, input
Supply voltage for the IC. It is directly connected to V_{DD} and has a threshold voltage of 15.8 V where the IC is turned on, and is turned off when the voltage falls below 9.7 V.

3.1.6 Leading edge blanking circuit on the ZCD pin

During startup, the input voltage at ZCD pin will be lower than the trigging point and leakage inductance from the transformer's windings can cause voltage spikes and trig DRV before the bias winding current is 0 A. If this behavior occurs repeatedly the current in the primary side will increase gradually and result in a saturated transformer and an overloaded MOSFET that breaks. The leading edge blanking circuit, Fig. 8, works as a delay timer for the trigging and starts when the voltage at DRV turns low so voltage spikes occurring from switching will be blanked.

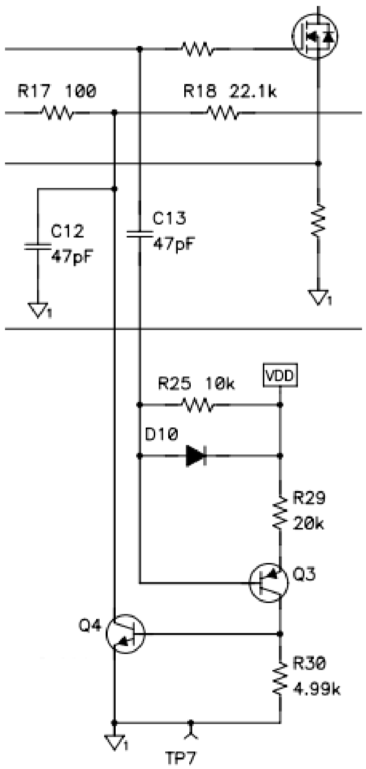


Fig. 8. Leading edge blanking circuit on ZCD pin. [7]

3.1.7 Transformer

The transformer, T_1, in the power supply is a four winding flyback transformer. The structure of the four windings is shown in Fig. 9 and the turn ratios between them are as follow:
1 – 3:5 –6:7 – 8:10 – 12 = 1:0.083:0.055:0.5

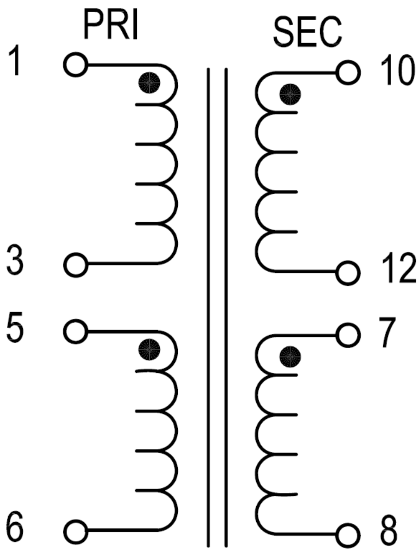


Fig. 9. The structure of the four winding transformer. [Appendix B.1]

This transformer was obsolescence and had to be replaced. A datasheet [Appendix B.1] was received from the manufacturer GCI Technologies. No equivalent replacement where to be found and after a long search it was decided that a new transformer should be ordered directly from a manufacturer. A request for quotations was posted online and a lot of manufacturers responded. The task of making the transformer was given to MagTop Technology Co. Ltd in China. They manufactured it and sent a sample of five transformers [Appendix B.2] which has been used in the circuit during the testing.

3.1.8 Output voltage from power supply

To calculate the output voltages of the power supply, V_{DC} and V_{AUX}, the knowledge of the reference voltage of U_3 was used [9]. This voltage is the same voltage as over R_{28} and with that knowledge V_{DC} could be calculated by voltage division (3.1.8.a), Fig. 10.

$$V_{DC} = \frac{R_9+R_{23}+R_{28}}{R_{28}} \cdot V_{R28} \tag{3.1.8.a}$$

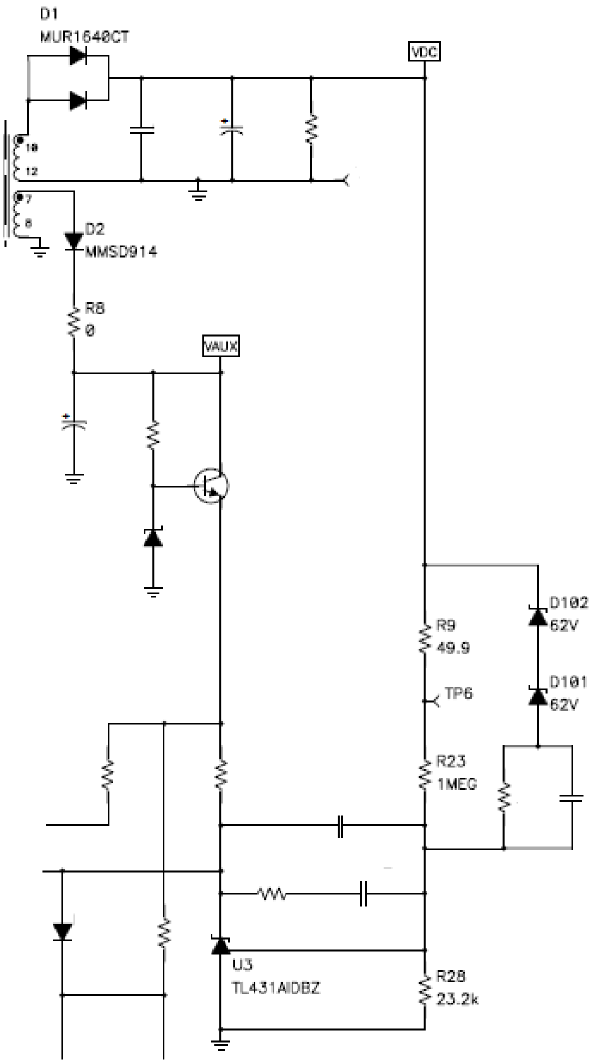


Fig. 10. Schematic over essential parts for calculations of V_{DC} and V_{AUX}. [7]

By using the knowledge of the transformer ratio between the windings at the secondary side, Fig. 9 page 16, the output voltage V_{AUX} could be calculated.

The voltage, V_{sec1}, for windings at sec_1 was achieved by simply adding the voltage drop from diode D_1 to V_{DC} (3.1.8.b).

$$V_{sec1} = V_{DC} + V_{D1} \tag{3.1.8.b}$$

By calculating the turn winding ratio between the two windings, sec_1 and sec_2

$$Ratio_{sec1/sec2} = \frac{Ratio_{sec1}}{Ratio_{sec2}} \tag{3.1.8.c}$$

the voltage V_{sec2} is then given by dividing V_{sec1} with the ratio between them.

$$V_{sec2} = \frac{V_{sec1}}{Ratio_{sec1/sec2}} \tag{3.1.8.d}$$

The final step was to withdraw the voltage drop over D_2 and V_{AUX} was set (3.1.8.e).

$$V_{aux} = V_{sec2} - V_{D2} \tag{3.1.8.e}$$

3.2 Driver

The driver can be divided into three essential parts, the IC TPS40210, a current detector and a set of external controls. An overview of reference driver can be seen in Fig. 11.

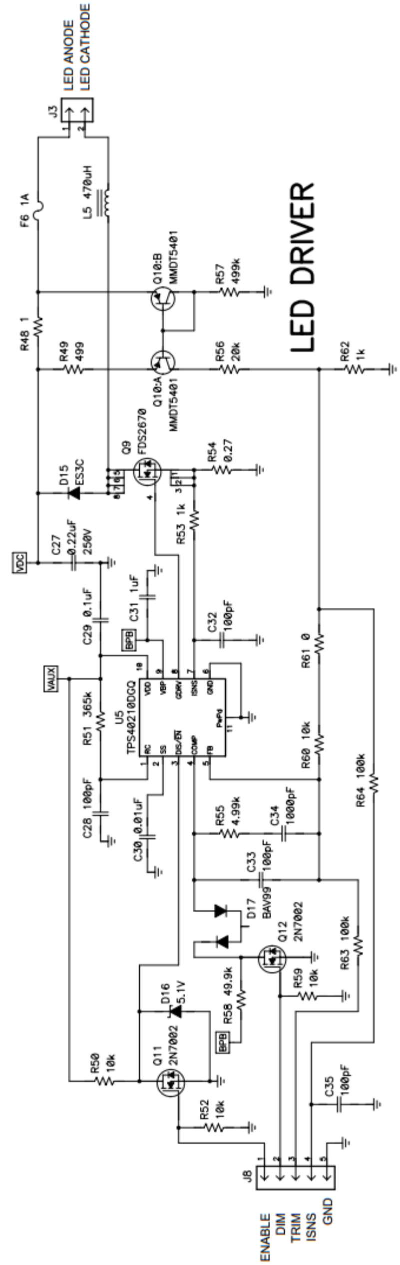


Fig. 11. The reference circuit of the driver with external controls. [7]

3.2.1 The IC in driver, TPS40210

The 10th pin of TPS40210 [10], Fig. 12, is a non–synchronous boost controller and can handle an input voltage between 4.5 – 52 V to work properly. The device includes features such as overcurrent protection, adjustable soft start and oscillator frequency. It is current mode controlled which both provides an improved simplified loop compensation and transient response. Block diagram for the IC can be found in Appendix C.2.

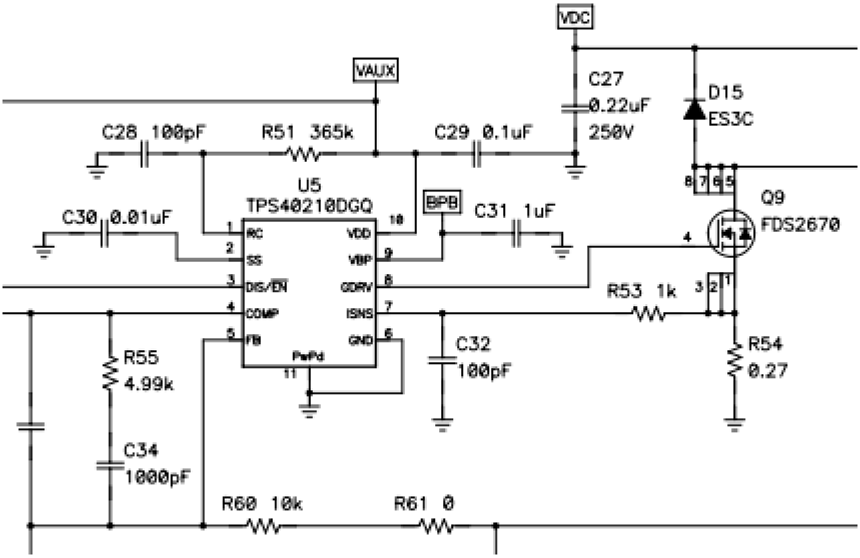


Fig. 12. The IC TPS40210 and its pin connections. [7]

Switching frequency (RC), 1, input
At this pin, R_{51} and C_{28} can be set to get the desired switching frequency.

Soft Start (SS), 2, input
A capacitor, C_{30}, of 0.01 µF is connected between the pin and ground. This works as a soft starter and a lower value of the capacitor will result in a faster startup of the TPS40210 and conversely. It is important to have enough soft start time to avoid overcurrent at startup. The soft starter is used when the circuit is switched on but can also be used as a restart timer. When an error occurs, like an overcurrent, the TPS40210 will turn of and the capacitor connected to SS will discharge. When discharged, the circuit will turn back on and try to start up again and the capacitor will charge. If the overcurrent protection is still enabled the capacitor will be discharged again and the process will be repeated until the protection is disabled.

Disable/Enable (DIS/EN), 3, input
Controls if the driver should be on or off. By putting the input voltage on the pin low, below 1.2 V, all the blocks in the IC will be enabled and by putting it high, above 1.2 V, they will be disabled.

Compensation (COMP), 4, output
Is an error amplifier and regulate the control loop compensation network between this pin and the FB pin.

Feedback (FB) pin, 5, input
This pin is used to control the switching output of the gate drive pin. It has a reference voltage at 700 mV. When the voltage at the pin exceeds the reference voltage, the output signal from the gate drive is shut down until the voltage has dropped below 700 mV. This feature is used to control the output current of the circuit.

Ground (GDN), 6
Connects the IC to ground.

Current Sense (ISNS), 7, input
Uses the input voltage to provide a current feedback in the control loop and to trig the overcurrent protection. The reference voltage is set to 150 mV and when exceeded, the entire circuit is shut down and a restart timer is started in forms of a discharge of the capacitor connected to the SS pin.

Gate Driver (GDRV), 8, output
The gate driver pin controls the MOSFET, Q_9, with PWM and uses the same voltage as the BP pin, nominal 8 V. The MOSFET will not switch for voltage below 5 V.

By Pass (BP) pin, 9, output
The IC has a linear regulator by pass that supplies the internal circuits with power. The PB voltage is nominally 8 V but can be lower depending on the input voltage, V_{DD}, for the circuit. The higher BP, the faster the startup time for the circuit will be. As long as the input voltage, V_{DD}, is over 8 V the BP output will be fixed at 8 V.

VDD, 10, input
The supply voltage for the system.

3.2.2 Current detector

The current detector, Fig 13, works as a current regulator, by using I_{ref} and I_L and scale it as (3.2.2.a) with the ratio between R_{49} and R_{48}, the regulated current, I_R , can be set. The regulated current is adjusted by a load resistance, R_{62} 4.2 The load resistance, page 28, and keeps the current through the LED lamp fixed. The reference transistor, $Q_{10:B}$, sets a common base voltage.

$$I_R = \frac{R_{48}}{R_{49}} \cdot \left(I_{ref} + I_L\right) \qquad (3.2.2.a)$$

To make sure the PN junctions behave identically, it is important that they have the same temperature during operation time. In this circuit, the transistors are manufacture together on an IC but if two separate components are used it can be fixed by gluing the transistors together. Due to the fact that the current through the transistors are different, they will not appear identically and a slightly difference in V_{BE}will occur.

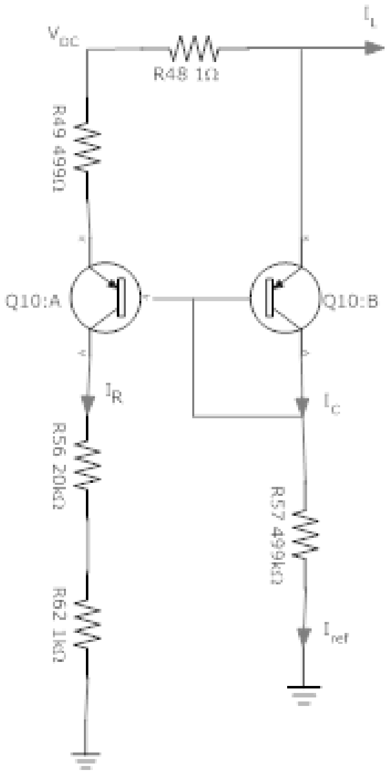


Fig. 13. The current detector with load resistance R_{62}.

The three reference currents in the detector, I_{ref}, can be expressed by (3.2.2.b) and is almost identical to I_C.

$$I_{ref} = \frac{V_{DC} - R_{48} \cdot I_L - V_{BE2}}{R_{57}} \qquad (3.2.2.b)$$

I_{ref} in the current detector showed out to be very small and was by that reason negligible in further calculations.

3.2.3 External controls

The main external controls that have been used during the thesis work are the enable and the dimming functions.

ENABLE

The ENABLE pin is used for turning the driver on and off by changing the state of the MOSFET, Q_{11}, Fig. 11 page 19. When ENABLE is high, around 2.5 V, the MOSFET starts to conduct and current goes through it to ground. The voltage at DIS/EN becomes low, which will enable the TPS40210. If instead ENABLE is kept low, the voltage at DIS/EN will stay high and the IC is turned off. The voltage level for on/off at DIS/EN pin is set to 1.2 V by the IC.

DIM

DIM is used for dimming the LED lamp by using PWM. A square wave can be applied on DIM to control the gate of Q_{12}, Fig. 11, page 19. When DIM is high, Q_{12} starts to conduct and the voltage at BP will be low. This stops the switching output at GDRV of TPS40210 and resume as soon as DIM is low. By controlling the duty cycle of the pulse applied to DIM, the average output current of the circuit can be controlled and thereby the LEDs can be dimmed. The fact that the IC just turns the switching off and not the whole circuit is a big advantage. Otherwise the driver would have to start up at every pulse of the PWM. This would make the dimming less effective and inaccurate.

3.2.4 Switching frequency

TPS40210 has a frequency range between 35 – 1000 kHz, but as a user of the IC it is important to keep in mind that this is a designed frequency range that works in theory and is not production tested by TI.

The RC pin is an input that is used to set the switching frequency. From the RC pin there is a capacitor, C_{28}, connected to ground and a resistor, R_{51}, between the RC and VDD pin, Fig. 14. These two can be set to choose the wanted frequency. The capacitor will be charged to approximately 1/20 of the voltage at V_{AUX} and then discharged trough an internal transistor in the IC. The suggested value for the capacitor is between 68 – 120 pF for most applications and by choosing a desired frequency, the resistance can be calculated using (3.2.4.b) with R_{51} in kΩ, switching frequency, f_{sw}, in kHz and C_{28} in pF. The resistance, R_{51}, is recommended to be within the range of 100 kΩ – 1 MΩ.

$$f_{SW} = \frac{\sqrt{5}\cdot\sqrt{R_{51}\cdot((4221\cdot C_{28}{}^{2}+13.5\cdot 10^{3}\cdot C_{28}+624.5\cdot 10^{3})\cdot R_{51}+4\cdot 10^{9})}-5\cdot(29\cdot C_{28}+70)\cdot R_{51}}{4\cdot R_{51}} \qquad (3.2.4.a)$$

Where f_{SW} has been solved from

$$R_{51} = \frac{1}{5.8\cdot 10^{-8}\cdot f_{SW}\cdot C_{28}+8*10^{-10}\cdot f_{SW}{}^{2}+1.4\cdot 10^{-7}\cdot f_{SW}-1.5\cdot 10^{-4}+1.7\cdot 10^{-6}\cdot C_{28}-4\cdot 10^{-9}\cdot C_{28}} \text{[10] (3.2.4.b)}$$

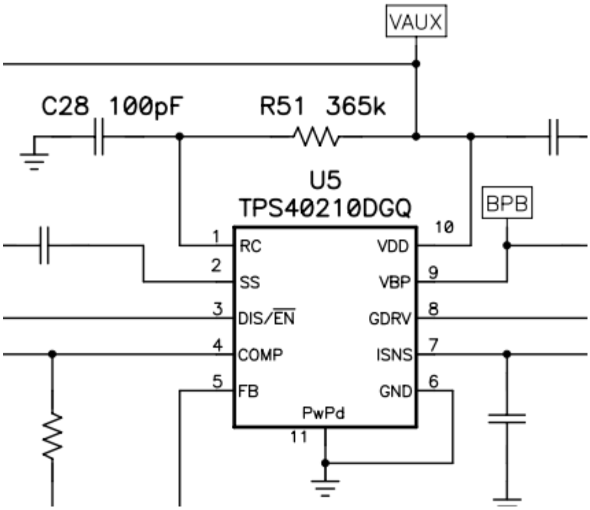


Fig. 14. R_{51} and C_{28} that can be set to get a desired switching frequency.

3.2.5 PWM frequency

The frequency of the PWM to control the dimming has to be larger than 200 Hz to avoid visible flicker and also significantly lower than the switching frequency for TPS40210. In addition to this, the duty cycle has to be long enough for the effect of the rise time to be negligible. To achieve this, the duty cycle is decided to be at least ten times the rise time of the output current. The maximum PWM frequency, $f_{max\ PWM}$, is calculated by (3.2.5.a) where t_{PWM} is the pulse time of the PWM and t_{rise} is the rise time of the output current.

$$f_{max\ PWM} = \frac{1}{t_{PWM}} \qquad (3.2.5.a)$$

where:

$$t_{PWM} = \frac{t_{rise} \cdot 10}{duty\ cycle} \qquad (3.2.5.b)$$

4 Circuit changes

To find the requirements for the three currents, 350 mA, 700 mA and 1050 mA, were calculations of the currents through the current detector, the load resistance and the overcurrent protection estimated.

4.1 The currents through the detector

Because of the PNP transistors' united base voltage (4.1.a), V_B , an expression can be made for the currents through the detector (4.1.f), I_R, with the wanted current through the LED lamp, I_L. With a common V_B, V_{BE1} and V_{BE2} will be close to equal in the identical transistors.

$$V_{B1} = V_{B2} \tag{4.1.a}$$

$$V_{B1} = V_{DC} - R_{49} \cdot I_R - V_{BE1} \tag{4.1.b}$$

$$V_{B2} = V_{DC} - R_{48} \cdot I_L - V_{BE2} \tag{4.1.c}$$

$$V_{DC} - R_{49} \cdot I_R - V_{BE1} = V_{DC} - R_{48} \cdot I_L - V_{BE2} \tag{4.1.d}$$

$$R_{49} \cdot I_R = R_{48} \cdot I_L + V_{BE1} - V_{BE2} \tag{4.1.e}$$

$$I_R \approx \frac{R_{48} \cdot I_L}{R_{49}} \tag{4.1.f}$$

4.2 The load resistance

The required current through the LED lamp can be set by adjusting the load resistance, R_{62}, Fig. 15, from its original value to make sure the maximum voltage over FP pin, 0.7 V, are reached at the right current. To estimate a value for R_{62}, Ohm's law is applied (4.2.a). The voltage drop over resistor R_{60} is very small and negligibly in the calculations.

$$R_{62} = \frac{V_{FB}}{I_R} \tag{4.2.a}$$

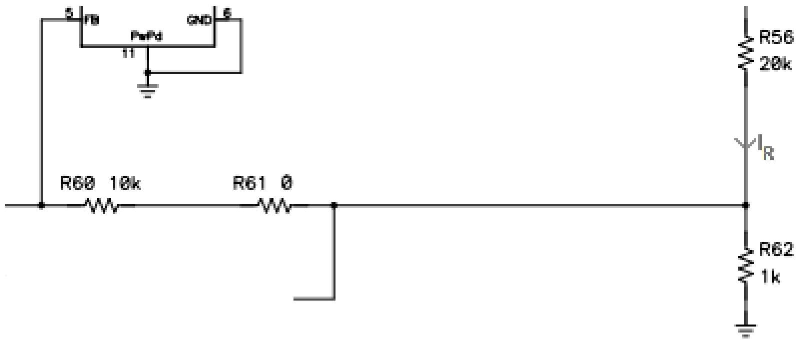


Fig. 15. Resistor R_{62} and the FB pin that regulates the fixed current through the LED lamp according to FB pin and I_R.

4.3 Overcurrent protection

The overcurrent protection is set by the ISNS pin on the IC in the driver and gets trigged when the pin's threshold voltage, V_{ISNS}, exceeds 150 mV. By putting a resistor, R_{54}, in series with the switching power MOSFET, Fig. 16, the voltage over the ISNS pin can be adjusted to the current according to (4.3.a). The result of this equation is the highest value the resistor can adopt without trig the overcurrent protection. The fixed current through the LED lamp is an average value and therefore a peak value, I_{peak}, from simulations must be used in calculations to avoid that the peaks in the ripple trigs the overcurrent protection.

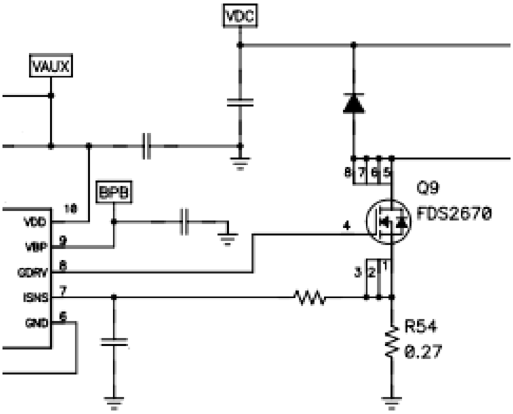


Fig. 16. The components that regulate the overcurrent protection, resistor R_{54} and the ISNS pin. [7]

$$R_{54} = \frac{V_{ISNS}}{I_{peak}} \tag{4.3.a}$$

When the overcurrent protection is trigged the capacitor at the SS pin will discharge, 3.2.1 *The IC in driver, TPS40210,* and the circuit shuts down.

5 Simulations in TINA–TI

After restrictions, simulations were only performed on the driver circuit.

5.1 Missing components in TINA–TI

Some of the components needed for the circuit was not included in TINA–TI and had to be added to the program. The easiest way to do this was to try to find a SPICE model online for the missing component and import it. The imported model is called macro. When imported to TINA–TI the component did not have any package or shape. When this occurs TINA–TI will automatically suggest a shape with the right amount of pins that can be used. There is also a possibility to choose a different one than the one suggested or draw one by hand. A lot of manufacturers make these kinds of SPICE models for their components so they could be used in many different kind of SPICE based software.

5.2 Voltage inputs

Since no changes were made in the power supply it was not simulated in TINA–TI. The two outputs of the power supply, V_{DC} and V_{AUX}, were instead replaced with voltage generators to supply the driver with the desired voltage.

5.3 Zener diode as LED substitute

During these simulations, a load consisting of a series backwards turned 4.7 V zener diodes were used. In simulations the voltage was measured to 4.8 – 5.2 V over each zener diode depending on the output current, a higher current led to a bigger voltage drop. For this reason, 16 diodes were connected in series for a maximum load of 76.8 – 83.2 V. By using 16 diodes, the load test of 100, 75, 50, 25 and 0 % could easily be executed. Zener diodes were used because it was an easy way to model the load without building a huge LED lamp with several rows of LEDs. Maximum current of the LEDs found in TINA–TI was 30 mA with a junction voltage of 1.5 V which had resulted in 35 rows with 54 LEDs in each row to be able to get the 1050 mA and 80 V.

5.4 Simulation execution

The simulations were run on 2 ms when the dimming function was turned off and up to 20 ms when the dimming was on at the lowest PWM frequency, 200 Hz. Without dimming, the measurements and data were taken after 1 ms and with dimming on after 5 ms to make sure the simulations had been stabilized. All simulations in this chapter were performed with V_{DC} at 110.3 V and V_{AUX} at 11.6 V.

The average current through the load was examined to make sure the right current was achieved. To get the average current through the load, the easiest way was to use the built–in function in TINA–TI called *averages*. The voltage at the FB and ISNS pin on TPS40210 was inspected to see that the current limitation and overcurrent protection worked as intended, Fig. 17.

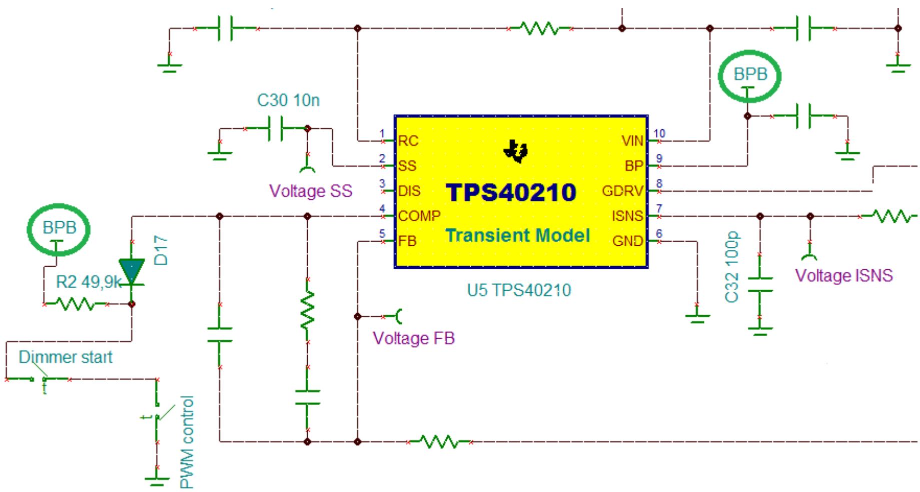


Fig. 17. Measuring points and implemented dimmer.

Measurements were made on the SS pin to see the charge and discharge of the capacitor, C_{30}, in series with the SS pin. This is shown if an overcurrent protection occurs. The overcurrent protecting should not be trigged under normal circumstances but was provoked in some simulations for analyzing by dimension R_{54} with higher value than calculated from (4.3.a).

The LED driver was first simulated with the calculated value for resistor R_{62}, as reference value, and subsequently the simulations was remade with adjustments on the resistor until the required current $I_{L-Avarage}$ was achieved. $I_{L-Avarage}$ is the measured average current after it has been stabilized between 1 and 2 ms, Fig. 18, and should be 350 mA, 700 mA and 1050 mA.

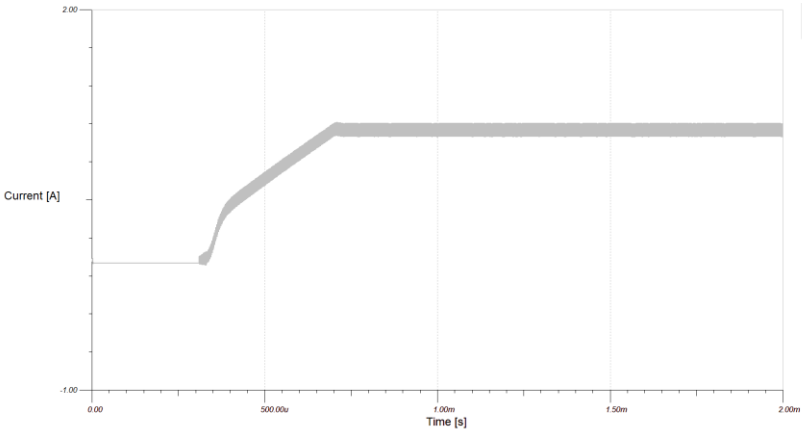


Fig. 18. Graph over the current at 1050 mA.

The dimmer function was also tested to see if it worked properly. Instead of using a pulse signal from the external control, DIM, to trig the original transistor, two time controlled switches were used. They were placed in series between the COMP pin at TPS40210 and ground, Fig. 17. The first switch, *PWM control*, was controlling the dimming by switching on and off in short preprogrammed interval. In this switch, both frequency and duty cycle can be set. The other switch, *dimmer start*, sets the start and stop time, if the dimming should be constant or just over a period of time. When both switches are closed, the voltage at BPB, which is connected to BP pin at TPS40210, goes to zero and the driver is turned off until *PWM control* is opened and the voltage at BPB is back. Compared to the previous simulations, these were made during 10 ms to make sure the driver would be turned on and off a few times and at the same time get a reliable average current. 1, 3, and 5 kHz PWM with duty cycles of 10, 25 and 50 % was used in these tests.

The rise time of the output current is always the same regardless of which dimming frequency is used. But it would most probably be affected by different current and was measured for each output current to see its duration compared to the duty cycle. It is expected to see a longer rise time related to higher output current. This is because it will take longer to reach the steady state and nothing else will be changed in the circuit which might affect the rise time. The maximum PWM frequency for each current was then calculated by (3.2.5.a), used in new simulations and compared to the previous results and to verify that the calculated frequency was good enough.

6 Efficiency and varying input voltage for the driver

To investigate the efficiency of the power supply, information was taken from TI's test result [11] since no changes or simulations were made. The driver was simulated with different switching frequency and loads to study the ripple and efficiency. Simulations with varying input voltage to the driver was also estimated to examine how the current through the load would react to disorders.

6.1 Efficiency of the power supply

From TI's test results, a graph could be obtained over the efficiency for the power supply, Fig. 19.

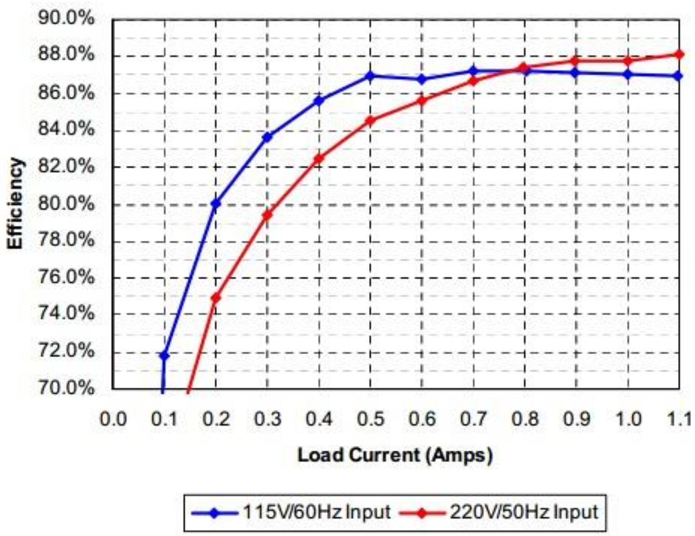


Fig. 19. The efficiency for the flyback converter in power supply. [11]

6.2 Efficiency of the driver

Simulations were executed for the three currents with a varying load between 0 – 100 % connected to the output to study the ripple relative to the size of the load and the efficiency. The ripple was then compared to the test results from TI, Fig. 20. The voltage generators had fixed input voltages of $V_{DC} = 110.3\ V$ and $V_{AUX} = 11.6\ V$. To calculate the efficiency, voltage and current meters were placed on both input voltages and also over the load. The amount of incoming energy, P_{in}, was then calculated by (6.2.b) and the amount of the output energy over the load, P_{out}, with (6.2.c). From these two, the efficiency could be calculated using (6.2.a).

$$Efficiency = \frac{P_{out}}{P_{in}} \tag{6.2.a}$$

$$P_{in} = V_{DC} \cdot I_{DC} + V_{AUX} \cdot I_{AUX} \tag{6.2.b}$$

$$P_{out} = I_{average} \cdot V_{load} \tag{6.2.c}$$

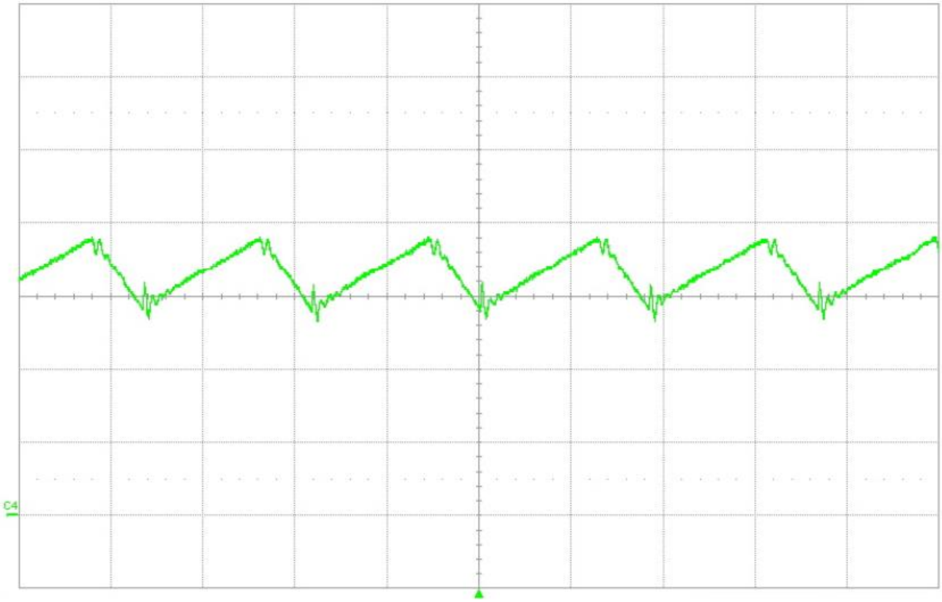


Fig. 20. Current ripple from test results made by Texas Instruments, 100 mA/Div. [11]

The TPS40210 can work in the frequency range of 35 – 1000 kHz and for this circuit by changing resistor R_{51} the lowest recommended frequency is 170 kHz [10]. This is based on the minimum value of R_{51} at 100 kΩ. The switching frequency was swept over an interval of 170 – 900 kHz to see how it affected both current ripple and efficiency. Simulations were performed in this interval for 700 mA with 100 % load.

6.3 Power losses in the driver

The main power losses is related to the switching transistor, Q_9. The switching causes small but notable rise and fall times for the voltage over the transistor in each period. This generates losses in each switch and is called *crossover losses*. These can be estimated by (6.3.a). With $V_{average}$ as the average voltage at the quite linier rise and fall time, $I_{average}$ is the current, f_{SW} the switching frequency, t the rise and fall time.

$$P_{crossover\_losses} = \left(P_{rise} \cdot t_{rise} + P_{fall} \cdot t_{fall}\right) \cdot f_{SW} \qquad [24] \qquad (6.3.a)$$

$$P_{rise} = V_{rise\_average} \cdot I_{rise\_average} \qquad (6.3.b)$$

$$P_{fall} = V_{fall\_average} \cdot I_{fall\_average} \qquad (6.3.c)$$

Other losses that are generated by the transistor are the so called *turn-on losses*. The transistor has some unwanted parasitic capacitances that will be charge every time the transistor starts to conduct. These losses can be calculated by (6.3.d).

$$P_{turn-on\_losses} = \frac{c_p \cdot V^2 \cdot f_{SW}}{2} \qquad [24] \qquad (6.3.d)$$

Where V is the drain voltage, c_p the parasitic capacitance and f_{SW} the switching frequency.

Losses will also occur over the diode, D_{15}, according to (6.3.e).

$$P_{diode\_losses} = \frac{1}{4} Q_{rr} \cdot V_{Drr} \cdot f_{SW} \qquad [24] \qquad (6.3.e)$$

Where Q_{rr} is the reverse recovery charge of the diode, V_{Drr} is the voltage across the diode at the time it switches on and f_{SW} is the switching frequency. All three losses are affected by the voltage and switching frequency. A higher switching frequency, f_{SW}, and voltage will result in more losses. A higher switching frequency or voltage should be notable as a lower total efficiency of the driver during simulations.

Losses that will not be affected by these changes are for example the heat losses in R_{48} at 1 Ω. The losses here will only be affected by the current going through it and is set by (6.4.f).

$$P_{resistor\_losses} = R \cdot I^2 \tag{6.4.f}$$

6.4 Lamp comparison

The dimming function was also tested in the simulations to see the difference in efficiency between the output current of 350 mA and a 700 mA with 50 % dimming of the load. The dimming was performed with a PWM of 200 Hz during 20 ms.

6.5 Variety of the input voltage to the driver

The test results from TI shows that the output voltage fluctuates about +/– 5 V from its initial value [11]. To keep the driver stable at all time, regardless of voltage ripple, it is of interest to know if the output current is affected by this voltage ripple. Simulations with different V_{DC} between 90 – 120 V and V_{AUX} calculated from (3.1.8.e) and 100 % load was performed for analyze how sensitive the driver was to disorders in the voltage supply. This interval should cover unpredictable variations of the voltage.

7 Components

The reference design had a Bill of Materials (BOM) with all components of the circuit [Appendix 1.A]. Farnell, Mouser and Elfa were used to find the same or similar substitutes and chosen by the following criteria:

1. Satisfied the minimum requirement/values
2. Lowest cost
3. Smallest size
4. Avoid parts planned to be obsolescence

The new BOM can be seen in Appendix A.2.

7.1 Bill of Materials (BOM)

The BOM contain the following information for each component:

Count – Number of components of the specific type
RefDes – Reference design name on the schematic
Value – Value or name of the component
Description – A short description, for example *“Diode, Zener, 18V, 3W”*
Size – Component size, such as SOT23, 0603, SMA etc.
Part number – Manufacturers part number
Mfr – Manufacturer

8 The LED lamp

A provisional 80 W LED lamp was constructed and built to be able to test the power supply and driver.

8.1 LED lamp design

The criteria was an 80 V LED source that would work in the range from the lowest current, 350 mA, up to the highest, 1050 mA. A small LED lamp, MX–6S, from CREE was chosen for this purpose [12]. The LED lamps were chosen because of their characteristics of the forward current and forward voltage, Fig. 21. These were suitable because of their high efficiency with a forward voltage of 21.5 V over each small LED lamp with a forward current of 100 mA [12]. By dividing the current in 10 rows with 4 small LED lamps in each row an adequate LED lamp, Fig. 22, could be constructed.

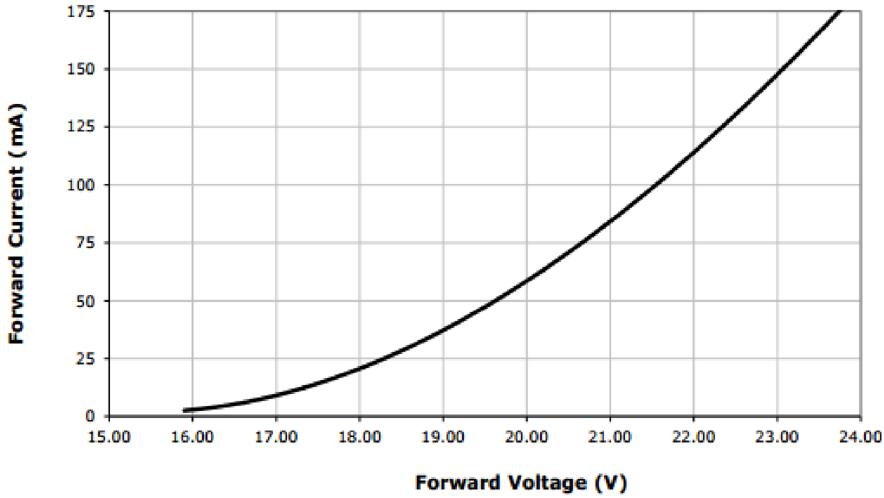


Fig. 21. Characteristics for the small LED lamp. [12]

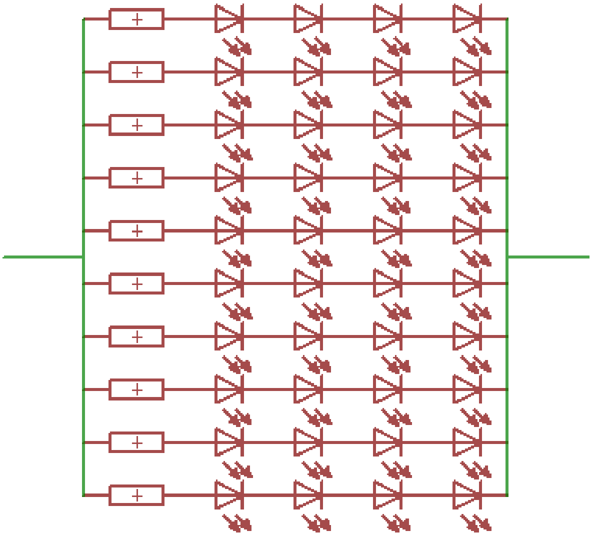


Fig. 22. Schematic over the LED lamp.

To avoid an uneven distribution of the current between the rows in the LED lamp, one resistor was added to each row. The resistors were chosen after consultation with one of the advisors for the master thesis [13] and decided to be 20 Ω with minimum 0.22 W power tolerance. By adding resistors, the small LED lamps were a bit more protected from uneven current distribution with high current in some rows and lower in others. If one LED lamp breaks the current would rise in the other rows and the beginning of a vicious circle starts. The rows in the LED lamp would shine with different strength and be unpleasant to look at. The fact that the current in each row could be measured by studying the voltage drop over each resistor, and then by Ohm's law calculate the current, was another advantage by adding resistors. This was a temporary solution of building an LED lamp, better methods are discussed in 13.7 *The LED lamp design*.

8.2 Construction and cooling of the LED lamp

Each LED will have an effect of 2 W and generate a lot of heat. This was taken into consideration when designing the board layout for the lamp. CREE has a recommendation of a two layer board with thermal dissipation through vias to the bottom layer and use of a heat sink. But at the time of the production of the lamp there was only possible to make a one layer design without thermal vias. With no vias CREE shows that a possible thermal resistance through a 0.8 – 1.6 mm FR4, glass–reinforced epoxy laminate, board would be between 21 – 32 °C/W [14] with a top trace larger than 15 mm. The board used to make the lamp was measured to be 1 mm thick and was estimated to 25 °C/W, Fig. 23. This is still a very high value and by using thermal vias the resistance could be as low as 5-7 °C/W [14], depending on the number of vias and their diameter.

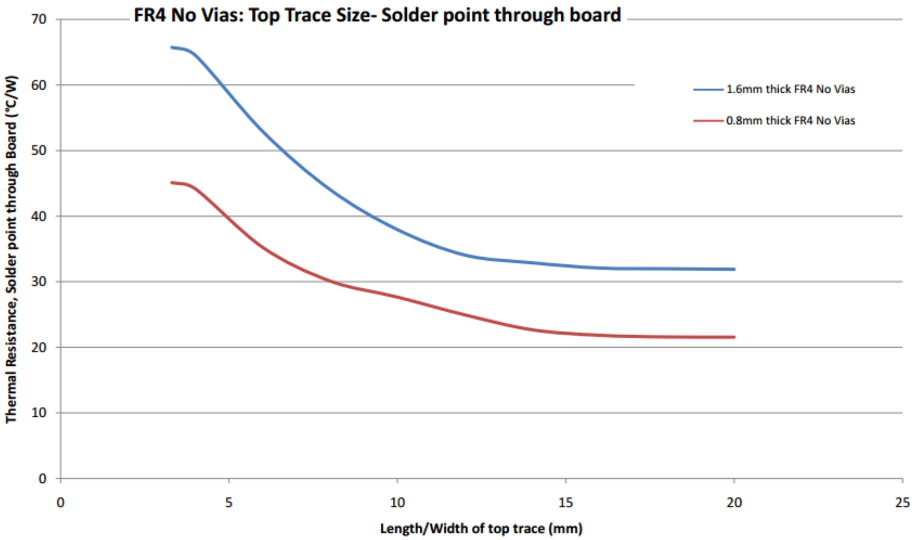


Fig. 23. Thermal resistance for FR4 PBC without vias. [14]

The size of the top layer was fitted for the biggest accessible PCB, 16 x 26 cm, in the etch laboratory. The finished LED lamp was then mounted with isolated screws on an 18 x 30 cm heat sink, Fig. 24. Since the heat conduction goes both laterally and perpendicularly it was hard to achieve any reliable calculations due to the lack of percentage heat distribution. A thermal imager was used and the lamp was supplied from a power box at 1 A to get a perspective of the heat conductions.

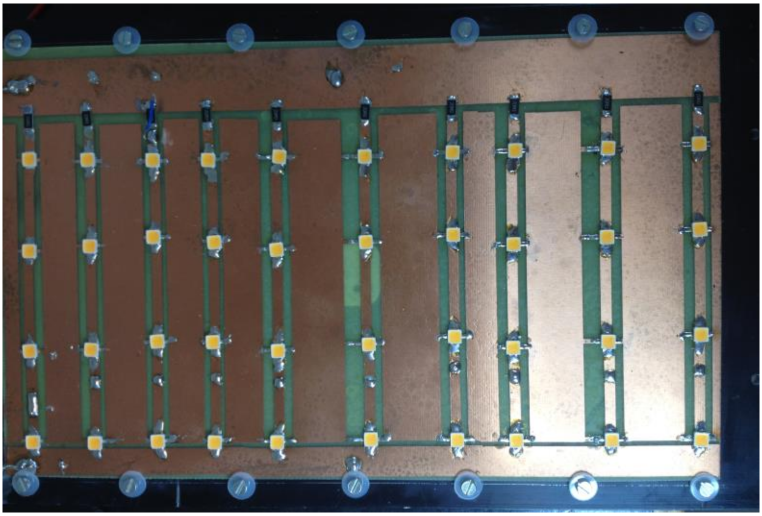


Fig. 24. The LED lamp mounted on a heat sink.

8.3 Eye safety

These LEDs gives a warm white light and is a more pleasant light for the eyes than many others. Even though it is a nicer light, the eyes should always be protected when working with any operating LEDs. If the intensity is high enough all light sources might be harmful for eyes and/or skin. It is important to never look straight into an operating LEDs. [15]

9 Bread board laboratory work

Due to the fact that the transformer was bespoken from China it was decided to build up the circuit on bread boards for testing the transformer. All the components datasheets were examined to find through hole substitutes. Those components without through hole options was ordered as Surface Mounted Device (SMD). Footprints for each component were made in EAGLE, 10.4.1 *EAGLE* page 48, and bitwise produced on small Printed Circuit Boards (PCB) and then attached with cables for connection to the bread board.

The positions of the components were noted on a bread board schematic with the coordinates for each component marked in the circuit's schematic for easy identifying of a component. The power supply and the driver were built on separate bread boards for simplification of testing each part.

Knowing that couplings on a bread board would result in a lot of extra noise, it was attempted to prevent unnecessary distortions by routing high current wires as short as possible, routing ground, V_{DD}, and the LED connections close together to avoid a big induction loop. The grounds in the driver should be routed with the ground of the transistor closest to the reference ground and then the IC's ground to avoid disorders in the ground trace. While investigating the driver, several capacitances were added, or changed to a higher value. This was an attempt trying to reduce noises on the inputs of FB and ISNS to prevent trigging the overcurrent protection.

The power supply and the driver were both tested separately as well as together.

10 Printed Circuit Board (PCB)

A PCB is used to build electronic devices and connects electrical components through conductive traces. The layout is made in a CAD software program and after the manufacturing, only the pads and holes for the components are exposed while all routing is covered under a green soldering mask. Connections between the layers are made with vias, plated–through holes.

10.1 Designing a PCB

There are many things to keep in mind when designing a PCB. The strength of the current, trace width, how things should be connected to ground, should ground plane be used and the best placement of the components.

10.1.1 Different kind of software

There are a lot of different kinds of software when it comes to designing a PCB layout. KiCAD, EAGLE CAD, TinyCAD, FreePCB is just a few, some of them are free and others are really expensive. It is important to know that some of the programs, especially the freeware, might have limitations in number of pins or layers, if the program has auto routing and have platforms dependence. One of the first things to do while designing a PCB is to look up some manufacturers to see their size and clearance criteria for manufacturing PCBs. There are software that are directly connected to manufacturer which might be a good choice if that is something the manufacturer offers. If this is the case, the manufacturer's limitations are already applied such as minimum drill size, trace width and clearance between two traces etc. If not, it is always a good idea to look up what limitations the manufacturer has before starting to make the layout. By knowing this, the design rules can be set at the start and a lot of time can be saved in the end.

10.1.2 Trace width and area

The larger the current that flows in the traces is, the smaller the areas within the traces have to be to avoid an unnecessary big magnetic field that causes noise in the circuit. It is also essential to have these high current traces close together, and separated from the rest of the circuit if possible, to avoid high voltage and current spikes to affects other parts.

A larger width of a trace will decrease the resistance for the current and thereby also reduce the heat. The trace width has to be calculated to make sure it can handle the maximum current that flows through it. If the trace is too thin the heat might get to a point where the traces burn.

Even a small difference in a trace width, like when making a turn, may cause problem. By that reason it is important to keep the wires' width constant by avoid sharp angles. A recommendation is to not make an angle sharper than 45° to maintain the width [16]. The PCB will also have a more professional look and by avoiding right angle the length on the traces can be shortened. Most of the CAD software has a setting for this that will make all traces with a 45° angle.

Not only high current needs to be taken into consideration, even high voltage can cause problems. It is crucial to make sure the spacing between high and low voltage paths, for example input voltage and low potential such as ground, is big enough to avoid sparks that may cause short circuits on the board.

The more pins a component has, the more space on the board will be needed around it since each pin will be connected to a wire and drawn out from the pads. By not putting components with many pins too close to each other, it will make the routing on the PCB and any manual soldering much easier.

To avoid cross over lines it is recommended to draw the wires horizontal on one layer and vertical on the next, and if several layers are used continue alternating between each layer [17].

Traces to ground will always have a resistance and causes a small voltage drop. This may end up in small voltage variations in the ground at different places in the circuit. It is good idea to create a ground plane where the whole layer or big parts of it is covered with copper. Unnecessary traces to ground can then be avoided and components can be directly connected to ground through vias. A ground plane will also avoid big loops and reduce unwanted noise.

Another advantage of a ground plane is the heat dissipation. The bigger the layer of copper is the bigger heat dissipation. However, this can be a problem when soldering the components but can be reduced by having the ground plane as a square grid with thin solid copper plane and resulting in easier soldering. There is also an option to use a solid layer but using small grids around the vias to simplify the soldering, so call heat relief pads, Fig. 25.

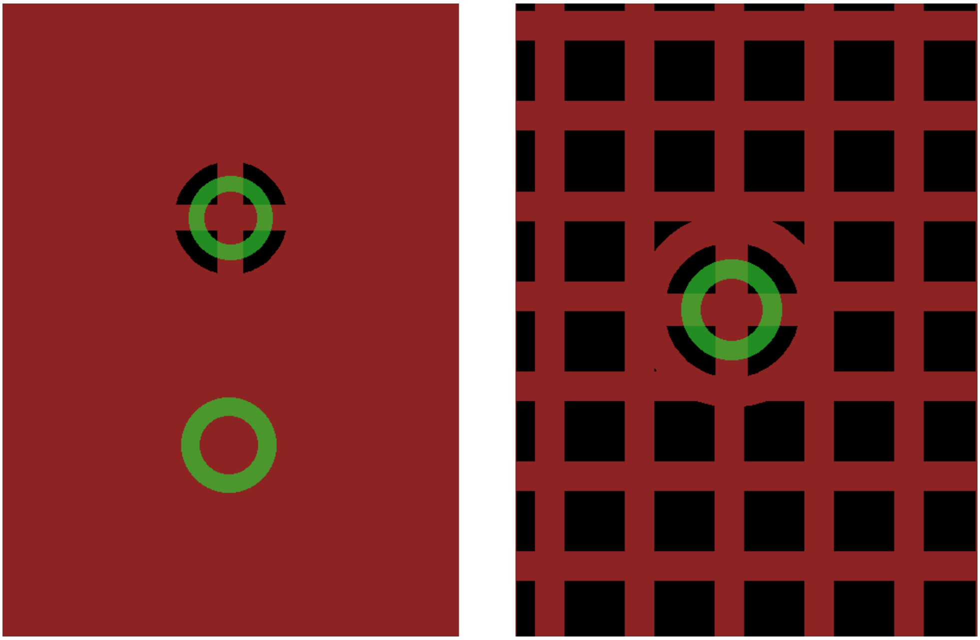


Fig. 25. A comparison of a solid ground layer to the left and a square grid layer to the right. Upper left via shows the opportunity to use a heat relief pad.

By paying attention to the components thermal resistance, components that become warmer should be isolated from sensitive components. If the heat dissipation on the PCB is not enough, a heat sink needs to be added.

10.2 Gerber format

When the layout is done and it is time for manufacturing, the PCB Gerber files need to be created. Gerber is a standardized file format used worldwide by PCB manufacturer. Those files contain different information such as were the pads, copper and solder layer, should be. The drill holes' size and location are stored in special files called Excellon files. An example of a Gerber file can be seen in Fig. 26 where the pads of our design are displayed.

Fig. 26. A Gerber file containing information about where pads are located on our PCB design.

10.3 Soldering

It is important to take care when manually soldering the components on the PCB. Always start with the smallest components that usually are Surface Mount Devices (SMD) and end up with the biggest through hole components to make sure no component covers the access to another pad.

10.3.1 Soldering technique

A flux pen is a no–clean pen which means it does not leave any residue behind and it makes soldering much easier. By removing the oxide layer from the metal the hot solder can get stuck directly on the pin. The soldering must take place directly after applying the flux because it starts to vaporize immediately.

10.3.2 Drag soldering

When soldering a SMD with several pins, such as SOIC–8 package, it is suggested to first solder two legs so it can not move when the rest of the pins are soldered. After putting flux over the pins the solder tip should be applied with a little solder and quickly be dragged over the pins to equally spread the solder. To make sure there are no short circuits a magnifier can be used for inspection.

It is crucial to be careful with the amount of solder applied, too much can create bridges between the pins and cause short circuits. The bridges can be removed with solder wick but will take some extra time. A faster way is to use the flux. Just apply some extra flux and drag the soldering tip over the pins, avoid pressing too hard as it can cause damage to the component.

10.4 The PCB layout in EAGLE

One of the requirements of this thesis was to make the PCB layout in EAGLE CAD. LTH was able to provide a Standard Education license to be used instead of the limited free version.

10.4.1 EAGLE

The reference design, PMP4862, had a four layer design and its dimension was 130 x 90 mm. EAGLE, with the Standard Education license, was able to handle six layers and 160 x 100 mm routing area. Several components on the BOM were not to be found in the standard libraries. Some of them could be found on various places on the internet, for example at www.farnell.com. Those not found had to be made manually by information collected from datasheets of each component to design the package, known as footprint, Fig. 27, and the symbol, Fig. 28. The footprint is used in the board layout and the symbol in the schematic. The footprint, the symbol and simulation data is assembled to a component called device.

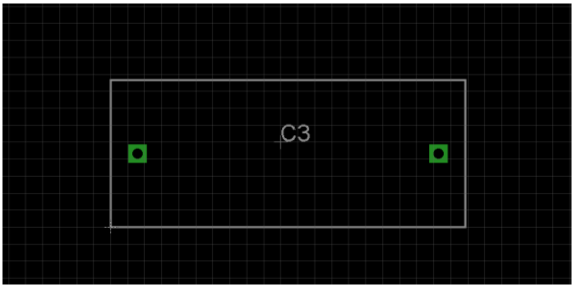


Fig. 27. Footprint for the capacitor C_3 in board layout.

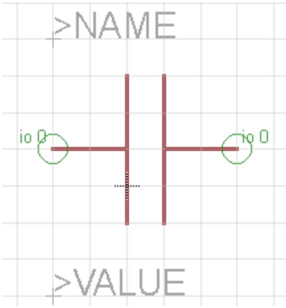


Fig. 28. Symbol for a capacitor in the schematic layout.

10.4.2 Connection check and routing

When the schematic was done, with all the right component sizes, an Electrical Rule Check (ERC) had to be run in EAGLE to check for errors. The ERC are able to show a lot of different *errors* and *warnings*. Some of the warnings, for example "part D1 has no value", can be ignored since this schematic is not supposed to be used in any simulations. Errors on the other hand are more critical and have to be taken care of before the layout and routing on the PCB starts.

The fewer components on the new layout, due to the fact that only one driver is used, made room for using a two layer design. Layer 2, 3 and 4 from the reference design were combined into just one layer, the bottom layer, and the design could now be done with only two layers. Two ground planes were used on each layer, GND and AGND. It has to be guaranteed that the top and bottom ground planes are well connected through several vias, so there are no areas of a ground planes trapped inside traces that has no connection to the whole ground plane.

Most of the components were placed similar to the way they were placed on the reference design. Some changes had to be made due to different package sizes than the reference circuit's components and also because there was some components missing in the reference design. Some of the routing in the reference design was connected wrong by TI, this made it extra important to be observant when making the new routing. The TI reference driver is also based on a third of the current through the LEDs in this design and because of this, high current traces were made wider in the driver for routing higher currents.

The traces width was calculated by (10.4.2.a)

$$Width_{mm} = {\frac{I}{k \cdot T_{rise}{}^{b}}}^{1/c} \cdot \frac{1}{t_{\mu m} \cdot 1.550} \qquad (10.4.2.a)$$

with the k, b and c as constants [18], T_{rise} as the accepted temperature rise in the trace given in °C, *I* the current through the trace given in ampere and $t_{\mu m}$ as the copper thickness in µm. (10.4.2.a) is converted from the original equation (10.4.2.b) to be able to give the answer in mm instead of mils, 1/1000 of an inch.

$$Width_{mils} = {\frac{I}{k \cdot T_{rise}{}^{b}}}^{1/c} \cdot \frac{1}{t_{oz} \cdot 1.378} \qquad [18] \qquad (10.4.2.b)$$

t_{oz} is here given in ounces per square foot and 1.378 is the thickness in mils of 1 ounce of copper which is also equal to 35 µm.

There is also a command, *run length–freq–ri*, which can be given in EAGLE to check the maximum current in each trace on the board when the layout is finished. The resistance, maximum and minimum, width in each trace can also be seen in Fig. 29. Every row specify a single trace and *Signal* is the name, *Imax* the maximum current in ampere, *R* the resistance in mΩ, *w min* the minimum width and *w max* is the maximum width in mm.

Cu thickness = 0.035 mm

Signal	Imax [A]	R [mOhm]	w min [mm]	w max [mm]
N$1	5.23	1.53	1.778	1.778
N$10	5.23	1.17	1.778	1.778
N$2	5.23	6.69	1.778	1.778
N$21	5.23	1.28	1.778	1.778
N$4	5.23	7.83	1.778	1.778
N$5	5.23	11.02	1.778	1.778
N$73	5.23	9.97	1.778	1.778
N$9	4.94	10.14	1.676	1.778
N$64	4.20	1.58	1.422	1.422
N$65	4.20	4.28	1.422	1.422
D4	3.75	2.19	1.270	1.270
N$68	3.75	0.94	1.270	1.270
N$69	3.75	0.93	1.270	1.270
N$7	3.75	1.80	1.270	1.270
N$70	3.75	0.94	1.270	1.270
N$3	3.01	37.49	1.016	1.778
N$43	3.01	1.18	1.016	1.016
N$44	3.01	7.03	1.016	1.016

Fig. 29. Some of the information given by use of the command *run length–freq–ri* in EAGLE.

Not only had the current through the traces be taken into consideration, even the voltage potential between two adjacent traces was of importance. To avoid leakage current between the traces, clearance had to be wide enough. By following the recommended clearance from Association Connecting Electronics Industries (IPC) [18] the clearance of each trace were designed to minimize the risk of leakage current and short circuits.

The reference design used a lot of “short cuts” by routing traces under and between the pads of other components. Two examples of this can be seen in Fig. 30. One of the outputs from TPS40210 is routed under R_{53} and one of the pads on C_{29} has a trace going under C_{31}. This way of routing was avoided during the whole layout process for two reasons, the risk of short circuits when soldering and also clearance between pads and traces would be too small for the manufacturer. The minimum clearance for the manufacturer was 0.20 mm, and by routing under other components the clearance could be as small as 0.15 mm. Instead the routing was done around the components when it was possible and for some cases the bottom layer was used to connect the pads by going through vias.

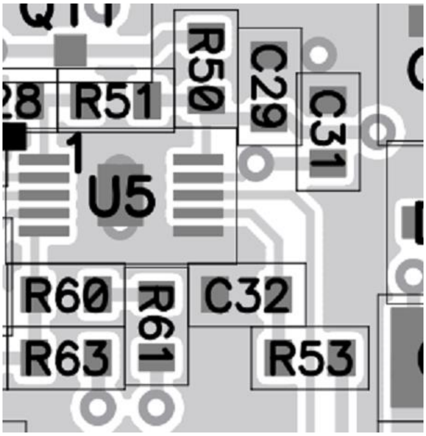


Fig. 30. Routing extracted from the reference circuit. [19]

When many pads where connected close to each other on single trace for high current, small planes around the pads where made instead of traces.

10.4.3 Jumpers for changing current

The circuit should be able to handle three output currents by using different values of R_{62}. To be able to use the same circuit for all three currents, jumpers was implemented, one for each resistor. By doing it this way, the current can easily be chosen by moving the jumper head to the jumper attached to connect the required resistor, Fig. 31.

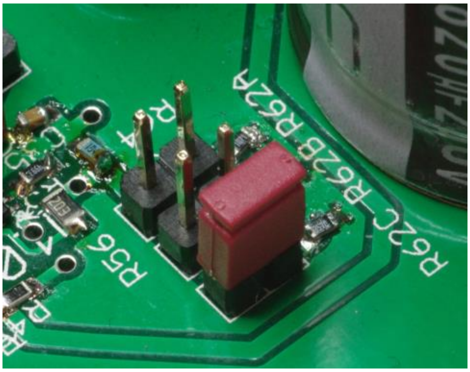


Fig. 31. The jumper solution for switching between the desired currents.

10.4.4 Test points

Three extra test points, in addition to those who already existed, were added to the schematics to be able to test the circuit. They were situated on locations based on where the measuring on the bread board prototype gave the most essential values for analyzing and troubleshooting of the circuit. The first one was placed on the output of DRV on UCC28810 in the power supply, Fig. 32, to be able to see if the switching worked as intended. The second one was placed on the FB pin on TPS40210, Fig. 33, to see how the feedback worked and if the input voltage would exceed the limitation of 0.7 V. The last test point was placed on the source of MOSFET, Q_9, to be able to see the switching of the driver, Fig. 33.

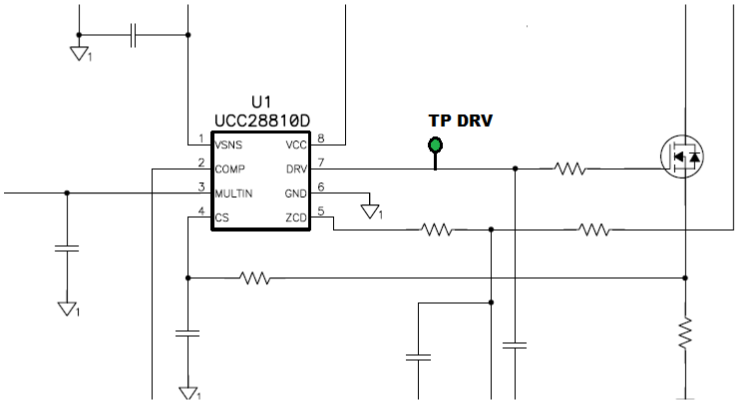


Fig. 32. The test point on the gate driver in the power supply.

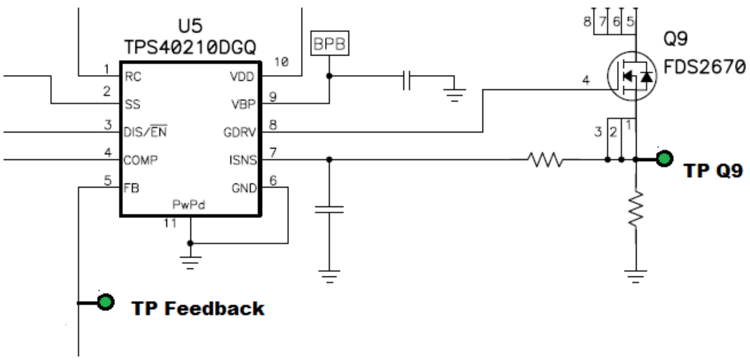


Fig. 33. The test points on the feedback pin and after the switch in the driver.

11 Testing the PCB

With all the components soldered to the PCB, the circuit was tested to see how and if it worked as planned.

The circuit was tested part by part to see how it behaved at different locations and to make it easier to troubleshot the circuit. Places with known voltages, like V_{DD} and V_{AUX}, were sometimes replaced with fixed voltages from an external power supply. By doing this, parts could be tested without voltage issues at these places.

By applying a small voltage of 2.5 V at the external control, ENABLE, to the gate of Q_{11} so it started to conduct, the DIS/EN pin on TPS40210 was pulled low to get the driver circuit enable.

To limit the incoming current two 15 Ω aluminum housed power resistors, each at 50 W, were connected in parallel just before the input of the line voltage in the power supply. These were mounted on a big heat sink for great heat dissipation.

A 40 W lamp was added between V_{DC} and ground as precaution when the driver was disable. This was done because we wanted to have some kind of load connected to the circuit and minimize the risk of broken components. The lamp also worked as a visual feedback.

12 Results

12.1 Output voltage from power supply, V_{AUX} and V_{DC}

V_{DC} was calculated by knowing that $V_{ref\ U23}$ is 2.5 V [9] by (3.1.8.a).

$$V_{DC} = \frac{R_9+R_{23}+R_{28}}{R_{28}} \cdot V_{ref\ U23} \quad (3.1.8.a)$$

$$V_{DC} = 110.3\ V$$

Where:
$V_{ref\ U23} = 2.5\ V$
$R_9 = 49.9\ \Omega$
$R_{23} = 1\ M\Omega$
$R_{28} = 23.2\ k\Omega$

The voltage drop from the diode, D_1, was added to get the right voltage at the sec_1 winding of the transformer.

$$V_{sec1} = V_{DC} + V_{D1} \quad (3.1.8.b)$$

$$V_{sec1} = 111.0\ V$$

Where:
$V_{D1} = 0.7\ V$ @ 1 A [20]

The ratio between the two secondary windings, sec_1 and sec_2, were then given by

$$ratio_{sec1/sec2} = \frac{ratio_{sec1}}{ratio_{sec2}} \quad (3.1.8.c)$$

$$ratio_{sec1/sec2} = 9.09$$

where:
$ratio_{sec1} = 0.5$ [Appendix B.2]
$ratio_{sec2} = 0.055$ [Appendix B.2]

to get the voltage

$$V_{sec2} = \frac{V_{sec1}}{Ratio_{sec1/sec2}} \tag{3.1.8.d}$$

$$V_{sec2} = 12.3\ V$$

and simply subtract the voltage drop from D_2

$$V_{AUX} = V_{sec2} - V_{D2} \tag{3.1.8.e}$$

$$V_{AUX} = 11.6\ V$$

where:
$V_{D2} = 0.7\ V$ @ 10 mA [21]

12.2 Ratio in the current detector

The ratio of the current detector sets a regulated current, I_R, (3.2.2.a).

$$I_R = \frac{R_{48}}{R_{49}} \cdot \left(I_{ref} + I_L\right) \tag{3.2.2.a}$$

$$I_R = \frac{1}{499} \cdot \left(I_{ref} + I_L\right)$$

Where:
$R_{48} = 1\ \Omega$
$R_{49} = 499\ \Omega$

12.3 Calculated value for the frequency

The calculated switching frequency, f_{SW}, in kHz for the circuit with R_{51} in kΩ and C_{28} in pF is given by (3.2.4.a).

$$f_{SW} = \frac{\sqrt{5} \cdot \sqrt{R_{51} \cdot \left(\left(4221 \cdot {C_{28}}^2 + 13.5 \cdot 10^3 \cdot C_{28} + 624.5 \cdot 10^3\right) \cdot R_{51} + 4 \cdot 10^9\right)} - 5 \cdot \left(29 \cdot C_{28} + 70\right) \cdot R_{51}}{4 \cdot R_{51}} \tag{3.2.4.a}$$

$$f_{SW} = 433\ kHz$$

Where:
$R_{51} = 365$ kΩ
$C_{28} = 100\ pF$

12.4 Circuit changes

12.4.1 Calculations for the current detector

The three currents on the reference side in the current detector, I_{ref}, can be expressed by (3.2.2.b), is almost identical to I_C and adequately small to be excluded in the calculations.

$$I_{ref} = \frac{V_{DC} - R_{48} \cdot I_L - V_{BE2}}{R_{57}} \tag{3.2.2.b}$$

$I_{ref\_350mA} = 0.219\ mA$
$I_{ref\_700mA} = 0.218\ mA$
$I_{ref\_1050mA} = 0.217\ mA$

Where:
$V_{DC} = 110\ V$
$R_{48} = 1\ \Omega$
$R_{57} = 499\ k\Omega$
$V_{BE2} = 0.6\ V$
$I_{L\_350mA} = 350\ mA$
$I_{L\_700mA} = 700\ mA$
$I_{L\_1050mA} = 1050\ mA$

An expression for the current through the current detector, I_R, (4.1.f) were established on the common base voltage, V_B, (4.1.a).

$$V_{B1} = V_{B2} \tag{4.1.a}$$

$$V_{B1} = V_{DC} - R_{49} \cdot I_R - V_{BE1} \tag{4.1.b}$$

$$V_{B2} = V_{DC} - R_{48} \cdot I_L - V_{BE2} \tag{4.1.c}$$

$$V_{DC} - R_{49} \cdot I_R - V_{BE1} = V_{DC} - R_{48} \cdot I_L - V_{BE2} \tag{4.1.d}$$

$$R_{49} \cdot I_R = R_{48} \cdot I_L + V_{BE1} - V_{BE2} \tag{4.1.e}$$

$$I_R \approx \frac{R_{48} \cdot I_L}{R_{49}} \tag{4.1.f}$$

$I_{R\_350mA} \approx 0.701\ mA$
$I_{R\_700mA} \approx 1.403\ mA$
$I_{R\_1050mA} \approx 2.104\ mA$

Where:
$R_{48} = 1\ \Omega$
$R_{49} = 499\ \Omega$
$I_{L\_350mA} = 350\ mA$
$I_{L\_700mA} = 700\ mA$
$I_{L\_1050mA} = 1050\ mA$
$V_{BE1} \approx V_{BE2}$

12.4.2 Calculations for the load resistance

Using Ohm's law (4.2.a) to calculate the value for R_{62} by knowing the reference voltage, V_{FB}, for TPS40210 and the calculated value for current through the current detector, I_R.

$$R_{62} = \frac{V_{FB}}{I_R} \tag{4.2.a}$$

$R_{62\_350mA} = 998\ \Omega$
$R_{62\_700mA} = 500\ \Omega$
$R_{62\_1050mA} = 333\ \Omega$

Where:
$I_{R\_350mA} = 0.701\ mA$
$I_{R\_700mA} = 1.403\ mA$
$I_{R\_1050mA} = 2.104\ mA$
$V_{FB} = 0.7\ V$ [10]

12.4.3 Calculations and decision for the overcurrent protection

The highest value the resistance R_{54} can take without trig the overcurrent protection for the three currents (4.3.a).

$$R_{54} = \frac{V_{ISNS}}{I_{peak}} \tag{4.3.a}$$

$R_{54\_350mA} = 376\ m\Omega$
$R_{54\_700mA} = 202\ m\Omega$
$R_{54\_1050mA} = 139\ m\Omega$

Where:
$V_{ISNS} = 150\ m\mathrm{V}$
$I_{peak\_350mA} = 399\ mA$
$I_{peak\_700mA} = 741\ mA$
$I_{peak\_1050mA} = 1080\ mA$

For keeping the design simple, the resistor R_{54} will have the same value for all three currents and by considering the availability of resistors at the suppliers, it was set to 100 mΩ. According to (4.3.a) the overcurrent protection will get trigged at 1500 mA for all three currents.

12.5 Simulations

All three currents could each be perfectly simulated in TINA–TI with precise result for the current through the LED lamp.

12.5.1 Overcurrent protection in simulation

To trig the overcurrent protection at the ISNS pin during simulations, resistor R_{54} was increased above the calculated value in (4.3.a), page 29, for the chosen current. Voltage charge and discharge of capacitor C_{30} after trigging the overcurrent protection can be seen in Fig. 34. When discharged, it tried to restart the circuit and a new overcurrent protection state occurred. This pattern was repeated over and over again until the simulation was done.

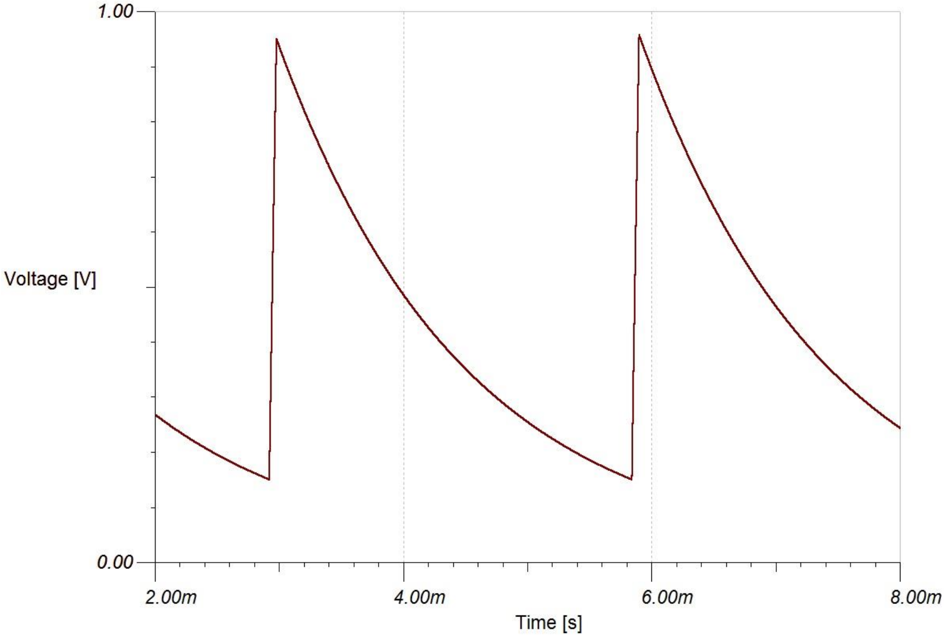


Fig. 34. The charge and discharge of the capacitor C_{30} connected to the SS pin on TPS40210 when overcurrent occurs.

12.5.2 Simulated value for the load resistance, R_{62}

The results after simulating with different values for the load resistance, R_{62}, in TINA–TI can be seen in Table 1, 2 and 3. The calculated values from (4.2.a) is the first value in each table, marked with italic, and the chosen resister value in bold. All simulations were made with a fixed value for R_{54} at 100 mΩ.

Table 1. Measurements from simulation for different resistor R_{62} for reaching an average current of 350 mA.

R_{62_350mA} [kΩ]	$I_{L-Average}$ [mA] $V_{DC} = 110.3\ V$
0.998	376
1.01	372
1.05	358
1.07	351
1.08	348
1.1	342
1.15	326

Table 2. Measurements from simulation for different resistor R_{62} for reaching an average current of 700 mA.

R_{62_700mA} [Ω]	$I_{L-Average}$ [mA] $V_{DC} = 110.3\ V$
500	747
510	733
520	719
530	705
534	700
535	699
540	692

Table 3. Measurements from simulation for different resistor R_{62} for reaching an average current of 1050 mA.

R_{62_1050mA} [Ω]	$I_{L-Average}$ [mA] $V_{DC} = 110.3\ V$
333	1112
340	1089
350	1059
353	1050
354	1047
355	1044
360	1030

12.5.3 The simulated frequency

Simulated switching frequency for R_{51} at 365 kΩ, C_{28} at 100 pF, V_{AUX} at 11.6 V and V_{DC} at 110.3 V gave a frequency of 465 kHz.

12.5.4 Current ripple and efficiency for different frequencies

The ripple and efficiency of the output current for 700 mA at 100 % load with different switching frequencies, f_{SW}, can be observed in Table 4, Fig. 35 and Fig. 36 where the dots represents the circuit's original switching frequency and its corresponding values. The relation between the ripple and the efficiency can be seen in Fig. 37.

Table 4. The ripple and efficiency for different switching frequencies at 700 mA and 100 % load. This circuit's original switching frequency is marked in bold.

f_{SW} [kHz]	I_{Δ} [mA]	Efficiency
170	268	96.1 %
200	227	95.7 %
260	171	94.7 %
340	131	93.9 %
420	101	92.8 %
465	**95**	**92.3 %**
530	82	91.3 %
680	64	90.7 %
770	55	88.2 %
900	48	86.7 %

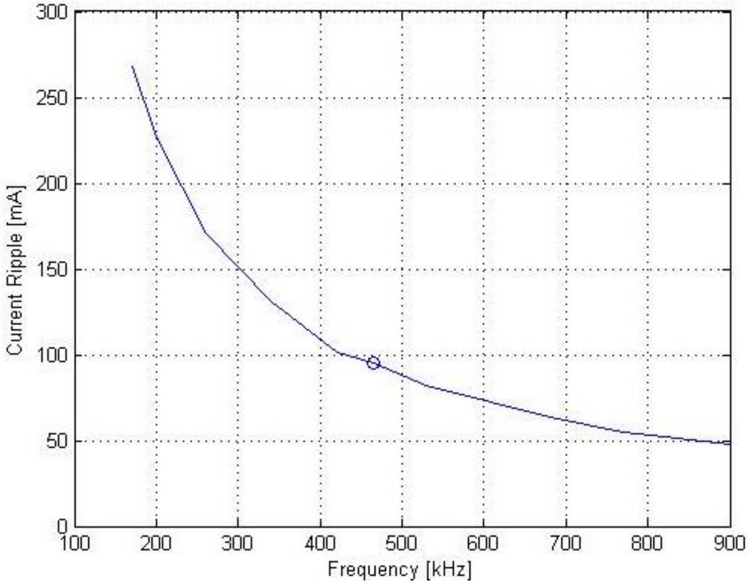


Fig. 35. The graph shows the current ripple related to the switching frequency at 700 mA. The dot represents the circuit's original simulated switching frequency.

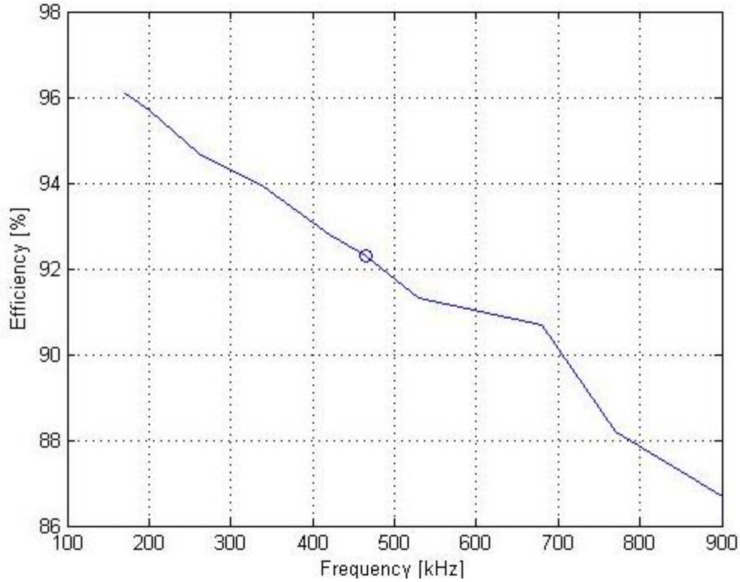


Fig. 36. The graph shows the efficiency related to the switching frequency at 700 mA. The dot represents the circuit's original simulated switching frequency.

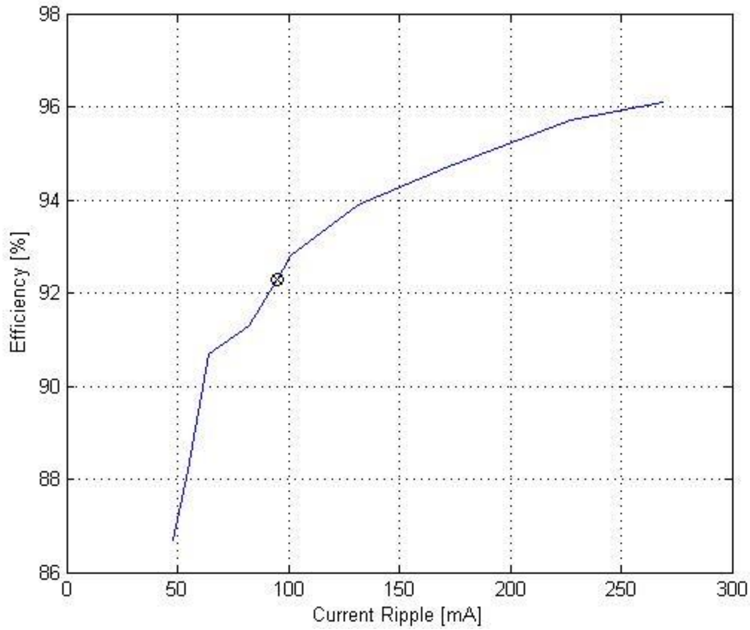


Fig. 37. A comparison between the efficiency and current ripple. The dot represents the circuit's original simulated switching frequency for this circuit.

12.5.5 Dimming

Test of the dimming function were made during 10 ms simulations with 1, 3 and 5 kHz PWM, f_{PWM}, and 10, 25, 50 and 100 % duty cycle. The current through the load with a 50 % duty cycle at 5 kHz can be seen in Fig. 38 and the dimming simulations and the results are represented in Table 5, 6, 7.

Table 5. Dimming for 350 mA.

f_{PWM} [Hz]	Duty cycle	Average current [mA]	Percentages of max current
5000	100 %	350	100.0 %
5000	50 %	165	47.1 %
5000	25 %	78	22.3 %
5000	10 %	26	7,4 %
3000	50 %	169	48.3 %
3000	25 %	82	23.4 %
3000	10 %	29	8.3 %
1000	50 %	173	49.4 %
1000	25 %	86	24.6 %
1000	10 %	33	9.4 %

Table 6. Dimming for 700 mA.

f_{PWM} [Hz]	Duty cycle	Average current [mA]	Percentages of max current
5000	100 %	700	100.0 %
5000	50 %	323	46.1 %
5000	25 %	149	21.3 %
5000	10 %	46	6.5 %
3000	50 %	334	47.7 %
3000	25 %	154	22.0 %
3000	10 %	54	7.7 %
1000	50 %	345	49.2 %
1000	25 %	170	24.2 %
1000	10 %	65	9.3 %

Table 7. Dimming for 1050 mA.

f_{PWM} [Hz]	Duty cycle	Average current [mA]	Percentages of max current
5000	100 %	1050	100.0 %
5000	50 %	468	44.6 %
5000	25 %	206	19.6 %
5000	10 %	057	5.4 %
3000	50 %	490	46.7 %
3000	25 %	228	21.7 %
3000	10 %	72	6.9 %
1000	50 %	514	49.0 %
1000	25 %	252	24.0 %
1000	10 %	95	9.0 %

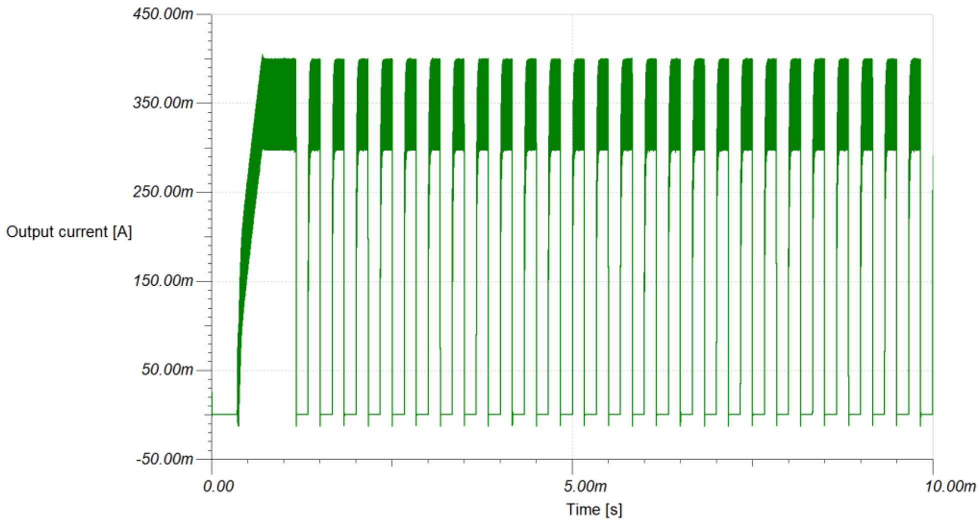


Fig. 38. Simulation of the dimming at 5 kHz and 50 % duty cycle for 350 mA.

12.5.6 Rise time for the PWM dimming

The rise time, t_{rise}, of each output current is presented in Table 8.

Table 8. The rise time for the three currents.

Current [mA]	t_{rise} [µs]
350	14
700	22
1050	26

12.5.7 Maximum PWM frequency

The maximum PWM frequency was calculated for each current at the shortest duty cycle of 10 % (3.2.5.a).

$$f_{max\ PWM} = \frac{1}{t_{PWM}} \qquad (3.2.5.a)$$

$$t_{PWM} = \frac{t_{rise} \cdot 10}{duty\ cycle} \qquad (3.2.5.b)$$

$f_{\max\_PWM\_350mA} = 714$ Hz
$f_{\max\_PWM\_700mA} = 454$ Hz
$f_{\max\_PWM\_1050mA} = 384$ Hz

Where:
t_{rise_350mA} = 14 µs
t_{rise_700mA} = 22 µs
t_{rise_1050mA} = 26 µs
$duty\ cycle$ = 0.1

Dimming simulated with the calculated maximum PWM frequency is presented in Table 9, 10 and 11.

Table 9. Dimming for 350 mA with calculated maximum PWM frequency.

$f_{max\_PWM\_350mA}$ [Hz]	Duty cycle	Average current [mA]	Percentages of max current
714	50 %	174	49.7 %
714	25 %	86	24,6 %
714	10 %	34	9,7 %

Table 10. Dimming for 700 mA with calculated maximum PWM frequency.

$f_{max\_PWM\_700mA}$ [Hz]	Duty cycle	Average current [mA]	Percentages of max current
454	50 %	348	49.7 %
454	25 %	173	24.7 %
454	10 %	68	9.7 %

Table 11. Dimming for 1050 mA with calculated maximum PWM frequency.

$f_{max\_PWM\_1050mA}$ [Hz]	Duty cycle	Average current [mA]	Percentages of max current
384	50 %	521	49.6 %
384	25 %	258	24.6 %
384	10 %	101	9.6 %

12.6 Efficiency and ripple for different loads in the driver

Regardless of the load's size on the output, the average currents through the load were unchanged, 350mA, 700 mA or 1050mA. The results can be seen in Table 12, 13 and 14. In each simulation, the same interval, 1.9 – 2 ms, has been analyzed and the highest point, I_{max}, and the lowest, I_{min}, in this interval was noted. The efficiency was based on an average measure value during the interval of 1 – 2ms by using (6.2.a).

$$Efficiency = \frac{P_{out}}{P_{in}} \qquad (6.2.a)$$

$$P_{in} = V_{DC} \cdot I_{DC} + V_{AUX} \cdot I_{AUX} \qquad (6.2.b)$$

$$P_{out} = I_{Average} \cdot V_{load} \tag{6.2.c}$$

In the measurements I_{AUX} was insignificantly small, 1 mA for all measurements, and with $V_{AUX} = 11.6$ V, $V_{AUX} \cdot I_{AUX}$ is neglectable in the efficiency calculations.

Table 12. Measurements and efficiency for 350 mA with different loads.

Load	I_{max} [mA]	I_{min} [mA]	I_{Δ} [mA]	I_{DC} [mA]	V_{Load} [V]	Efficiency
0 %	375	323	52	–	–	–
25 %	386	313	73	91	19.2	66.8 %
50 %	404	293	111	154	33.4	68.7 %
75 %	409	287	122	215	57.5	84.9 %
100 %	399	295	104	275	76.7	88.6 %

Table 13. Measurements and efficiency for 700 mA with different loads.

Load	I_{max} [mA]	I_{min} [mA]	I_{Δ} [mA]	I_{DC} [mA]	V_{Load} [V]	Efficiency
0%	724	679	45	–	–	–
25 %	740	659	81	174	20.2	73.6 %
50 %	758	641	117	310	40.4	82.6 %
75 %	760	638	122	428	60.5	89.7 %
100 %	746	651	95	555	80.7	92.3 %

Table 14. Measurements and efficiency for 1050 mA with different loads.

Load	I_{max} [mA]	I_{min} [mA]	I_{Δ} [mA]	I_{DC} [mA]	V_{Load} [V]	Efficiency
0 %	1078	1025	53	–	–	–
25 %	1091	1002	89	261	20.9	76.3 %
50 %	1105	985	120	458	41.7	86.9 %
75 %	1110	990	120	659	62.5	91.0 %
100 %	1090	1010	80	855	83.4	93.1 %

The output current had an irregularly behavior without load, this can be seen in Fig. 39 for 700mA without load and be compared to Fig. 40 for 700 mA with 75 % load.

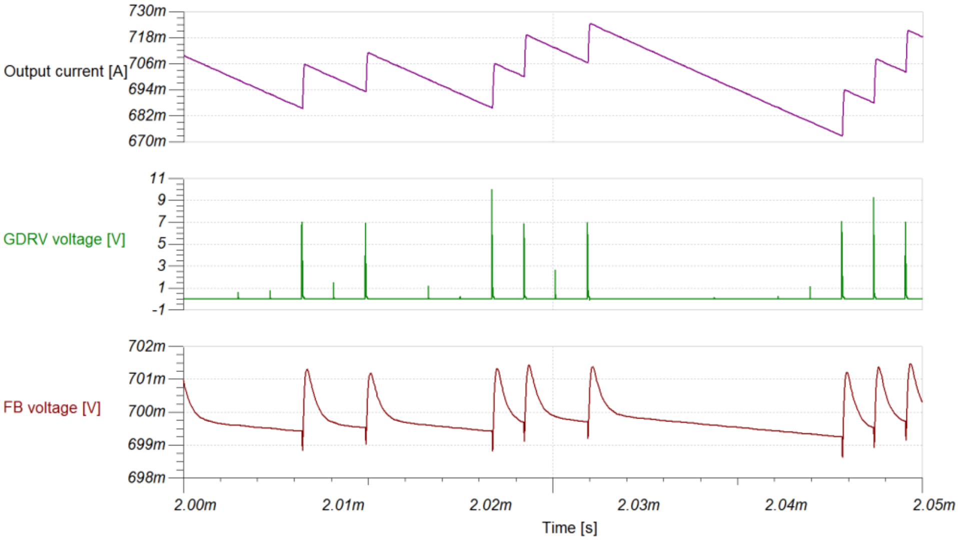


Fig. 39. The output current of 700 mA, the voltage on the GDRV pin and the voltage at the FB pin without any load.

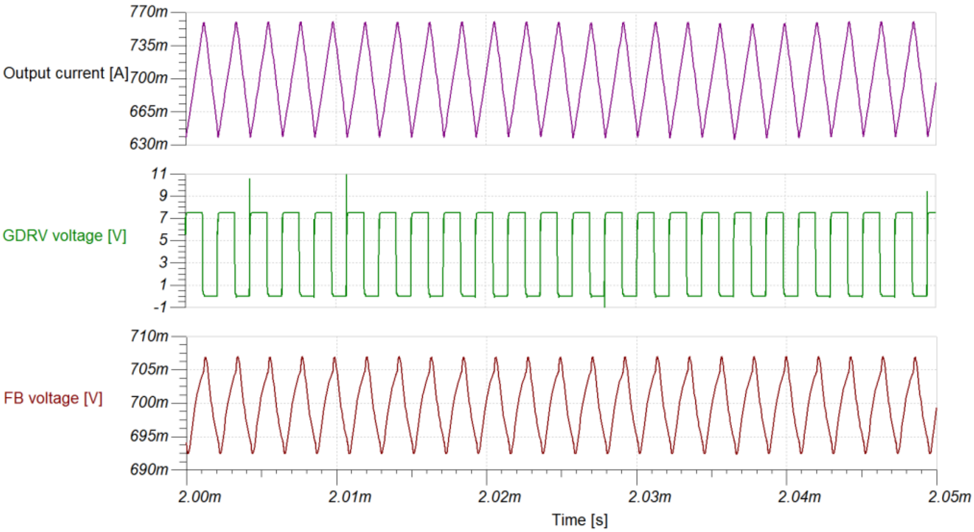


Fig. 40. The output current of 700 mA, the voltage on the GDRV pin and the voltage at the FB pin with 75 % load.

12.7 Power losses in the driver

The crossover losses in the MOSFET, Q_9, was calculated in (6.3.a) with values from Fig. 41.

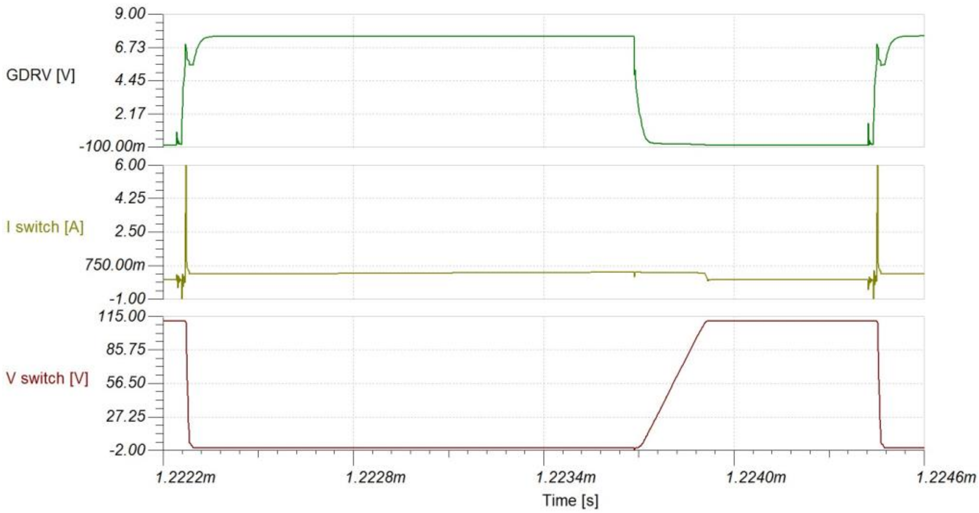


Fig. 41. The MOSFET's on and off stage can be seen in GDRV, The current through the MOSFET in I switch and the voltage over the MOSFET in V switch. This figure is for 700 mA.

$$P_{crossover\_losses} = \left(P_{rise} \cdot t_{rise} + P_{fall} \cdot t_{fall}\right) \cdot f_{SW} \quad (6.3.a)$$

$$P_{crossover\_losses} = 2.34\ W$$

$$P_{rise} = V_{rise\_average} \cdot I_{rise\_average} \quad (6.3.b)$$

$$P_{fall} = V_{fall\_average} \cdot I_{fall\_average} \quad (6.3.c)$$

Where:

$V_{rise\_average} = 53\ V$

$V_{fall\_average} = 53\ V$

$I_{rise\_average} = 350\ mA, 700\ mA\ or\ 1050\ mA$

$I_{fall\_average} = 350\ mA, 700\ mA\ or\ 1050\ mA$

$t_{rise} = 115\ ns$

t_{fall}=20 ns

$f_{SW} = 465\ kHz$

The turn-on losses were calculated as follow:

$$P_{turn-on\_losses} = \frac{c_p \cdot V^2 \cdot f_{SW}}{2} \quad (6.3.d)$$

$$P_{turn-on\_losses} = 0.37\ W$$

Where:
$c_p = 129\ pF$
$V = 110.3\ V$
$f_{SW} = 465\ kHz$

The diode losses at normal operation condition was calculated as follow

$$P_{diode\_losses} = \frac{1}{4} Q_{rr} \cdot V_{Drr} \cdot f_{SW} \quad (6.3.e)$$

$$P_{diode\_losses} = 0.45\ W$$

Where:
$Q_{rr} = 35\ nF$ [25]
$V_{Drr} = 110.3V$
$f_{SW} = 465\ kHz$

12.8 Lamp comparison

Efficiency difference between an undimmed lamp connected to 350 mA and a 50 % dimmed lamp connected to 700 mA, Table 15.

Table 15. Efficiency comparison between 350 mA and 700 mA dimmed load.

Load	**Efficiency for 350 mA**	**Efficiency for 700 mA, 50% dim**	$I_{average}$ **[mA] for 700 mA, 50 % dim**
100 %	88.6 %	92.0 %	349

12.9 How varying input voltages affects the load

The results for the three currents and how the load is affected in the simulations with varying voltage supply for the driver can be seen in Table 16, 17 and 18. The estimated value from the power supply is V_{DC} at 110.3 V and V_{AUX} at 11.6 V.

Table 16. The current and voltage for the load and efficiency with varying supply voltage for 350 mA.

V_{DC} [V]	V_{AUX} [V]	I_{Load} [mA]	V_{Load} [V]	Efficiency
90	9.3	358	76.8	91.2 %
100	10.4	354	76.8	90.0 %
110.3	11.6	350	76.7	88.6 %
120	12.6	348	76.7	86.5 %

Table 17. The current and voltage for the load and efficiency with varying supply voltage for 700 mA.

V_{DC} [V]	V_{AUX} [V]	I_{Load} [mA]	V_{Load} [V]	Efficiency
90	9.3	707	80.8	93.6 %
100	10.4	704	80.7	93.0 %
110.3	11.6	700	80.7	92.3 %
120	12.6	697	80.7	90.8 %

Table 18. The current and voltage for the load and efficiency with varying supply voltage for 1050 mA.

V_{DC} [V]	V_{AUX} [V]	I_{Load} [mA]	V_{Load} [V]	Efficiency
90	9.3	1058	83.5	94.3 %
100	10.4	1054	83.4	93.6 %
110.3	11.6	1050	83.4	93.1 %
120	12.6	1047	83.4	92.6 %

12.10 Heat convention in the LED lamp

Without having the PCB mounted on the heat sink, the temperature in the small LEDs rose fast, up to 129 °C, Fig. 42, and after just a minute, it started to smell burnt. The upper left value shows the temperature for the middle point in the picture.

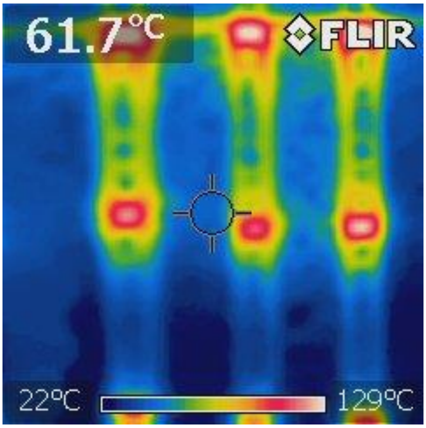


Fig. 42. Picture from the thermal imager with PCB just lying on the heat sink.

With the PCB mounted on the heat sink, the highest temperature on the PCB was at the beginning 100 °C, Fig. 43, and after operating during 20 minutes the temperature had made a small increase to 105 °C, Fig. 44.

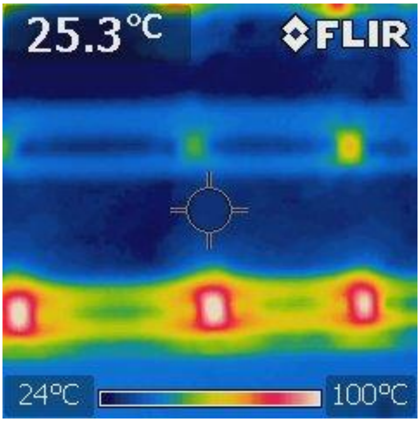


Fig. 43. Picture from the thermal imager with PCB mounted on the heat sink just after the LED lamp is switched on.

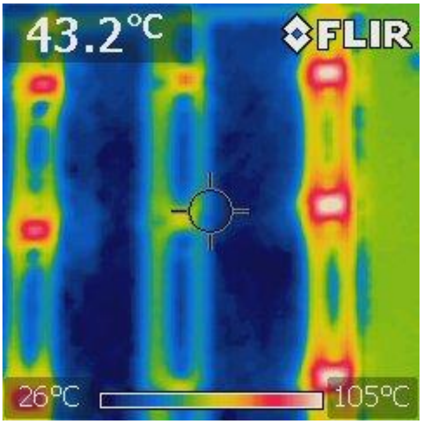


Fig. 44. Picture from the thermal imager with PCB mounted on the heat sink after 20 minutes on.

12.11 Bread board

There was too much noise in the bread board circuit to get any relevant measurement results. For example, the current through the LED lamp was much lower than expected caused by the exceeded voltage at the ISNS pin that will trig the overcurrent protection if it exceeds 150 mV. This problem was both detected visually by observing flicker and also by studying the voltage at the SS pin. Fig. 45 shows when the light is flickering. The ISNS voltage is the blue signal, 200 mV/Dev, and is high when the LED lamp is on and the soft start is the red signal, 2 V/Dev, shows how the capacitance connected to the SS pin is discharging as a restart timer trying to start up again.

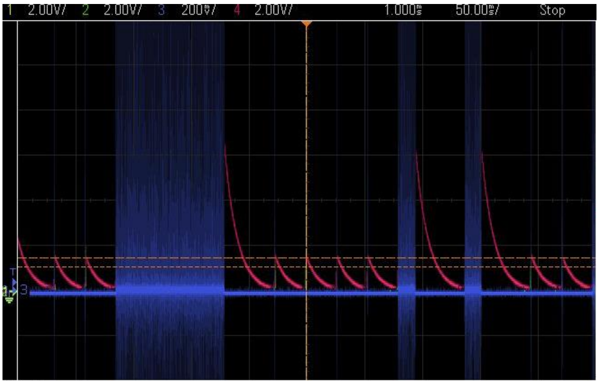


Fig. 45. Flicker light from the LED. The blue, channel 3, 200 mV/Div, is ISNS that trigs the overcurrent protection at 150 mV, the red, channel 4, 2 V/Div, is soft start's state of charge.

12.12PCB board layout

The final PCB board layout in EAGLE with top layer is shown in Fig. 46 and bottom layer in Fig. 47. The PCB was designed with a solid ground layer and small grids just around the vias and pads.

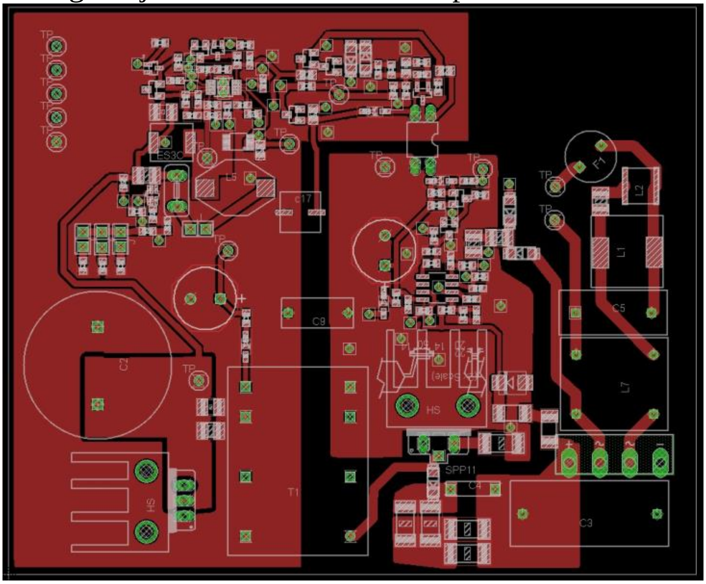


Fig. 46. Top layer of the final PCB in EAGLE.

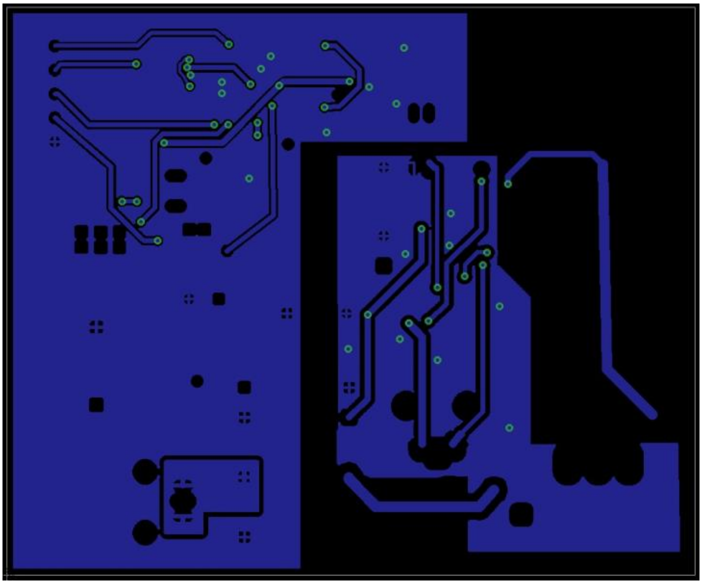


Fig. 47. Bottom layer of the final PCB in EAGLE.

12.13 Testing the PCB

An unwanted behavior appeared early during the tests, the circuit seemed to have problems with the switching power supply. It was switching in slow intervals and the 40 W lamp was turned on during half a second and then turned off for few seconds with repeatedly behavior. When measurements were made on V_{DD}, Fig. 48, it was clear that the supply voltage for the IC was falling below the lower threshold of VCC pin at 9.7 V and thereby turned off the IC. After it was turned off, V_{DD} started to rise up to VCC pin's upper threshold voltage, 15.8 V. When it was reached the IC was turned on and V_{DD} started to decrease again. The measurements can be seen in Fig 49. V_{DD} is represented as channel 1, channel 2 shows the voltage at the DRV pin and channel 4 is the voltage at the COMP pin. The supply voltage for the IC should however be held at 18 V when the switching is activated due the 18 V zener diode, D_{100}, and voltage supply from the second winding on the transformers primary side when MOSFET, Q_2, is saturated in the switching on state. The problem was resistor R_{12} that constantly broke which resulted in a voltage drop at V_{DD}. To avoid this problem, R_{12} was replaced with another resistor of a higher value, 47 Ω. After the change, V_{DD} could be held at a steady level of 18 V and the switching worked as intended.

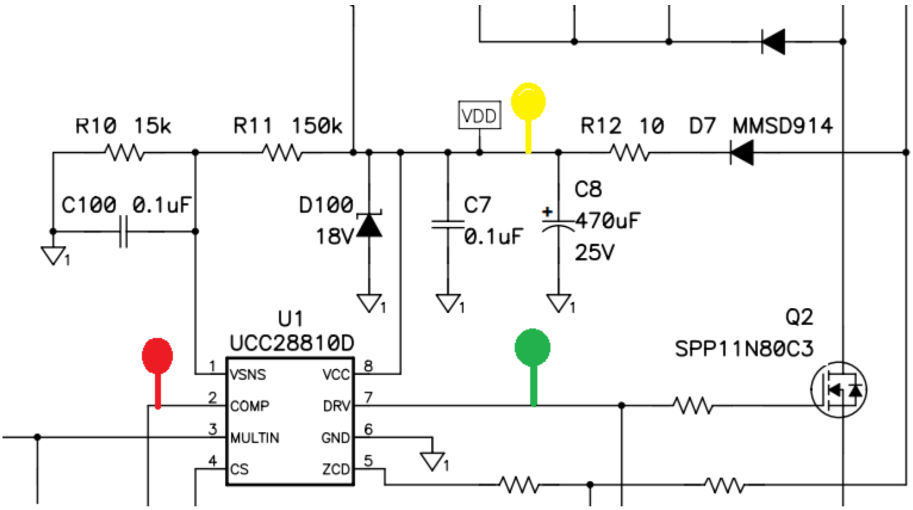


Fig. 48. Test point locations of V_{DD}, DRV and COMP.

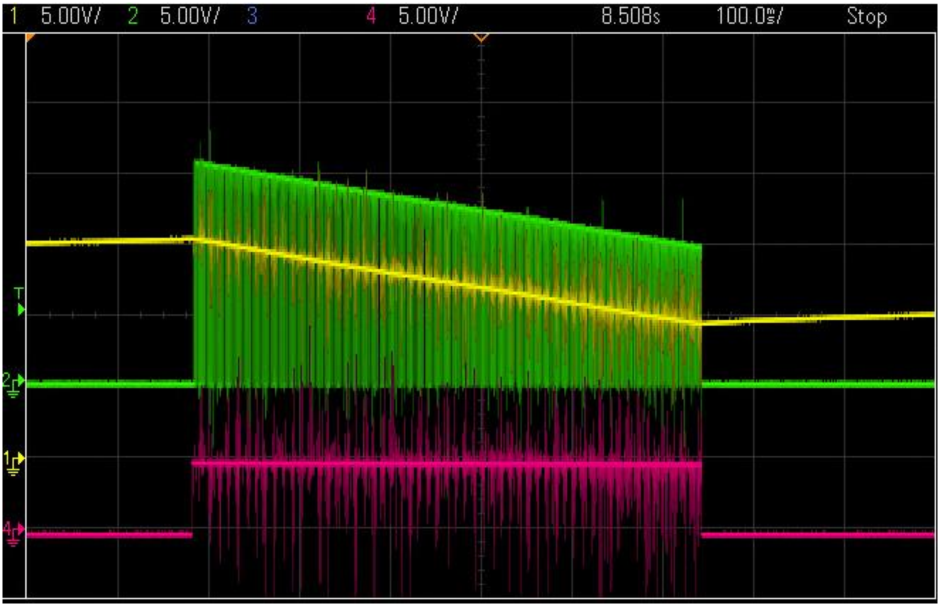


Fig. 49. The voltage drop of V_{DD}. Channel 1 is V_{DD}, channel 2 is the DRV pin and channel 4 is the COMP pin of UCC28810. 5 V/Div and 100 µs/Div.

With the switching working, the voltage at the secondary side of the transformer, V_{DC}, stayed stable and the lamp did no longer had a visible switching between on and off. Unfortunately, V_{DC} was rising way above its estimated value of 110.3 V, up to 240 V. The voltage feedback from the optoisolator that was supposed to regulate V_{DC} was not working. The problem was caused by the non-existing voltage at V_{AUX} which led to malfunction of the feedback. No current did flow through the diode in the optoisolator and there would never be any conduction on the other side giving feedback to UCC28810. When no current is flowing through the optoisolator, the voltage at the COMP pin will remain high and the switching continued. Even when an external power supply was connected to V_{AUX} to keep it high, the feedback still did not work properly. The voltage at the COMP pin was not always falling below 2.3 V which is required to stop the switching. V_{DC} fell admittedly down to 130 - 140 V with a power supply replacing the V_{AUX} but was still higher then intended.

When new measurements were performed on the optoisolator, with the driver disable and without an applied V_{AUX}, V_{DC} rose to a point where it exceeded the reversed voltage limit of the diode, D_{15}, and broke it. This also broke the MOSFET, Q_9, and a short circuit between V_{DC} and ground was a fact, Fig. 50. Several components broke and some traces on the PCB got destroyed. After this incident unfortunately no further measurements could be done.

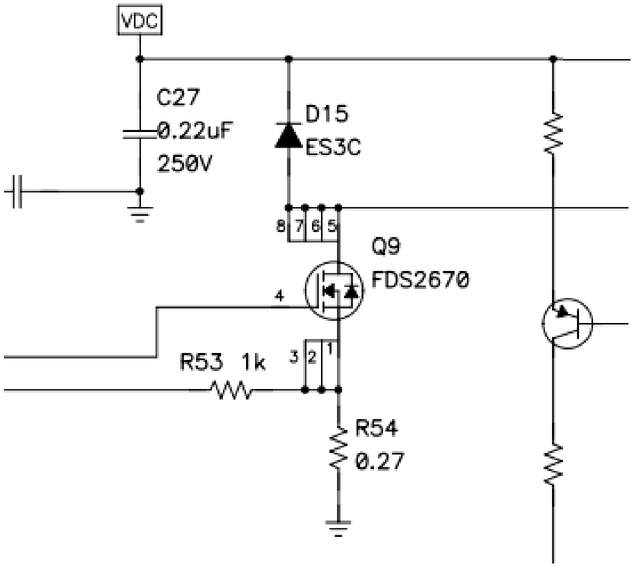


Fig. 50. Shortcut was created through D_{15}, and Q_9 because of a too high V_{DC}.

13 Discussion

13.1 The external controls

Since a lack of documentation for the external controls, only the two operative controls, ENABLE and DIM were included in the simulations. The TRIM and ISNS function in the reference circuit probably work with ISNS as an input and TRIM as an output. This can be an explanation to the reference circuit's choice of resistor R_{62} at 1 kΩ for a reference current of 350 mA compared to our choice of R_{62} at 1.07 kΩ. Instead of mathematically insert the exact resistor value, TRIM is polishing the feedback for the desired current.

13.2 The load resistance

The calculated value for R_{62} is a good reference value and calculated with the most depending components in the circuit. But other components will affect as well and therefore the simulated value for R_{62} was chosen as final value to work with. When ordering components, it was of great significance to make sure the tolerance was kept low, at least 1 %, to make sure the value of the resistor was as accurate as possible to avoid wrong output current.

13.3 Overcurrent protection

By selecting a common R_{54}, at 100 mΩ suited for 1050 mA, for the three different load currents, the testing became simplified by not needing extra components that had to be switched for each current. The overcurrent protection will be trigged when the peak value of the current through the LED lamp exceeds 1500 mA. Since the same lamp was used for all three currents during the tests there were no risks of damaging the lamp even when using an output of 350 or 700 mA.

Normally the overcurrent protection does not trig during simulations since it is more or less an ideal model. This is a situation that will only occur if something in the circuit breaks or if there are lots of interferences in the circuit as there were all the time in the bread board prototype. During simulations the overcurrent protection was forced to trig, just so it could be studied.

13.4 The frequency for the circuit

The difference between the calculated switching frequency of the driver, 433 kHz, and the simulated frequency, 465 kHz, depends on the input voltage for the TPS40210, V_{AUX}. If V_{AUX} instead was changed from its value of 11.6 V and raised to the maximum input value for the IC, 52 V, it would result in a frequency of 402 kHz. This makes it hard to calculate an exact switching frequency and thereby we recommend that the calculated frequency always is simulated and tested before any construction is made to make sure it is not too far from what is desired.

It was expected that a higher switching frequency would give a smaller ripple since faster switching results in shorter pulses where the current increases, Table 4. The table also shows that a higher frequency results in a lower efficiency. The input current, I_{DC}, increases at the same time as the output remains the same, which means the losses increases with higher switching frequencies. This is something that has to be considered wisely, for each application. It might be essential with a high efficiency, but for what cost? Is it reasonable to get three times higher current ripple to achieve 3.8 % higher efficiency? Or shall the ripple be cut in half but instead lose 5.6 % efficiency? Fig. 37 shows that our simulated frequency of 465 kHz has a quite good balance of efficiency and current ripple. One important criteria that needs to be fulfilled is that the ripple can not be as high as it will trig the overcurrent protection. Due to the fact that the lamp we used in this thesis work was known, the trigging point for the overcurrent protection was common for all three currents and set to 1500 mA. To be able to apply different load to the circuit the resistor R_{54} must be dimensioned according to (4.3.a) for trig the overcurrent protection before breaking the lamp.

Our recommendation is that if it is not necessary to achieve a higher efficiency or lower current ripple, then the switching frequency should not be changed from its original value of 465 kHz. But if it is changed, then the user has to make sure that the current protection is properly dimensioned.

13.5 Dimming and PWM frequency

The simulations of the dimming, with 1 – 5 kHz, shows that the output current follows the duty cycle of the PWM but with a small difference in the output. The longest rise time was related to the highest current, which was expected. The effect of the rise time can clearly be seen in the dimming with small duty cycles, Table 5, 6, 7. The longer the duty cycle is, the more accurate is the dimming and the difference seems to flatten out with lower frequency. This is because the rise time represents a shorter part of the total pulse time. The dimming was simulated with a very high frequency, compared to what is recommended for not be disturbing for the eyes.

With the calculated maximum PWM frequencies, all were way below the circuit’s simulated switching frequency of 465 kHz. The PWM frequency can take any value between 200 Hz and the maximum frequency for the three currents, 714 Hz for 350 mA, 454 Hz for 700 mA and 384 Hz for 1050 mA. But the lower the PWM frequency is, the better the dimming will become. When simulations were made with the calculated PWM frequency, the output current behaved as we wanted and only had a small difference of 0.3 – 0.4 % compared to the duty cycle. The rise time can never be zero and will always have a small impact on the output current, but such a small difference as 0.4 % could be tolerated. We can not find any reason to use a PWM frequency unnecessary high and if the user want to be sure to get an accurate dimming and at the same time avoid flicker, then 200 Hz or slightly higher works just fine.

13.6 Efficiency, ripple and voltage variety

The ripple of the output current was only uneven when no load was connected. Otherwise the ripple has no significant difference between the three output currents at the same frequency. The results shows that even if we use a higher current than TI in the original driver, the ripple stays at the same level, comparing Fig. 20 from TI’s test results with a ripple of 100 mA and Table 12, 13 and 14 where the ripple is almost the same for the three currents at the same load. For 100 % load was the ripple 104 mA for 350 mA, 95 mA for 750 mA and 80 mA for 1050 mA. From this we concluded that no extra filter had to be added on the output to reduce current ripple.

Comparing Fig. 39 and Fig. 40 the current ripple on the output without any load has a different look compared to the one with a load. There are several factors that affect this irregular ripple. The output of GDRV has small voltage spikes up to 10 V. The MOSFET only switches if the pulse from the GDRV pin is above 5 V. But the fact that there are small regular pulses, even if they do not trig the transistor, shows that the switching frequency remains the same for the circuit but the switching is limited. The limitation is cause by the fast rising current on the output relative to its slow decreasing when the switch is off, which can be observed in Fig. 39 at 2.025 ms. The FB pin can sense the high current and cuts the voltage at GDRV before it has reached 5 V, which prevents the MOSFET from conduct until the current has reached a more stable level, Fig. 39, between 2.030 – 2.045 ms.

13.7 Lamp comparison

A lamp for 700 mA with 50 % dimming is, based on the simulations, a better choice and gives a higher efficiency than an undimmed lamp for 350 mA, Table 15. This would prove that it would be better idea to set the driver to 700 mA and dim it by 50 % than set it to 350 mA. But in reality, other factors, such as the payback time of the investments in bigger LED lamps, needs to be taken into consideration as well. If an LED lamp constructed for 700 mA cost more than one for 350 mA, maybe it might be a good idea to use the one at 350 mA, even if it has a lower efficiency. As an example; Both cases have 100 % load, an input voltage, I_{DC}, at 275 mA and an input voltage, V_{DC}, at 110.3 V. An estimated value for the street light to be lit is 10 h/day 365 days/year and results in a consumption of 110.7 kWh/year. The lamp for 350 mA with 88.6 % efficiency has 12.6 kWh in losses each year and a 700 mA lamp with 50 % dimming has annual losses of 8.9 kWh. With a price of 1 sek/kWh, it would result in a profit of 3.7 sek/year using a 700 mA lamp and dim it instead of a 350 mA. An LED has an estimated lifetime of 50 000 hours which is 14 years when lit 10 h/day. The profit during the lifetime is approximately 52 sek.

Another thing that would be affected by using 700 mA at 50 % is the dimming. The lowest dimming level would require a 5 % duty cycle of the 700 mA to achieve the same amount of light as a 10 % dimming of a 350 mA. This would affect the maximum PWM frequency by a factor 0.5 down to 227 Hz. It is still over the minimum of 200 Hz but not far away. So it still might be a better idea to use the proper settings at 350 mA instead of 50 % dimming at 700 mA.

13.8 Voltage variations

The calculated voltage from the power supply is 110.3 V and a difference in V_{DC} between 90 – 120 V is a range that should cover a variety in the driver's supply voltage. In this range, the current through the load varies with 10 mA, Table 16, 17 and 18, while the voltage over the load is unchanged and shows that the driver can handle the voltage variety without any major problem. It also shows that the efficiency got a bit higher with a lower V_{DC}. This was expected and as mentioned in 6.3 *Power losses in the driver*, the losses are directly connected to the input voltage, a higher voltage will result in more losses.

This can be related to the tests with different loads, Table 12, 13 and 14, which showed that the lowest efficiency was related to the smallest load. The relation between V_{DC} and the load voltage has a clear impact on the efficiency. A smaller difference between these two results in better efficiency. This can be used to improve the efficiency of the driver by adjusting the input voltage from the power supply. But keep in mind that V_{DC} has to be slightly higher than the intended load voltage to avoid issues if the voltage fluctuates. The test result from TI showed that the voltage might fluctuate in the range of 10 V and we recommend that an additional 10 V is used as security. For example: if the load is intended to be 80 V, than the lowest acceptable V_{DC} shall be 100 V. This should cover unpredictable voltage ripple from the power supply. LEDs are sensitive to voltage drops and will be completely turned off if the voltage level is lower than the threshold.

13.9 Losses

As expected the MOSFET, Q_9, had the highest power losses in the driver with 2.34 W for the crossover and 0.37 W for the turn-on losses. The losses over the diode, D_{15}, were estimated to 0.45 W and combined they give a loss of 3.16 W. For 700 mA with 100 % load with the output voltage of 80.7 V, Table 13, the was efficiency 92.3 %. This is a total power loss of 4.7 W. The calculated losses from the MOSFET and the diode stand for 67 % of the losses in this example. Other losses, such as heat losses in the 1 Ω resistor, R_{48}, and the 0.27 Ω, R_{54}, also has an impact on the efficiency. But these are not affected by changes in voltage or frequency.

The calculated efficiencies for the driver are the theoretical maximum values of the different scenarios but depending of the quality for the lamp, in reality the efficiency will decrease. These calculations are based on the output current and voltage and not on the actual emitted light and heat losses of the LEDs.

13.10 The LED lamp design

Instead of building an LED lamp, a big zener diode could have been used as a load, but for this thesis work the visual feedback has been essential. The ability to immediately see if the circuit worked or not and if the light was flickering or instable was needed. If a zener diode had been used, all measuring would have been much harder to perform and the circuit more difficult to analyze.

In the LED lamp design, resistors were used to distribute the current evenly between the rows. It is an acceptable and temporary solution but not optimal and was chosen because it was a quick and easy way to build a functional lamp to make measurements with. The current was a little bit unevenly distributed with the highest current in the rows closest to the connection point for the input current. For designing a more long lasting and stable LED lamp with several rows of LEDs, one solution could have been to use one separate driver for each row. The disadvantage with this solution is partly a higher cost for all extra components but also the reliability for each driver and the size of the circuit becomes remarkably larger. Another less expensive and less space occupying solution is to add one PNP transistor on each row, connected as a current mirror with common reference base current set by one of the rows. That would result in equal current flow through all transistors, with the same size as the desired current. A problem that could appear would be if the reference row has lower voltage due to differences in forward voltage, than the other rows. The purpose of the current mirror would then be unsuccessful and not operate as it should. This can be prevented by adding one resistor in the reference row to make sure that row’s reference voltage always is a little bit higher than the others. Another problem can be a slight difference in the transistors V_{BE}. To stabilize, a small value resistor can be added after the emitter in each row. [22]

13.11 Cooling of the LED lamp

The cooling of the LED lamp was much better after it had been screwed to the heat sink, instead of just lying on top of it. The concave shape of the PCB led to too much air between it and the heat sink. By reduce the air resistance, with very low conductivity factor, the heat sink could be more useful and the temperature in the LEDs remarkably decreased. The estimated thermal resistance from CREE of 25 °C/W, Fig. 23 page 41, is an optimized value. Our LED lamp could have had a lower thermal resistance by making the isolation distance between the lines and cooling pads smaller, Fig. 51, to minimize the thermal resistance. There will always be a small air gap between the two surfaces but to minimize it, silicon grease can be applied for a better heat conduction. This in combination with a thinner PCB, due to the fact that the FR4 material works more like an isolator with its low conductivity factor, and add thermal vias would result in a very successful LED lamp.

Fig. 51. The vertical LED lines with cooling pads between.

13.12 Bread board

Because of all noises, it was really hard to get any relevant feedback from the circuit, even though we tried to decrease the noises in the driver by adding capacitors as filters and shortened wires. The ICs are very sensitive and affected by just the smallest disturbances on its pins. A lot of time was spent on the IC in the driver to make it work properly but its behavior could not be controlled. The circuits on the bread boards was never exposed to maximum voltage due to security reasons, and the fact that it did not work as expected, we did not want to jeopardies and break the components. Another aspect of the miss function might be that the bread board itself was not made for handling high voltage and we can not make sure that no sparks between components occurred. Coupling on a bread board was not a great success and was a time consuming moment in this work compared to the received feedback.

13.13PCB and layers

The reference design was consisting of one power supply and three drivers. In the new design there is only one driver which means much less components and connections on the board. Due to the fact that a two layer PCB is cheaper than a four layer, the two layers are preferred from that point of view. There is one advantage of using four layers instead of two layers, one of the layers can be completely dedicated to be a ground plane. Even with this in consideration, in the end a two layer PCB was chosen to keep the cost as low as possible.

There are some 90° angles traces on the routing in our design. Those angles were chosen due to the simplification of the design and there is only low current and voltage going through those traces, which means there will not be any high temperature in those curves.

13.14Test on the PCB

When the PCB was tested, it soon became clear that those three extra implemented test points to the design was not satisfying enough. Measurements had to be done at locations on the PCB that did not have any test points or suitable components to attach probes to. The testing would have been a lot easier if more test points had been implemented before the PCB was manufactured.

The conclusions made out of the measurements, before it broke, were that the feedback did not work properly and V_{DC} got out of hands. The circuit's inability to keep V_{AUX} high may be explained to a broken diode connected between V_{AUX} and the transformer. The output of the transformer at pin 7 seemed to be correct but V_{AUX} was still low.

As the testing proceeded, more and more components broke. A lot of time was spent on searching for failure, broken components and short circuits. It would have been a good idea to re-solder the whole circuit with new components instead of kept going with the trouble shooting. But a lack of extra components and economy prevented us from soldering a new PCB.

14 Further work

The simulations were only made with zener diodes as load for the driver, but for more accurate results it would be a good idea to build a realistic lamp with the right LEDs.

In the final design, it is recommended to use three different resistors for the overcurrent protection if different lamps are used for each current, and switch between them for the desired output current to protect it. This can be done by using simple jumpers as we did to control the output current.

If the user's purpose is to use a lower load, the input voltage to the driver, V_{DC}, can advantageous be decreased to receive a better efficiency. A new V_{DC} can be calculated by (3.1.8.a) and change the ratio between the resistors R_9, R_{23}, and R_{28}. For an exact average current through the load, the simulations for resistor R_{62} must be remade after the calculated value with the new input voltages to obtain the exact resistor value.

For easier testing of the PCB during troubleshooting we suggest more test points to be implemented in the circuit design. A suggestion of these test points can be seen in Fig. 52. Test point 1 will be used to measure the voltage after the rectifier, number 2 for V_{DD}, number 3 to measure the feedback from the optoisolator to COMP, number 4 to see if there is any current flowing through the optoisolator and number 5 to measure the voltage over R_{28} which is also the voltage to the shunt regulator, U_3.

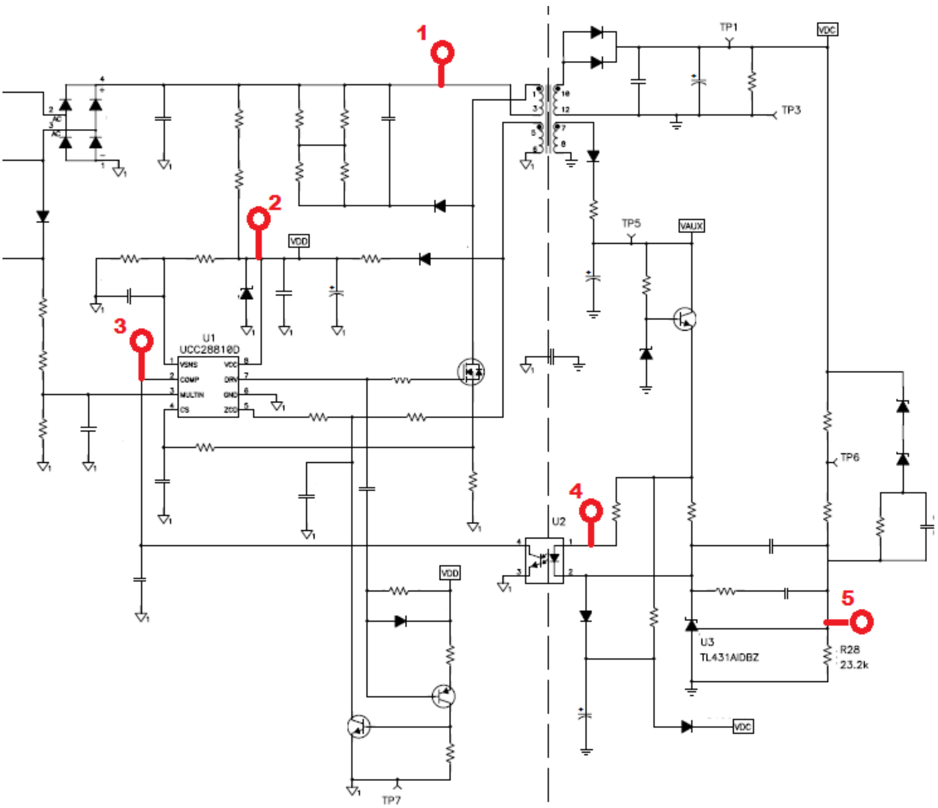


Fig. 52. Suggestions of implementations of additional test points.

For further work with this circuit, it is recommended to take the environment where it is going to be used into consideration, like moisture and pollution. How well isolated it will be and things like that. In this thesis work the purpose was to make a prototype on a low budget and it has not been prioritized to get components with a high estimated lifetime, this is also something that might be important. When manufacturing a driver like this for commercial use, it is important to make sure the right standards and regulations for trace width and clearance are followed.

Another substantial thing to do before taking a prototype into production is to test it for Electromagnetic Compatibility (EMC). This means the product shall be functional in an environment together with other electrical equipment. The Electromagnetic Interference (EMI) from the product itself has to be limited to a level where it does not disturb the equipment around it, but also make sure it works as intended with EMI from the surrounding environment. It is quite easy to notify if a product fails the test of EMI to its surrounding, but much harder to make sure it pass the test. Usually these EMC tests are outsourced and done by professional test houses.

If the efficiency shall be improved, without changing parameters such as input voltage or switching frequency of the driver, it might be a good idea to search for better components. A switching transistor with a shorter rise time and lower parasitic capacitance for example would not be affected as much as the ones used in our design.

15 Investigation areas

A summarize of the investigation areas, 1.4 *Investigation areas of the different currents*, is presented below:

- **Appropriate PWM interval for the driver**
  To get an accurate dimming, the PWM frequencies are suggested to be kept within these intervals:
  350 mA: 200 – 714 Hz
  700 mA: 200 – 454 Hz
  1050 mA: 200 – 384 Hz

- **Variety in the load**
  A variation of the load, between 0 – 100 %, shows that higher load correspond to a higher efficiency and vice versa. The current ripple has a small change with the highest value at 50 and 75 %.

- **Variety in the supply voltage**
  A variation of the input voltages to the driver between 90 – 120 V makes no significant difference of the current ripple but shows that the efficiency rises with a lower input voltage.

- **Study the efficiency**
  The efficiency of the driver is affected by two parameters at full load, the input voltage from the power supply and switching frequency. The efficiency decreases with both higher frequency and input voltage.

- **Study the ripple**
  The output ripple of the current is more or less only affected by a change in the switching frequency. A small difference can however be seen when the load are changed, but this small variation will not affect the circuit. A higher switching frequency gives a lower current ripple. But keep in mind that higher frequency also impairs the efficiency.

16 Conclusion

This thesis work has given us great experience by examine and understanding datasheet and circuit schematics. The influence different parts have on each other and small changes can cause huge differences on the outcome. Designing a PCB layout was a whole new experience for us, including soldering SMD. The restrictions of trace width, clearance, routing and the fact that we have never worked with such small components before have been giving us a new perspective of electronics. Another lesson is to always have extra components when building a circuit for testing due to the fact that one mistake easily leads to another and components might need to be replaced.

Even though we did not get a successful result from our PCB, the simulation showed us that this circuit works in theory and should work in reality.

References

[1] http:// www.greinon.se/ (30–10–2014)

[2] http://www.ti.com/tool/PMP4862.1 (03-11-2014)

[3] Texas Instruments, *Getting Started with TINA–TI*, Texas Instruments, August 2007

[4] http://www.cadsoftusa.com/ (12–09–2014)

[5] http://www.grahlighting.eu/en/learning–centre/street–lighting–technology–comparison (25–04–2014)

[6] http://bsl.lacity.org/led–news–media.html#prettyPhoto (25–04–2014)

[7] Texas Instruments, *PMP4862 Schematics*, August 2011
http://www.ti.com/lit/df/slur512/slur512.pdf (22–10–2014)

[8] Texas Instruments, *UCC28810/UCC28811 - LED Lightning Power Controller*, October 2008
http://www.ti.com/lit/ds/symlink/ucc28810.pdf

[9] NXP Semiconductors, *TL431 family*, Rev. 4, June 2011
http://www.nxp.com/documents/data_sheet/TL431_FAM.pdf

[10] Texas Instruments, *TPS40210/TPS40211 - 4.5–V to 52–V input current mode boost controller*, Rev D. March 2008 – October 2011
http://www.ti.com/lit/ds/symlink/tps40210.pdf

[11] Texas Instruments, *PMP4862 Test Results*, Rev D, April 2013
http://www.ti.com/lit/pdf/sluu636

[12] Cree, *Cree XLamp MX–6S LEDs*, July 2013
http://www.farnell.com/datasheets/1810598.pdf (07–08–2014)

[13] Mail correspondence with Bertil Larsson 07–08–2014

[14] Cree, *Optimizing PCB Thermal Preformance for Cree XLamp LEDs*, November 2010

[15] Cree, *Eye Safety With LED Components*, 2014

[16] A. Weiler and A. Pakosta, *High–Speed Layout Guidelines,* Texas Instruments, November 2006

[17] http://www.onmyphd.com/?p=pcb.design (15–10–2014)

[18] – *IPC–2221A, Generic Standard on Printed Board Design,* Association Connecting Electronics Industries (IPC), February 1998,

[19] Texas Instruments, *PMP4862 PCB,* August 2011
http://www.ti.com/lit/df/sluu635/sluu635.pdf (22–10–2014)

[20] ON Semiconductors, *MUR1620CT/D,* Rev. 7, September 2008
http://www.onsemi.com/pub_link/Collateral/MUR1620CT–D.PDF

[21] Fairchild Semiconductors, *MMSD914 Small Signal Diode*, February 2005
https://www.fairchildsemi.com/datasheets/MM/MMSD914.pdf

[22] M. Murugesu, *Current Distribution in Parallel LED Strings,* OSRAM Opto Semiconductors, June 2011

[23] D. Cooper, *LED lighting must work with legacy dimming technologies,* LEDs Magazine, June 2011

[24] –D. Graovac, M. Pürschel, A. Keip, *MOSFET Power Losses Calculation Using the DataSheet,* Infineon, V 1.1, July 2006

[25] Fairchild Semiconductors, FDS2670 200V N-channel PowerTrench® MOSFET, Rev C1, August 2001
https://www.fairchildsemi.com/datasheets/FD/FDS2670.pdf

Appendix A

A.1 Bill of material for the reference circuit PMP4862.

COUNT	RefDes	Value	Description	Size	Part Number	Mfr
4	C1, C18, C27, C36	0.22uF	Capacitor, Ceramic, 250-V, X7R, 20%	1210	C3225X7R2E224M	TDK
6	C10, C11, C15, C25, C34, C43	1000pF	Capacitor, Ceramic, 50V, X7R, 10%	0603	C1608X7R1H102K	TDK
2	C12, C13	47pF	Capacitor, Ceramic, 50V, C0G, 5%	0603	C1608C0G1H470J	TDK
1	C14	0.47uF	Capacitor, Ceramic, 50V, X7R, 10%	0603	C1608X7R1H474K	TDK
1	C16	4.7uF	Capacitor, Ceramic, 16V, X7R, 15%	0805	Std	STD
1	C17	68uF	Capacitor, Aluminum Electrolytic, 16V	0.200 * 0.435 inch	25V ZL 68uF 6.3 X 7	Rubycon
12	C19, C23, C24, C26, C28, C32, C33, C35, C37, C41, C42, C44	100pF	Capacitor, Ceramic, 50V, X7R, 10%	0603	C1608X7R1H101K	TDK
1	C2	820uF	Capacitor, Aluminum, 200V	1.380 dia * 1.970 inch	ECOS2DA821EA	Panasonic
3	C21, C30, C39	0.01uF	Capacitor, Ceramic, 50V, X7R, 10%	0603	C1608X7R1H102K	TDK
3	C22, C31, C40	1uF	Capacitor, Ceramic, 16V, X7R, 20%	0603	C1608X7R1C105M	TDK
1	C3	1uF	Capacitor, EMI Suppression, MKP	1.240 x 0.492 inch	B81130-C1105-M	Epcos
1	C4	0.01uF	Capacitor, Polyester Film, 400V, 10%	0.311 x 0.213 inch	ECQ-E4103KB	Panasonic
1	C5	0.22uF	Capacitor, EMI Suppresion, 275V	0.709 x 0.236 inch	B81130-C1224	Epcos
2	C6, C8	470uF	Capacitor, Aluminum Electrolytic, 25V	0.315 inch	25V ZL 470uF 8 X 20	Rubycon
6	C7, C20, C29, C38, C100, C102	0.1uF	Capacitor, Ceramic, 50V, X7R, 10%	0603	C1608X7R1H104K	TDK
1	C9	1000pF Y1	Capacitor, Cer. Disc, 250V, Y1	0.394 X 0.315 inch Max.	ECK-ANA102MB	Panasonic
1	D1	MUR1640CT	Diode, Dual Ultrafast, 400V, 16A	TO-220	MUR1640CT	ON Semi
1	D100	18V	Diode, Zener, 18V, 3W	SMB	1SMB5931BT3	On Semi
2	D101, D102	62V	Diode, Zener,62-V, 1W	SMA	1SMA5944BT3	ON Semi
1	D11	MMBD1501A	Diode, Standard, 200-V, 200-mA	SOT23	MMBD1501A	Fairchild
3	D12, D15, D18	ES3C	Diode, Ultrafast, 150V, 3A	DO-214AB	ES3C	Diodes Inc.
3	D13, D16, D19	5.1V	Diode, Zener, 5.1-V	SOT23	BZX84C5V1LT1	ON Semiconductor
3	D14, D17, D20	BAV99	Diode, Dual Ultra Fast, Series, 200-mA, 70-V	SOT23	BAV99	Fairchild
4	D2, D7, D9, D10	MMSD914	Diode, Switching, 100-V, 200-mA, 225-mW	SOD-123	MMSD914T1	On Semi
1	D3	GBU406	Diode, Bridge Rectifier, 4A, 600 V	0.935 X 0.280 inch	GBU406	Diodes Inc.
2	D4, D5	MURA160T3	Diode, Rectifier, 1A, 600V	SMA	MURA160T3	ON Semiconductor
1	D6	US1K-13	Diode, Rectifier, 1A, 800V	SMA	US1K-13	Diodes Inc
1	D8	6.8V	Diode, Zener, 6.8-V, 225-mW	SOT23	MMBZ5235BLT1	ON Semi
1	F1	5A/250V	Fuse, TR5 Series, 5A, 250V	0.335	3701500041	Wickmann
3	F5, F6, F7	1A	Fuse, Fast-Acting, 125 VAC, 1A	4.00 x 8.5 mm	395 1100	Littelfuse
2	HS1, HS2	{Value}	Heatsink, TO-220/218 veritcal	0.640 x 0.640	634-10ABP	Wakefield
3	J2, J3, J4	PEC36SAAN	Header, Male 2-pin, 100mil spacing, (36-pin strip)	0.100 inch x 2	PEC36SAAN	Sullins
3	J7, J8, J9	PEC36SAAN	Header, Male 5-pin, 100mil spacing, (36-pin strip)	0.100 inch x 5	PEC36SAAN	Sullins
1	L1	470uH	Inductor, SMT, Isat 1.64-A, 786-milliohm	0.484 x 0.484 inch	MSS1278T-474KL	Coilcraft
1	L2	33uH	Inductor, SMT, 1.1A, 310 milliohm	0.250 x 0.250 inch	MSS6132-333ML	Coilcraft
3	L4, L5, L6	470uH	Inductor, SMT, 0.5A, 1.3ohm	0.51x0.37 inch	DO3316P-474	Coilcraft
1	L7	2.5mH	Inductor, Coupled, 1.2A	0.906 x 0.748 inch	UU15.6V-252LF	GCI
2	Q1, Q4	MMBT3904	Bipolar, NPN, 40-V, 200-mA, 225-W	SOT23	MMBT3904LT1	On Semiconductor
1	Q2	SPP11N80C3	MOSFET, N-ch, 800-V, 11-A, 450-milliOhms	TO-220V	SPP11N80C3	Infineon
1	Q3	MMBT2907	Bipolar, PNP, 60V, 600mA	SOT23	MMBT2907LT1	On Semiconductor
3	Q5, Q9, Q13	FDS2670	MOSFET, N-Chan, 200V, 130m-ohm, 3A	SO8	FDS2670	Fairchild
3	Q6, Q10, Q14	MMDT5401	Transistor, NPN, Dual, 150V, 200ma	SOT-363	MMDT5401	Diodes
6	Q7, Q8, Q11, Q12, Q15, Q16	2N7002	MOSFET, N-ch, 60-V, 115-mA, 1.2-Ohms	SOT23	2N7002W	Diodes
1	R1	33	Resistor, Chip, 1/2W, 1% anti-surge	1210	ERJ-P14J330U	Panasonic
1	R10	15k	Resistor, Chip, 1/16W, 1%	0603	Std	Std

COUNT	RefDes	Value	Description	Size	Part Number	Mfr
1	R100	100k	Resistor, 1/4W, 5%	1210	Std	Std
1	R11	150k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
2	R12, R20	10	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R13	4.99k	Resistor, Chip, 1/10W, 1%	0805	Std	Std
2	R14, R15	787k	Resistor, Chip, 1/4W, 5%	1210	Std	Std
1	R17	100	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R18	22.1k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
15	R19, R25, R33, R35, R42, R43, R50, R52, R59, R60, R67, R69, R76, R77, R101	10k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
2	R2, R5	51.1k	Resistor, Chip, 1/2W, 5%	2010	Std	Std
1	R21	1k	Resistor, Chip, 1/10W, 1%	0805	Std	Std
7	R22, R36, R45, R53, R62, R70, R79	1k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R23	1MEG	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R24	0.1	Resistor, Chip, 1-W, 1%	2512	Std	Std
1	R26	3.40k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R27	10k	Resistor, Chip, 1/10W, 1%	0805	Std	Std
1	R28	23.2k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
4	R29, R39, R56, R73	20k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
4	R3, R4, R6, R7	100k	Resistor, Chip, 1-W, 1%	2512	Std	Std
4	R30, R38, R55, R72	4.99k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
3	R31, R48, R65	1	Resistor, 1/4W, 5%	1210	Std	Std
3	R32, R49, R66	499	Resistor, Chip, 1/16W, 1%	0603	Std	Std
3	R34, R51, R68	365k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
3	R37, R54, R71	0.27	Resistor, Chip, 1/10W, 5%	0805	STD	STD
3	R40, R57, R74	499k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
3	R41, R58, R75	49.9k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
6	R46, R47, R63, R64, R80, R81	100k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
5	R8, R16, R44, R61, R78	0	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R9	49.9	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	T1	G095026LF	Transformer, Flyback 125 W	30.5 x 32.00 mm	G095026LF	GCI
5	TP1, TP2, TP4, TP5, TP6	5000	Test Point, Red, Thru Hole Color Keyed	0.100 x 0.100 inch	5000	Keystone
2	TP3, TP7	5001	Test Point, Black, Thru Hole Color Keyed	0.100 x 0.100 inch	5001	Keystone
1	U1	UCC28810D	IC, TM LED Lighting Power Controller	SO8	UCC28810D	Texas Instruments
1	U2	LTV-817M	OPTOISOLATOR 1CH 4-DIP	DIP-4	LTV-817M	LITE-ON Inc.
1	U3	TL431AIDBZ	IC, Precision Adjustable Shunt Regulator	SOT23-3	TL431AIDBZ	TI
3	U4, U5, U6	TPS40210DGQ	IC, 4.5V-52V I/P, Current Mode Boost Controller	DGQ10	TPS40210DGQ	TI

A.2 Bill of material for the new circuit.

Count	RefDes	Value	Description	Size	Part Nr	Mfr
2	C1, C27	0.22uF	Capacitor, Ceramic, 50V, X7R, 10%	1210	MC1210B224K500CT	MULTICOMP
4	C10, C11, C15, C34	1000pF	Capacitor, Ceramic, 50V, X7R, 10%	0603	MC0603B102K500CT	MULTICOMP
2	C12, C13	47pF	Capacitor, Ceramic, 50V, C0G, 5%	0603	MC0603N470J500CT	MULTICOMP
1	C14	0.47uF	Capacitor, Ceramic, 50V, X7R, 10%	0603	C1608X5R1H474K080AB	TDK
1	C16	4.7uF	Capacitor, Ceramic, 16V, X7R, 10%	0805	C0805C475K4PACTU	KEMET
1	C17	68uF	Aluminium electrolytic, 16V, 20%	5x11mm	EEUFM1C680	PANASONIC
1	C2	820uF	Capacitor, Aluminum, 200V	35x30mm	ECO-S2DA821EA	PANASONIC
4	C28, C32, C33,C35	100pF	Capacitor, Ceramic, 50V, X7R, 10%	0603	MCSH18B101K500CT	MULTICOMP
1	C3	1uF	Capacitor, EMI Suppression, MKP	26.5x11x20.5mm	B32923C3105M	EPCOS
1	C30	0.01uF	Capacitor, Ceramic, 50V, X7R, 10%	0603	GRM188R71H103KA01J	MURATA
1	C31	1uF	Capacitor, Ceramic, 25V, X5R, 10%	0603	C1608X5R1E105K080AC	TDK
1	C4	0.01uF	Capacitor, Polyester Film, 400V, 10%	2.5x9x5.5mm	B32560J6103K	EPCOS
1	C5	0.22uF	Capacitor, EMI suppresion, 275V	18x7.5x13.5mm	R46KI322045N0K	KEMET
2	C6, C8	470uF	Aluminium electrolytic, 25 V, 20%	10x12mm	UVR1E471MPD1TD	Nichicon
4	C7, C29, C100, C102	0.1uF	50V, X7R, 10%	0603	C1608X7R1H104K	TDK
1	C9	1000pF Y1	Capacitor, Ceramic, 440VAC, Y5U, 20%	7.5x5mm	AY2102M29Y5US63L7	VISHAY
1	D1	MUR1640CT	Fast recovery diode, 16A, 400V	TO-220	MUR1640CT	MULTICOMP
1	D100	18V	Diode, Zener, 18V, 3W	5.5x2.8mm	BZG03C18-TR	VISHAY
2	D101, D102	62V	Diode, Zener, 62V, 1.5W, 5%	SMA	SMAZ5944B-E3/61	VISHAY
1	D11	MMBD1501A	Diode, Standard, 200-V, 200-mA	SOT23	MMBD1401	Fairchild
1	D15	ES3C	Diode, 150V, 3A	SMC	ES3C	Fairchild
1	D16	5.1V	Diode, Zener, 5.1-V	SOT23	BZX84C5V1LT1G	ON Semiconductor
1	D17	BAV99	Diode, Dual Ultra Fast, Series, 200-mA, 70-V	SOT23	BAV99	Fairchild
4	D2, D7, D9, D10	MMSD914	Diode, Switching, 100V, 200mA, 225mW	SOD123	MMSD914T1	ON Semiconductor
1	D3	GBU406	Diode, Bridge Rectifier, 4A, 600 V	19.5x6.5mm	KBL406G	GeneSiC
2	D4, D5	MURA160T3	Diode, Rectifier, 1A, 1000V	SMA	MURA160T3	ON Semiconductor
1	D6	US1K-13	Diode, Rectifier, 1A, 1000V	SMA	MRA4007T3G	ON Semiconductor
1	D8	6.8V	Diode, Zener, 6.8V, 300mW	SOT23	MMBZ5235BLT1G	ON Semiconductor
1	F1	5A/250V	Fuse, 5A, 250V	8.5x8.5mm	0034.6051	Schurter
1	F6	2A	125 VAC, 2A Fast Acting	4x8.5x8 mm	39512000440	Littlefuse
2	HS1, HS2	-	Heatsink, TO-220/218 veritcal	16.26x16.26mm	1710603	MULTICOMP
4	JUMPERS	-		2.54 mm	Std	Std

1	L1	470uH	Inductor, Isat 1.4A, 540mOhm	12.3x12.3x10mm	MSS1210-474KEB	COILCRAFT
1	L2	33uH	Inductor, 1.6A, 250mOhm	6.1x6.1mm	MSS6132-333ML	COILCRAFT
1	L5	470uH	Inductor, 1.27 ohm, 0.5A	12.95x9.4x5.21	DO3316P-474MLB	COILCRAFT
1	L7	2.5mH	Inductou, Coupled, 1.3A	18x15.5mm	SU10VFC-R13025	KEMET
2	Q1, Q4	MMBT3904	Bipolar, NPN, 40V, 200mA, 225W	SOT23	MMBT3904LT1G	ON Semiconductor
1	Q10	MMDT5401	Transistor, PNP, Dual, 150V, 200mA	SOT-363-6	MMDT5401	Diodes Incorporated
2	Q11, Q12	2N7002	N-chan, 60V, 115mA, 1.2 ohm	SOT-323	2N7002W-7-F	DIODES INC.
1	Q2	SPP11N80C3	MOSFET, N-ch, 800V, 11A, 450mOhms	TO-220FP-3	SPA11N80C3	Infineon
1	Q3	MMBT2907	Bipolar, PNP, 40V, 500mA	SOT23	MMBT2907	Fairchild
1	Q9	FDS2670	N-chan, 200V, 130mOhm, 3A	SOIC-8	FDS2670	Fairchild
1	R1	33	Resistor, Chip, 1/2W, 1%	1210	ERJ-P14J330U	Std
1	R10	15k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R100	100k	Resistor, 1/4W, 5%	1210	Std	Std
1	R11	150k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
2	R12, R20	10	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R13	4.99k	Resistor, Chip, 1/10W, 1%	0805	Std	Std
2	R14, R15	787k	Resistor, Chip, 1/4W, 5%	1210	Std	Std
1	R17	100	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R18	22.1k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
7	R19, R25, R50, R52,R59, R60, R101	10k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
2	R2, R5	51.1k	Resistor, Chip, 1/2W, 5%	2010	Std	Std
1	R21	1k	Resistor, Chip, 1/10W, 1%	0805	Std	Std
2	R22, R53	1k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R23	1M	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R24	0.1	Resistor, Chip, 1W, 1%	2512	Std	Std
1	R26	3.4k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
1	R27	10k	Resistor, Chip, 1/10W, 1%	0805	Std	Std
1	R28	23.2k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
2	R29, R56	20k	Resistor, Chip, 1/16W, 1%	0603	Std	Std
4	R3, R4, R6, R7	100k	Resistor, Chip, 1W, 1%	2512	Std	Std
2	R30, R55	4.99k	1/10 W, 1%	0603	Std	Std
1	R48	1	1/2 W, 5%	1210	Std	Std

1	R49	499	1/10 W, 1%	0603	Std	Std
1	R51	365k	1/16 W, 1%	0603	Std	Std
1	R54	0.12	1/4 W, 5 %	0805	Std	Std
1	R57	499k	1/10 W, 1%	0603	Std	Std
1	R58	49.9k	1/10 W, 1%	0603	Std	Std
1	R62a	1.07k	1/10 W, 1%	0603	Std	Std
1	R62b	534	1/10 W, 1%	0603	Std	Std
1	R62c	352	1/10 W, 1%	0603	Std	Std
2	R63, R64	100K	1/10 W, 1%	0603	Std	Std
3	R8, R16, R61	0	1/10 W, 1%	0603	Std	Std
1	R9	49.9	Resistor, Chip, 1/16W, 1%	0603	Std	Std
14	TP	Test points	Copper		TEST-3	MULTICOMP
1	U1	UCC28810D	IC, TM LED Lighting Power Controller	SOIC-8	UCC28810D	Texas Instruments
1	U2	LTV-817M	OPTOISOLATOR 1CH 4-DIP	PDIP-4	LTV-817M	Lite-On
1	U3	TL431AIDBZ	IC, Precision Adjustable Shunt Regulator	SOT23-3	TL431AIDBZ	Texas Instruments
1	U5	TPS40210DGQ	IC, 4.5V-52V I/P, Current Mode Boost Controller	DGQ10	TPS40210DGQ	Texas Instruments
Count	**RefDes**	**Value**	**Description**	**Size**	**Part Nr**	**Mfr**

A.3 Bill of material for the LED lamp.

LED lamp BOM

Count	Value	Description	Size	Part Number	Mfr
40	MX6SWT-H1-0000-000AE8	High efficient LED lamp, White 87.4lm XLamp MX6 LED	5 x 5 mm	MX6SWT-H1-0000-000AE8	Cree, Inc.
10	20	Resistor, Chip, 2 W, 1%	2512	Std	Std

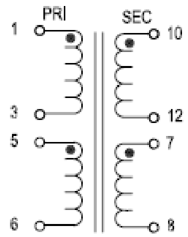
Appendix B

B.1 Original transformer

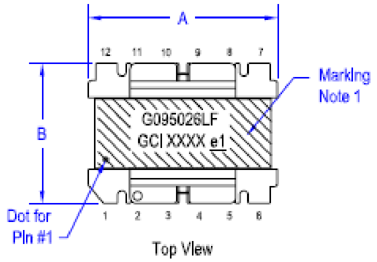
Electrical Specification @ 25°C:

Inductance:	(1-3): 150 μH ±10% @ 100KHz, 0.1V
Leakage Inductance:	(1-3): 5 μH Max @ 100KHz, 0.1V, short (5+6+7+8+10+12)
RDC:	(1-3): 185 mΩ Max (5-6): 125 mΩ Max (7-8): 95 mΩ Max (10-12): 75 mΩ Max
* Turns Ratio:	(1-3):(5-6)=1:0.083 ±2% @ 100KHz, 0.1V (1-3):(7-8)=1:0.055 ±2% (1-3):(10-12)=1:0.50 ±2%
Hipot:	(1-10): 1,500VAC, for 2 seconds, short (1+3) and (5+6+7+8+10+12) (1-Core): 1,500VAC, for 2 seconds, short all windings together

Schematic:

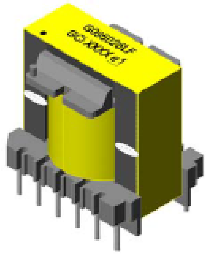


Mechanical Specification:



Top View

Dimensions Table	
Ref	mm
A	31.50 Max
B	23.50 Max
C	31.50 Max
D	3.50±0.50
E	Ø0.80±0.10
F	5.00±0.25
G	17.50±0.50

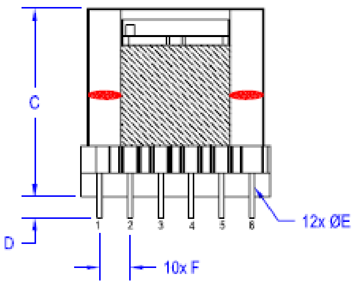


Notes:

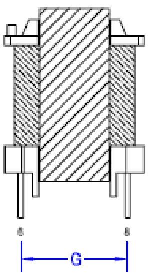
1. Marking shall include:
   GCI Part Number,
   GCI Name, Date Code,
   RoHS Symbol

Marking

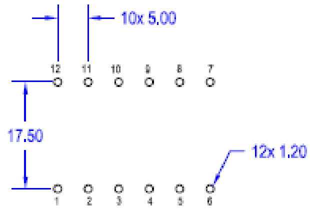
G095026LF GCI XXXX e1



Front View



R-Side View



Suggested PCB Layout

RoHS

"ALL CURRENT CHANGES INDICATED BY ASTERISKS"

Electrical / Mechanical Specification			125 W Flyback Transformer				
DESIGN ENG: Chris V.	APPD. BY: Greg W.		RELEASED BY:	REV: 6	DRAFTER: June W.		DATE: 06/15/09
S/O NUMBER: 095026		GCI PART NO: G095026LF		CUSTOMER PART NO:		SHEET 1 OF 1	

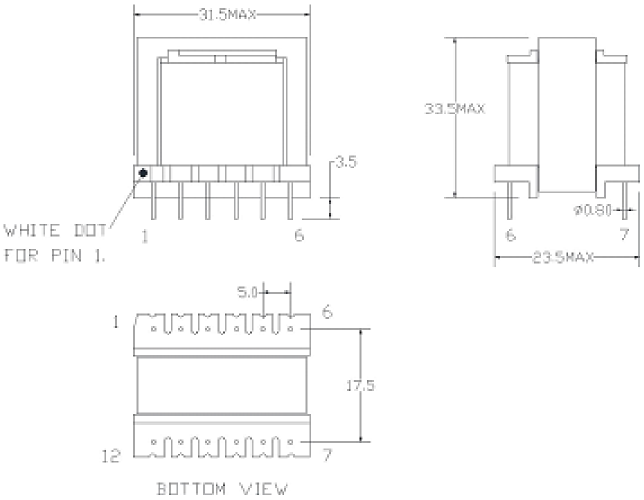
GCi technologies

1301 Precision Drive * Plano, Texas 75074 U.S.A. * Tel: (972) 423-8411 * Fax: (972) 423-4550 * email:gci@gcitechnologies.com * website:www.gcitechnologies.com

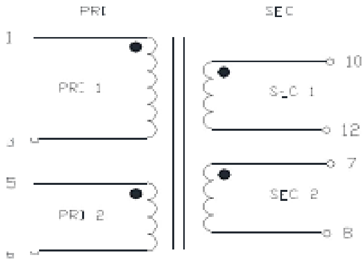
B.2 Transformer from China

惠州市迈特普电子有限公司
MAGTOPELECTRONIC CO., LTD
E-mail: sales@magtop.com.cn Web Site: http://www.magtop.com.cn TEL: +86 752 7803358 FAX: +86 752 5872218

1. DIMENSIONS: (UNIT: mm)



2. SCHEMATICS:



Prepared By:	Deng LiJun	Drawing Title	TRANSFORMER	Part Number	MTEER28-151K	Rev A
Checked By:	Hui Cai					
Approved By:	ZH Tong	Tolerances: .x±0.5, .xx±0.10, .xxx±0.050		Sheet 1 of3		

MTP-QR-014 REV:A/0

惠州市迈特普电子有限公司
MAGTOPELECTRONIC CO., LTD
E-mail: sales@magtop.com.cn Web Site: http://www.magtop.com.cn TEL: +86 752 7803358 FAX: +86 752 5872218

3. ELETCRONICAL CHARACTERISTICS:

1. Inductance (1`3) :	150 uH ± 10% @ 100.0 KHz, 0.10 Vrms
2. Leakage Inductance (3`1) :	5.0 uH MAX @ 100.0 KHz, 0.10 Vrms Tie 5,6,7,8,10,12
3. Turns Ratio :	1~3 : 5~6 : 7~8 : 10~12 = 1 : 0.083 : 0.055 : 0.50 +/-3%
4. DC Resistance :	1~3: 185 mohms MAX 5~6: 125 mohms MAX
	7~8: 95 mohms MAX 10~12: 75 mohms MAX
5. Insulation Strength :	3.0KV AC 5mA 3sec @ Pri to Sec 1.5KV AC 5mA 3sec @ Pri, Sec to Core

4. WINDING STRUCTURE:

WIND	S	F	WIRE	TURNS	TAPE	REMARKS
P1-1	1	3	∮0.40mm*1 2UEW	36	3	CLOSE
S1	10	12	∮0.60mm*1 TEX-E	18	1	CLOSE
S2	7	8	∮0.25mm*1 2UEW	2	3	Evenly
P1-2	1	3	∮0.40mm*1 2UEW	36	2	CLOSE
P2	5	6	∮0.25mm*1 2UEW	3	3	Evenly

Prepared By:	Deng LiJun	Drawing Title	TRANSFORMER	Part Number	MTEER28-151K	Rev A
Checked By:	Hui Cai					
Approved By:	ZH Tong	Tolerances: .x±0.5, .xx±0.10, .xxx±0.050			Sheet 2 of 3	

MTP-QR-014 REV.A/0

E-mail: sales@magtop.com.cn Web Site: http://www.magtop.com.cn TEL: +86 752 7803358 FAX: +86 752 5872218

5. MATERIAL LIST:

No.	DESCRIPTIONS	MATERIALS	MANUFACTURE
1	BOBBIN	EER28 6+6PIN DIP PHENOLIC	JI XIANG TENG
2	CORE	EER28 TP4	TDG
3	WIRE	2UEW 0.25mm	YI DA
		2UEW 0.40mm	
		0.60mm TEX-E	FURUKAWA
4	TAPE	POLYESTER TAPE 17.0mm	CHUN YI
		POLYESTER TAPE 10.0mm	
5	VARNISH	AC-43	CHANG XIAN
6	SOLDER	99.3Sn/0.7Cu	THOUSAND

6. REMARKS:

1. FIXED THE CORES WITH 10.0mm TAPE.
2. IDENTIFY PIN 1 WITH A WHITE DOT ON BOBBIN.
3. DIP THE UNIT INTO THE VARNISH.

Prepared By:	Deng LiJun	Drawing Title	TRANSFORMER	Part Number	MTEER28-151K	Rev A
Checked By:	Hui Cai					
Approved By:	ZH Tong	Tolerances: .x±0.5, .xx±0.10, .xxx±0.050			Sheet 3 of 3	

MTP-QR-014

DATA SHEET

Part No: MTEER28-151K

Date: 2014-06-13

Model: TRANSFORMER

Temperature 25 ℃ Humidity 65 %

Quantity: 5PCS

Test Equip: Chroma3250 RK2672A

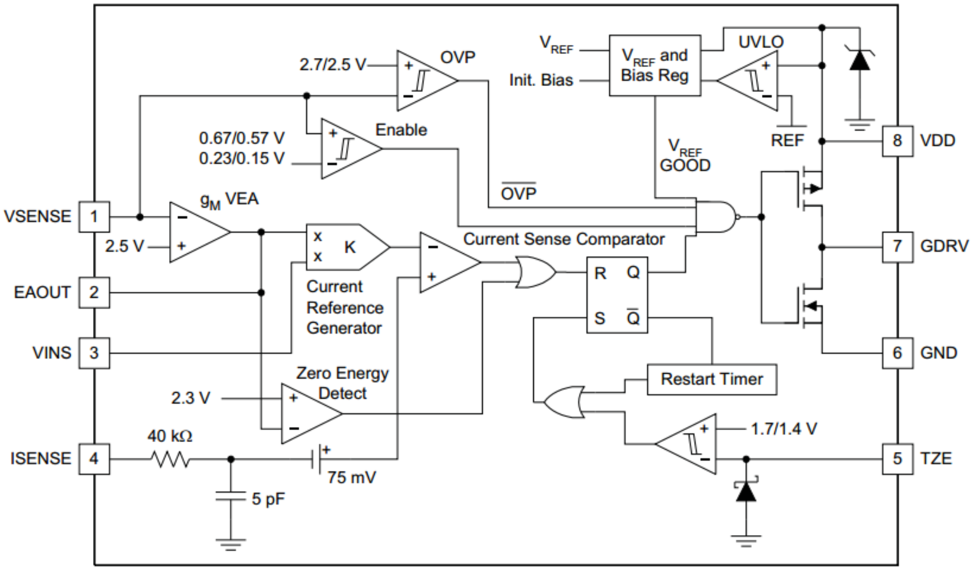
Tested	Confirmed	Approved
Deng LiJun	Hui Cai	ZH Tong

Character (特性) / Spec. (规格) / Unit (单位) / No.	L (1~3) 100KHz 0.10V	L/K(1~3) 100KHz 0.1V	DCR 1~3	DCR 10~12	DCR 5~6	TR 50.0KHz 1.0V	HI-POT	HI-POT
Spec.	150 ±10%	5.0 MAX	185.0 MAX	75.0 MAX	125.0 MAX	SPEC	Pri to Sec	Pri, Sec to Core
Unit	uH	uH	mohms	mohms	mohms		1.5KV 3Sec 5m	1.5KV 3Sec 5mA
1	146.4	1.89	119.1	55.2	66.9	OK	OK	OK
2	150.3	1.89	117.0	55.0	67.2	OK	OK	OK
3	152.4	1.90	117.5	54.0	65.4	OK	OK	OK
4	148.4	1.89	118.3	53.9	66.7	OK	OK	OK
5	152.2	1.75	118.2	53.7	67.7	OK	OK	OK
6								
7								
8								
9								
10								
11								
12								
13								
14								
15								
16								
17								
18								
19								
20								
X	149.94	1.86	118.02	54.36	66.78			
R	6.00	0.15	2.10	1.50	2.30			

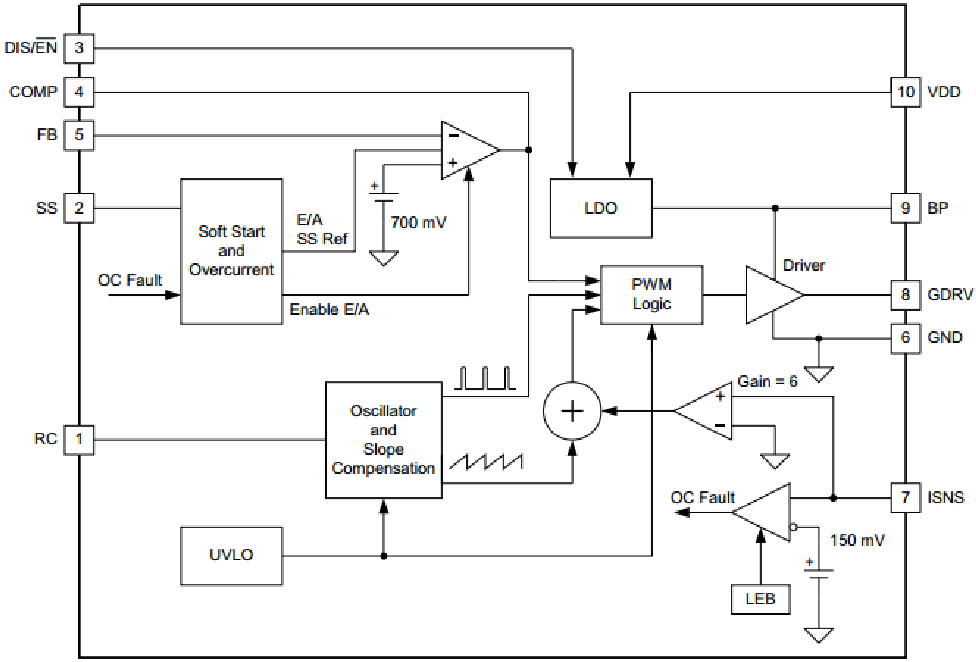
Notes:

Appendix C

C.1 Functional block diagram of UCC28810

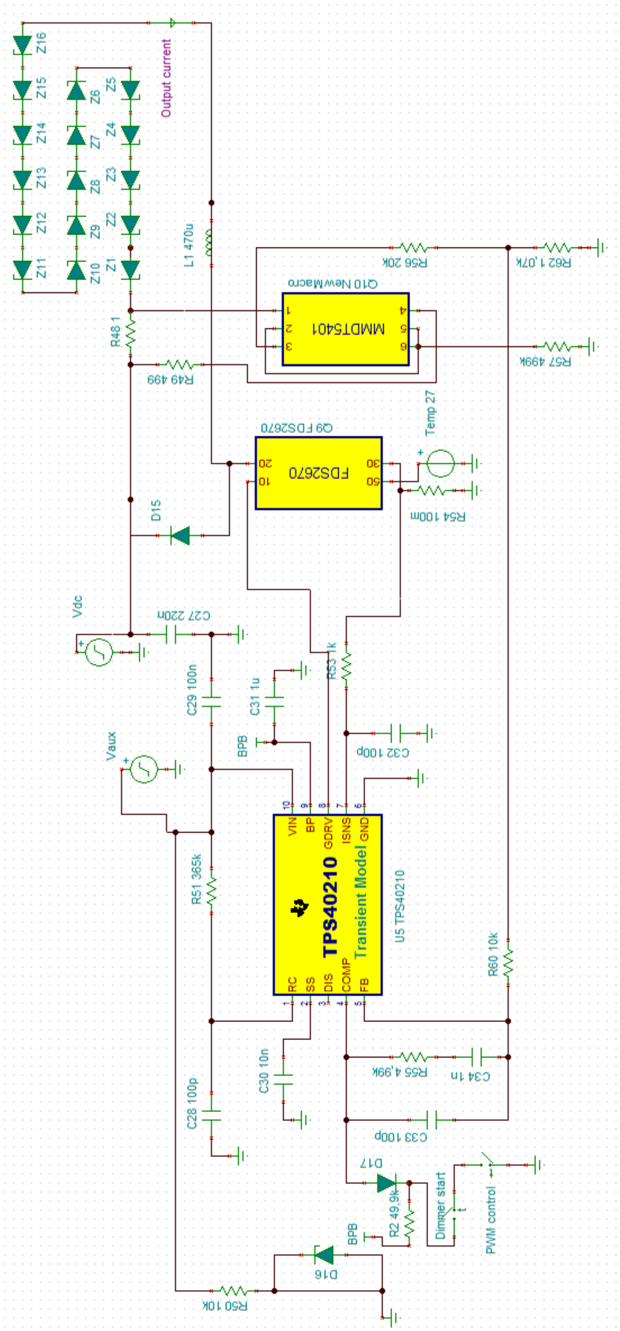


C.2 Functional block diagram of TPS40210.



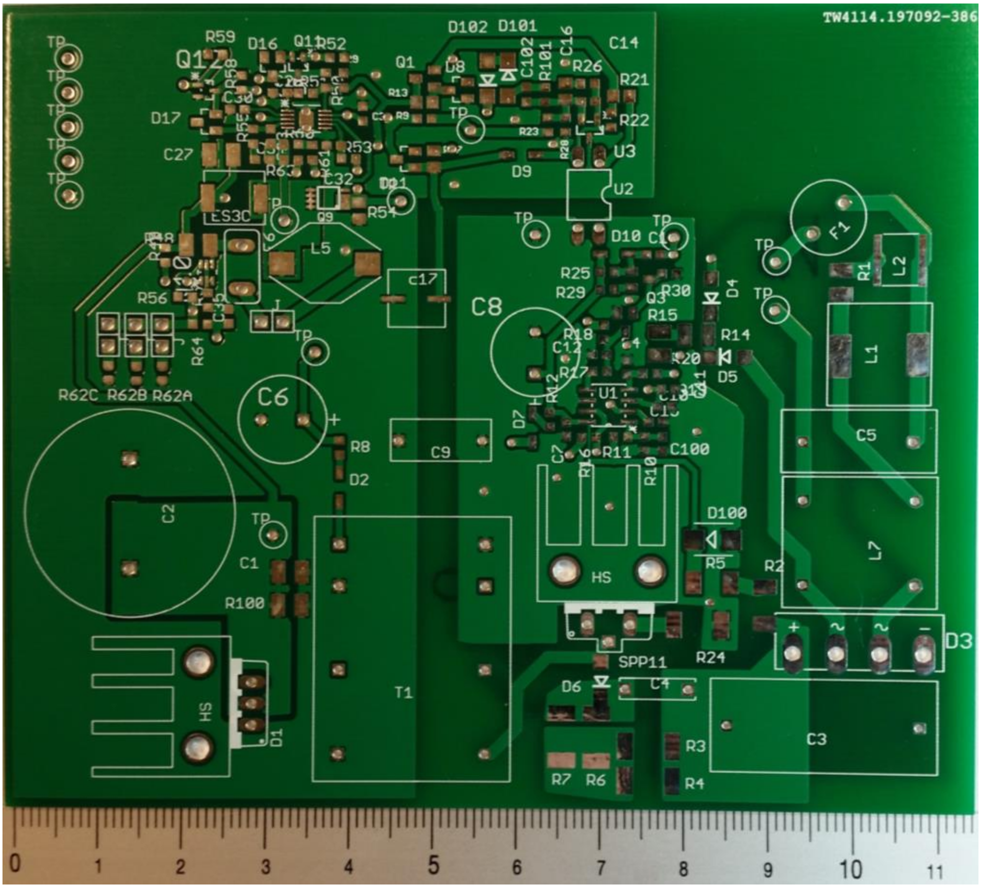
Appendix D

D.1 Simulation schematic of the driver in TINA-TI



Appendix E

E.1 The final PCB



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