TAMPERE UNIVERSITY OF TECHNOLOGY

AARNE HAIKKA

OPTIMIZING ENERGY-EFFICIENT BUCK CONVERTER AS LED CURRENT SOURCE

Master of Science Thesis

Examiner: Professor Teuvo Suntio
Examiner and topic approved in the Faculty of Automation, Mechanical and Materials Engineering Council meeting on 4. April 2012

ABSTRACT

TAMPERE UNIVERSITY OF TECHNOLOGY
Master's Degree Programme in Automation Engineering
HAIKKA, AARNE: Optimizing energy-efficient buck converter as LED current source
Master of Science Thesis, 66 pages, 12 Appendix pages
April 2013
Major: Switched-mode converter design
Examiner: Professor Teuvo Suntio
Keywords: LED current source, buck converter, light level control, PWM dimming, analog dimming, hybrid dimming.

Energy-efficiency demands get tighter and tighter and traditional lighting devices don't satisfy the demands. LED-lighting efficiency is high, therefore they are a good choice for replacing traditional lighting devices. There are a vast variety of LED-modules on the market, which are setting high demands for an LED current source, especially when high quality dimming is demanded. There is the intention to design a high quality LED current source in this thesis.

LED-modules need current flow through them, which is produced by a constant current step-down switched-mode converter. A switched-mode converter has a high efficiency and a small size. Buck converter output current is the same as inductor current. When inductor current is desired then output current is too. Buck converter control can be implemented in many ways, all of them based on precise inductor current feedback. A control system controls the main switch ON and OFF with a very high frequency and inductor current average value depending on the inductance of the inductor, the switching frequency and the duty ratio.

Although a buck converter efficiency is high, it produces heat, which has to be minimized. This heat shortens the lifetime of the components and increases the total power loss. However, switched-mode converters have a much better efficiency than linear regulators. Since the operating frequency is high, the converter produces electromagnetic interference to its input, output and to the air. These interferences have to be minimized by a good design.

An LED-modules light output is proportional to the current through it. Dimming can be implemented by reducing the average current flow through it. There are three dimming methods: blocking LED current flow rapidly with a specific duty ratio**,** reducing the current value linearly, or combining both earlier dimming methods. All dimming methods have pros and cons. Dimming may have the following problems: visible steps between light levels, flickering, stroboscopic effect, audible noise and possible color change.

On the basis of the study, a prototype device was designed in which efficiency and power losses was simulated and measured. The prototype device is hybrid dimmable and its output is optimized for human eye sensitivity. Especially the hybrid dimming results are encouraging, because the dimming is step-less for the human eye. In addition the main circuit is chosen so that the control circuit can be implemented by using low voltage levels, and then a microcontroller can be used.

TIIVISTELMÄ

TAMPEREEN TEKNILLINEN YLIOPISTO
Automaatiotekniikan koulutusohjelma
HAIKKA, AARNE: Optimizing energy-efficient buck converter as LED current source
Diplomityö, 66 sivua, 12 liitesivua
Huhtikuu 2013
Pääaine: Teholähde-elektroniikka
Tarkastaja: Professori Teuvo Suntio
Avainsanat: LED virtalähde, buck teholähde, valotason säätö, pulssinleveys himmennys, analoginen himmennys, hybridi himmennys.

Energiatehokkuusvaatimukset vaativat kehittämään vaihtoehtoja perinteisille valaistusratkaisuille. LED-valaistuksen hyötysuhde on korkea, joten se on hyvä vaihtoehto korvamaan perinteisiä valonlähteitä. Markkinoilla olevien LED vaihtoehtojen laaja kirjo asettaa LED-liitäntälaitteelle korkean vaatimustason. Kun siihen lisätään vielä tarve laadukkaalle himmennykselle, vaatimukset korottuvat entisestään. Työssä on tarkoitus suunnitella hyvin toimiva himmennettävä LED virtalähde.

LED-moduuleille syötetään virta, joka tuotetaan vakiovirtaisella jännitettä laskevalla buck-hakkuritopologialla. Hakkuria käytetään, jotta virtalähteestä saadaan kooltaan pieni ja mahdollistetaan korkea hyötysuhde. Buck-tyyppisessä hakkurissa kelan virta on sama kuin ulostulovirta, joten pitämällä kelan virta haluttuna myös ulostulovirta on oikea. Virtalähteen säätöön on monia tapoja, kaikissa tavoissa virranmittaus on keskeinen tekijä, joten mittauskytkentä on oltava hyvin suunniteltu. Säätöjärjestelmä ohjaa hakkurin kytkintä päälle ja pois nopealla taajuudella ja kelan virran keskimääräinen arvo määräytyy kelan induktanssin, kytkentätaajuuden ja pulssisuhteen funktiona.

Vaikka hakkurin hyötysuhde on korkea tuottaa se silti häviöitä, jonka määrä pyritään minimoimaan. Häviöteho huonontaa hakkurin hyötysuhdetta ja muuttuu lämmöksi, joka lyhentää komponenttien elinikää ja huonontaa hyötysuhdetta entisestään. Kuitenkin hakkureiden hyötysuhde on ylivoimainen verrattuna lineaarisiin virtalähteisiin. Hakkureiden korkeasta kytkentätaajuudesta johtuen niiden ongelmana on korkeataajuiset elektromagneettiset häiriöt niin sisääntuloon, kuin ulostuloon sekä säteilynä ilmaan. Nämä häiriöt voidaan minimoida hyvällä suunnittelulla.

LED-moduulia himmennetään pienentämällä sen läpi kulkevaa keskiarvovirtaa. LED-virtalähteen lähtövirtaa voidaan pienentää kolmella eri tavalla: katkotaan virran kulkua nopeasti tietyllä pulssisuhteella, pienennetään virran arvoa lineaarisesti tai näiden yhdistelmällä. Himmennystavoilla on omat hyvät puolensa ja niillä on myös omat ongelmansa. Himmennyksen ongelmiin kuuluvat selvät portaat valotasojen välissä, välkyntä, stroboskooppinen efekti, häiritsevä ääni ja mahdolliset valon värin muutokset.

Tutkimuksen pohjalta suunniteltiin prototyyppilaite, jonka hyötysuhteita ja tehohäviöitä tutkittiin ensin simulointiohjelmistolla ja sen jälkeen mittaamalla käytännössä. Prototyyppiin toteutettiin hybridihimmennys, jonka ulostulo pyrittiin optimoimaan silmän herkkyyden mukaan. Varsinkin hybridihimmennyksen tulokset ovat rohkaisevia, portaattomalta näyttävän himmennyksen ansiosta. Lisäksi piiriratkaisu valittiin siten, että sen ohjaus on helposti toteutettavissa mikrokontrollerilla.

PREFACE

This Master of Science thesis has been written for Helvar Oy, the examiner was Professor Teuvo Suntio from Tampere University of Technology and supervisors from Helvar was M.Sc. Hannu Vihinen and M.Sc. Jari Heine.

I want to thank Helvar and Max Björkgren for offering me interesting topic for this thesis. Also I want to thank Professor Teuvo Suntio for his advice and guidance during writing process and for examining. I want to thank Hannu and Jari for their valuable advice and guidance during writing, special thanks to Hannu for good tips for the prototype device design and technical guidance during writing. Also big thanks for Leo Haapavuo for the prototype device PCB design and Robert Hutson for spelling and grammar check. Huge thanks belong also to my other colleagues for support and advices.

I would also like to thank my family for their support during writing process and also during my studies.

Karkkila 29.04.2013

Aarne Haikka

TABLE OF CONTENTS

ABBREVIATIONS AND NOTATION

EMI	Electromagnetic interference.
IC	Integrated circuit.
PCB	Printed circuit board.
SELV	Safety extra low voltage.
Δi	Inductor current hysteresis.
Δi_{LED}	Output current ripple.
Δi_{off}	Inductor current change during OFF-time.
Δi_{on}	Inductor current change during ON-time.
C_{gd}	MOSFET parasitic capacitance between gate and drain.
C_{gs}	MOSFET parasitic capacitance between gate and source.
D	Duty ratio.
D'	Inverse of the duty ratio.
f_s	Switching frequency.
I_{in}	Input current.
I_{nom}	Nominal output current.
I_{out}	Output current.
I_{target}	Desired output current.
L	Inductance of the inductor.
L_n	N:th light level.
r_C	The parasitic resistance of the capacitor.
r_{DS}	The parasitic resistance of the MOSFET.
r_L	The parasitic resistance of the inductor.
r_{LED}	LED string dynamic resistance.
t_{off}	Main switch OFF-time.
t_{on}	Main switch ON-time.
T_{onpwm}	PWM signal ON-time.
T_{pwm}	PWM signal switching time.
T_s	Main switch switching time.
U_D	Voltage over a diode.
U_{ds}	MOSFET drain to source voltage.
U_{gs}	MOSFET gate to source voltage.
U_{in}	Input voltage.
U_{out}	Output voltage.

1 INTRODUCTION

Nowadays energy saving is important, because energy is getting more and more expensive and energy producing causes pollution. That's why governments all over the world are setting energy-efficient requirements for devices which use energy, and traditional lighting devices don't satisfy the demands. Lighting is one segment which has high energy efficient demands. LED lighting is a good solution to save energy, but an LED current source needs optimizing. The optimized current source is energy-efficient and produces a stable current for an LED string. The optimized device is also dimming LEDs step-less and logarithmically. An LED current source which meets those demands is very sophisticated and it produces high quality light.

This thesis describes how a basic buck-type switched-mode converter works and what kind of components it consists of. Converter power losses are also an important thing because they have to be minimized. It is important to understand how any component affects to the total power loss because increasing efficiency somewhere, decreases efficiency somewhere else.

High quality dimming is also setting many demands for the LED current source, because the dimming curve has to be logarithmic for the eye sensitivity and dimming has to be step-less. The lowest light output level might be 1%, or even lower 0,1%, which is very difficult to achieve by using traditional dimming techniques. This thesis describes traditional dimming techniques as well as some special techniques generally.

The objective of this thesis is to optimize a buck converter and making it modular. Then the same converter can used in many lighting devices which have different output powers. The buck converter has to be energy-efficient and the output current has to be suitable for the LED modules on the market. Also high quality dimming is demanded, which means that dimming has to be smooth for the human eye, therefore flickering and any steps is prohibited.

The thesis is divided into seven Chapters. The first is an introduction. The second chapter focuses on the basic buck converter and dimming methods theory. The third chapter describes what kind of problems a buck converter and dimming designer can face, and this Chapter also tries to find solutions for these problems. In the fourth chapter the simulation model and simulation results are shown. Simulations consist of operating frequency and efficiency simulations in different operation modes. The fifth chapter describes the prototype device and how components are chosen in the main circuit and the control circuit. The sixth chapter focuses on the results of how the prototype device is working. Efficiency measurements are introduced as well as general operation like operation frequency, inductor current ripple and output current ripple. These measurements are done with different loads. In Chapter six the dimming measurement are also shown. They consists of dimming curve and analog dimming frequency change measurements. Finally, the seventh chapter is the summing up.

2 THEORY

A Buck converter is a step-down-type DC-DC switched-mode converter which means that the output voltage is always lower than the input voltage, and efficiency is high because switches are only at ON- or OFF-modes. Because of that, buck converters are widely used in regulated power sources like constant current sources. (Mohan et al. 1995, pp. 164.). The needed load voltage is known and on the basis of that the input voltage can be chosen to be some amount higher than the needed load voltage to achieve proper operation. At the same time the load needs constant current, like LED current source solutions. Therefore the buck converter is a good choice.

The buck converter can regulate its output voltage or output current and the control method is dependent on that (Maxim 2001, pp. 5). This thesis focuses only on the constant current sources, because an LED load needs constant current while at the same time the only demand for output voltage is that it has to be over the LED string threshold voltage. There are also constant voltage LED-modules, which need constant voltage and their power source is a constant voltage source. Those LED-modules have their own current limiting element inside them: Usually this is done by using a resistor or a linear regulator.

2.1 Buck Alternatives

There are two basic buck alternatives which consist of an inductor, a filtering capacitor and two switches: a switching transistor and a diode. Both topologies have the same basic principles in which way they work and that’s why the same equations are valid.

Figure 2.1 shows the basic topology where the main switch, usually the MOSFET, and the inductor is high-side (Mohan et al. 1995, pp. 165). The diode offers a path for the inductor current when the main switch doesn’t conduct and the capacitor filters the output current and voltage.

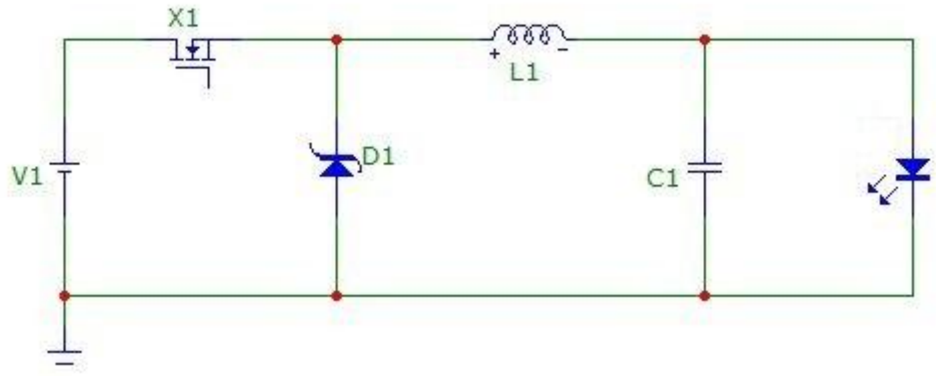

Figure 2.1. *A high-side buck alternative.*

The main switch is controlled ON and OFF at a high frequency, usually several hundred kilohertz, which means that the switching time is a few microseconds. At the ON-time current flows through the main switch to the inductor and continues to the capacitor and to the load at the same time the diode is reverse biased. Current rises linearly and a rising slope depends on the inductor inductance, as shown in equation 2.1, which is written without diode threshold voltage or any parasitic resistances. When increasing current flows through the inductor, the inductor magnetic field energy increases too. Voltage across the inductor is positive and the output voltage is lower than the input voltage. In equation 2.1 U_{in} is the input voltage, U_{out} is the output voltage and L is the inductor inductance. (Mohan et al. 1995, pp. 165-166.)

$$\frac{\Delta i_{on}}{t_{on}} = \frac{U_{in}-U_{out}}{L} \qquad (2.1.)$$

At the OFF-time the main switch is turned off and the inductor is keeping current flowing with its magnetic field energy storage. Falling current flows through the diode and falls with the slope in equation 2.2, which is written without diode threshold voltage or any parasitic resistances. Current falls because the inductor magnetic field energy is also decreasing. The voltage across the inductor is equal to the output voltage at the OFF-time but it is negative. In equation 2.2 U_{out} is the output voltage and L is the inductor inductance. (Mohan et al. 1995, pp. 165-166.)

$$\frac{\Delta i_{off}}{t_{off}} = \frac{-U_{out}}{L} \qquad (2.2.)$$

Current increases and decreases during every switching cycle, and the slopes depend on the voltage across the inductor and the inductance value of the inductor. In a steady state the absolute change of current during ON- and OFF-time has to be the same, otherwise the output current is not stable. The absolute current change can be calculated by multiplying the ON-time of the current rising slope or multiplying the OFF-time of the current falling slope as shown in equation 2.3. The absolute current change, which is also called the ripple, is an important thing when designing a switched-mode converter. When the output voltage is half of the input voltage the slopes are equal and then ON- and OFF-times are equal too, but this is a special case and usually slopes aren't equal and that's why ON- and OFF-times aren't equal either. In equation 2.3 Δi is the absolute value of the current change during ON- or OFF-time, D is the conversion ratio, which is the ratio between the main switch ON-time and the total switching time, D' is 1-D and T_s is the switching time. (Mohan et al. 1995, pp. 170-172.)

$$\Delta i = \left|\frac{\Delta i_{on}}{t_{on}} \cdot D \cdot T_s\right| = \left|\frac{\Delta i_{off}}{t_{off}} \cdot D' \cdot T_s\right| \tag{2.3.}$$

Figure 2.2 shows second basic alternative where the main switch and the inductor is low-side (ON Semiconductor 2011, pp. 2). This alternative is also called a floating or inverse buck.

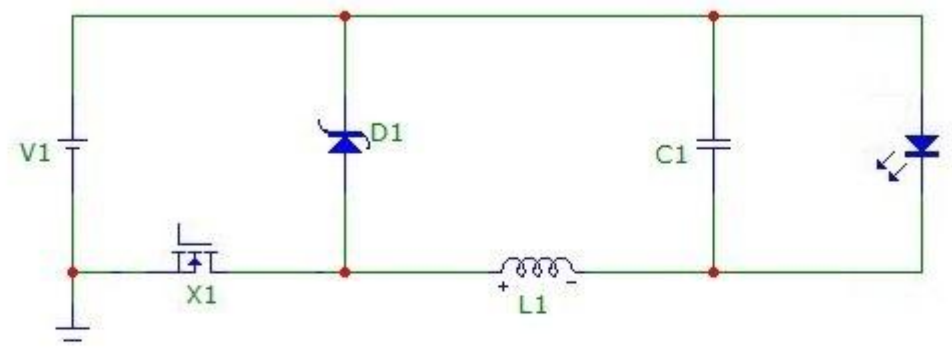

Figure 2.2. *A low-side buck alternative.*

Buck converter basic equations in the operating point are: a voltage equation and a current equation. The operating point is the steady state point, when all transients are attenuated and the output voltage and current are steady as well as the input voltage and current.

Equation 2.4 is the voltage equation, U_{out} is the output voltage, D is the conversion ratio, U_{in} is the input voltage, D' is 1-D, U_d is the diode threshold voltage, r_L is the inductor parasitic resistance and I_{out} is the output current. This equation shows that output voltage is always lower than input voltage because the conversion ratio D is between 0-1.

$$U_{out} = D \cdot U_{in} - D' \cdot U_d - r_L \cdot I_{out} \tag{2.4.}$$

Equation 2.5 is the current equation; I_{in} is the input current, D is the conversion ratio and I_{out} is the output current.

$$I_{in} = D \cdot I_{out} \qquad (2.5.)$$

2.2 Operation Modes

Switched-mode converters can be classified in many ways. One way is using operation modes which categorize converters with an inductor current waveform. This classification system has three different categories: a continuous conduction mode (CCM), a boundary conduction mode (BCM) and a discontinuous conduction mode (DCM). (Mohan et al. 1995, pp. 165-170.)

2.2.1 CCM

In continuous conduction mode, inductor current flows continuously through the inductor. In Figure 2.3 voltage V_{GS} is the MOSFET gate voltage which controls the MOSFET ON- and OFF-times. I_L is the inductor current, I_{in}, red, is the input current and current through the MOSFET and I_D, green, is the diode current. As Figure 2.3 shows, the inductor current rises during every ON-time and falls during every OFF-time, which causes some ripple. The inductor current rising and falling slopes are different which means that ON- and OFF-times have to be different too, as told in Section 2.1 The ripple amount depends on the slopes as well as the ON- and OFF-times. Average inductor current is the mean of the peak current and valley current. (Mohan et al. 1995, pp. 165-167.)

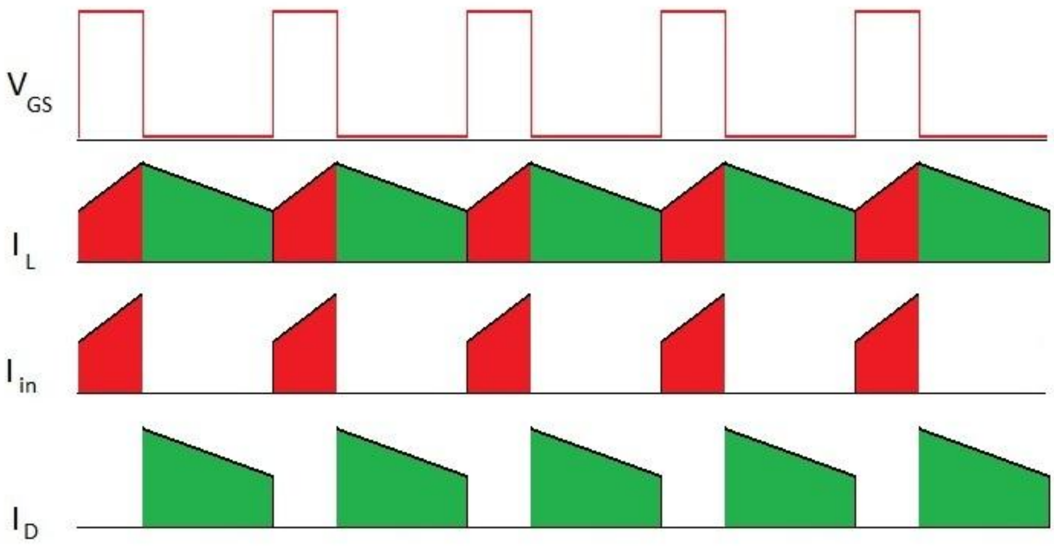

Figure 2.3. *Current waveforms in the continuous conduction mode.*

The CCM is a very common operation mode for a switched-mode converter, because current ripple can be low enough and demands for output filtering are lighter than in

BCM or DCM. Because of the low current ripple, the inductance of the inductor has to be higher than other operation modes to achieve the same operational frequency, and higher inductance means a bigger inductor.

2.2.2 BCM

In boundary conduction mode the inductor current ripple is the same as the peak current. The valley current value is zero as shown in Figure 2.4 When the inductor current reach zero the main switch is set ON and current starts rising again. In Figure 2.4 symbols and colors are the same as Figure 2.3. (Mohan et al. 1995, pp. 167-168.)

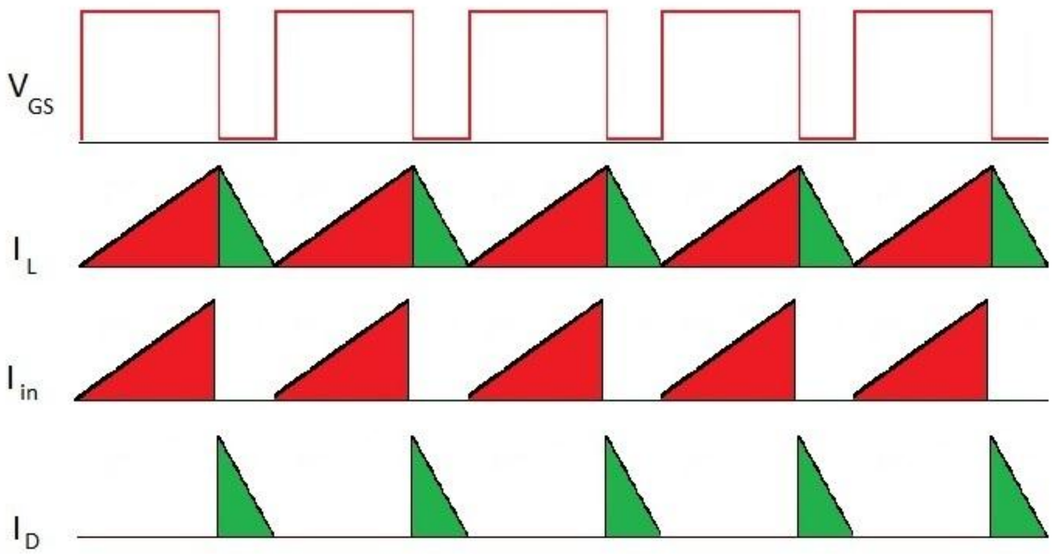

Figure 2.4. *Current waveforms in the boundary connection mode.*

In the BCM the current ripple is the highest it can be, as high as the peak current, which is setting high demands for output filtering. Respectively, the inductance of the inductor can be lower and this usually means a smaller size. The main switch is turned on when inductor current is zero, which means that turn on losses are minimized and the speed of the diode can be slower. Thus the diode can be more cost-effective.

2.2.3 DCM

In discontinuous conduction mode the inductor current reaches zero before the switching cycle ends, thus inductor energy is zero and no current flows to the load through the inductor as shown in Figure 2.5. In Figure 2.5 the symbols and colors are the same as Figure 2.3. (Mohan et al. 1995, pp. 168-170.)

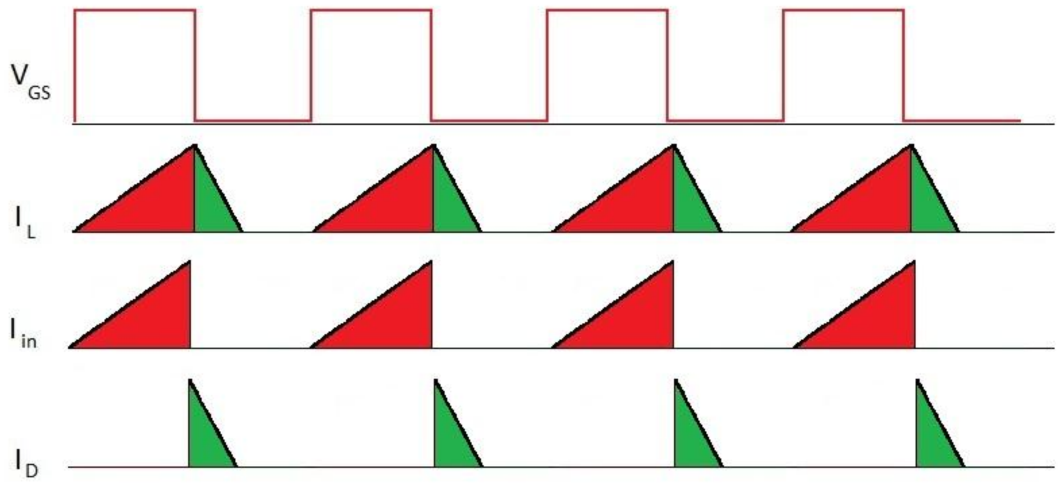

Figure 2.5. *Current waveforms in the discontinuous conduction mode.*

This operational mode is setting high demands for the output capacitor, which has to provide all the output current when the inductor current is zero. That's why the output capacitor has to have much higher capacitance compared to other operational modes, especially the CCM, but the inductor size can be smaller because the ripple is higher.

2.3 Control Methods

There are many methods to control a buck converter. The choice between different methods depends on the chosen buck alternative and how current is measured. A control circuit can be found from manufacturer's IC catalogs or it can be implemented by using discrete components. In this Section the most common control methods are presented.

2.3.1 Hysteresis

A hysteresis control is a method which is trying to keep the inductor current between specific thresholds; an area between peak- and valley-values is called a hysteresis band. In Figure 2.6 the hysteresis band is shown. I_L is the inductor current, B_1 is the peak value and B_2 is the valley value. When the main switch is ON, the inductor current rises and when it reaches the peak value the main switch is set OFF and current flows through the diode and starts falling. After some time current reaches the valley value and the main switch is set ON and then the cycle starts again. That's why the hysteresis band is setting the inductor current ripple. (Texas Instruments 2006.)

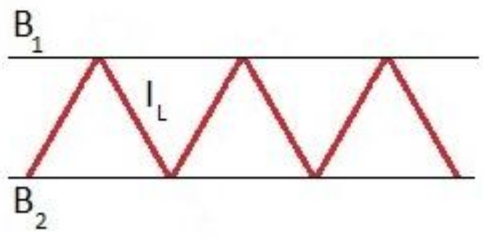

Figure 2.6. *A hysteresis band.*

The hysteresis control can be implemented by using an SR-flip-flop, which controls the MOSFET and two comparators having different reference voltages as shown in Figure 2.7. The first comparator output is connected to the flip-flop set input and another is connected to the reset input. The first comparator sets the flip-flop on when the voltage across the current sense resistor reaches the comparators reference voltage. Another comparator resets the flip-flop when the current sense resistor voltage reach its reference voltage. Thus the first comparator's reference voltage is setting the current valley value and the second comparator's reference voltage the peak value of current.

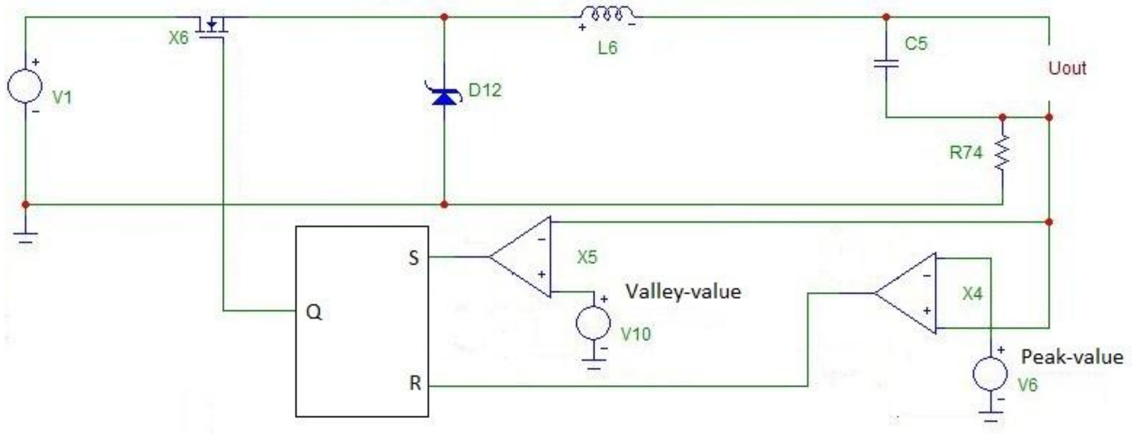

Figure 2.7. *A hysteresis control circuit.*

If more than one hysteresis band is demanded the reference voltages have to be adjustable or a current sense resistor has to be changeable to achieve that. Adjustable references or a changeable resistor allow many output current levels and changing hysteresis. This increasing the LED current source dynamics, because the same converter can drive LEDs with different forward currents. This control method with adjustable references allows analog dimming too.

The second way to implement hysteresis control is using a hysteretic comparator which is connected to the MOSFET-driver circuit (Texas Instruments 2006). This implementation needs only one reference voltage and hysteresis band to be determined by the comparator's hysteresis. If this reference voltage is adjustable, it allows many output current levels and analog dimming too.

In real life comparators aren't ideal, that's why thresholds are not precise. The comparator and other circuit delays are one source of inaccuracy. Current has some overshoot at the peak threshold and undershoot at the valley threshold as shown in Figure 2.8. In Figure 2.8 B_1 is the peak value, B_2 is the valley value, I_L the inductor current and t_d is the total delay. In some cases this might be a problem especially if the current speed of rise or fall is high enough and the delay is quite long compared to the switching time. Delays increases the current ripple and decreases the switching frequency and this must be noticed when designing the control system. (Texas Instruments 2006.)

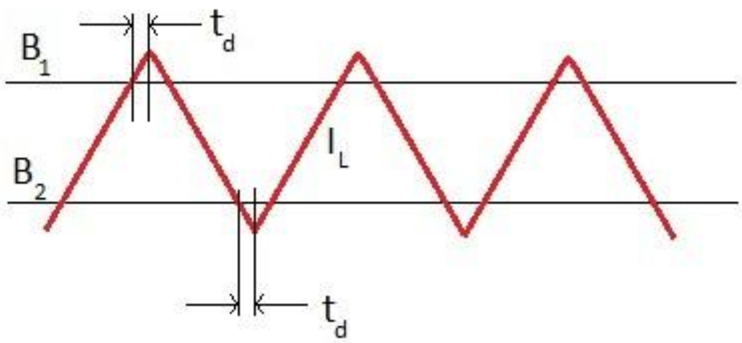

Figure 2.8. *Comparator delay.*

As shown in Figure 2.6 and 2.8, the comparator delay decreases the switching frequency and increases the current ripple. Current rise and fall is a little bit longer and it takes more time and causes some over- and undershoot. In Figure 2.8 rising and falling slopes are equal. This means that the mean of the inductor current is demanded. If slopes aren't equal the mean of the inductor current changes. If a rising slope is higher the mean is higher too, and respectively, if the falling speed is higher the mean is lower.

A buck converter, which is controlled by hysteresis controller with B_1 and B_2 above zero, operates in the CCM. Average current and current ripple is designed to meet the LED string and output filtering demands. In some cases the valley threshold B_2 is zero and then converter operates in the BCM.

2.3.2 Constant OFF-time

Constant OFF-time control is based on keeping the main switch's OFF-time in a steady state. Current falls during that OFF-time and the amount of that decrease can be calculated by using the voltage across the inductor, the inductance of the inductor and the OFF-time. The peak value is set by a comparator and its reference voltage. Constant

OFF-time affects that the valley value of inductor current is changing and it depends on load voltage, the inductor inductance and the OFF-time. The inductance and the OFF-time are constant, therefore the valley value is inversely proportional to the output voltage. (ON Semiconductor 2011.)

The constant OFF-time is a good choice to control a buck converter whose MOSFET and inductor is low-side, and then the current sense resistor voltage measures only the ON-time current. If the current sense resistor is in the positive supply line and the control circuit measures current with it, then the control circuit has to handle high voltages. Usually control circuits are more cost-effective when they use a low-side current sense resistor, and that's why the constant OFF-time is a good way to control the low-side buck converter. (ON Semiconductor 2011.)

The constant OFF-time is generated by an OFF-time generator, which can be implemented by charging or discharging a capacitor with constant current. Then the capacitor voltage increases linearly and time can be determined by comparing capacitor voltage to a reference. The time can be changed by modifying the charge or discharge current, changing the reference voltage or changing the timing capacitor. One way to generate constant current is using a current mirror whose current is determined by a resistor. When the converter is operating, current can be modified by a BJT which changes resistor's opposite end voltage. The OFF-time has to be controllable if the output voltage changes or the LED load is dimmed by analog dimming. Figure 2.9 shows the controllable constant OFF-time generator, which is adjusted with voltage source V_{var}.

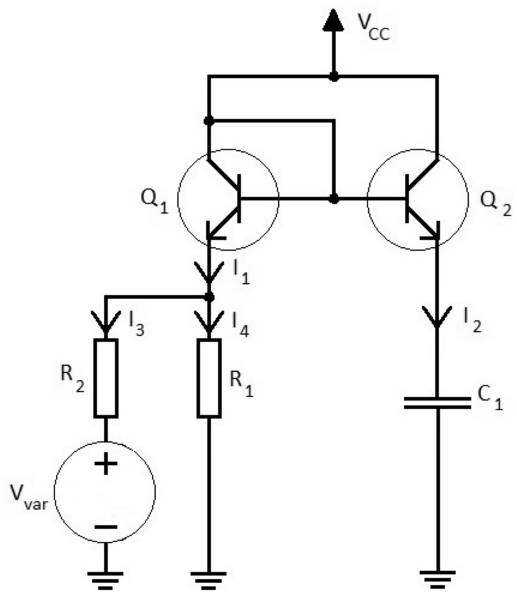

Figure 2.9. *A constant OFF-time generator.*

In Figure 2.9 Q_1 and Q_2 are the current mirror, C_1 is the timing capacitor, R_1 and R_2 are the timing resistors and V_{var} is the adjustable voltage source. This circuit charges the capacitor with constant current I_2 which is equal to I_1. I_1 can be determined by the adjustable voltage source and timing resistors. Resistor R_1 determines current I_4 which is the charging current default level, and with resistor R_2 and voltage source V_{var} the charging current can be increased. A lower voltage source voltage means a higher I_3 which increases charging current too. The capacitor voltage reaches a specific level within a specific time and this time is the constant OFF-time. When the capacitor has reached that level the main switch is turned on and the timing capacitor is discharged and its voltage is kept at zero level for the whole ON-time. At the beginning of the OFF-time the capacitor charging is started again and the cycle continues.

When using capacitor discharging, a sinking current mirror has to be used. OFF-time control can then be done by changing the discharge current but also changing the capacitor initial voltage. When using a microcontroller the constant OFF-time generator can be implemented by using an AD-conversion or timer interrupts.

2.3.3 Constant Frequency

In a constant frequency control the switching frequency is the same the whole time. The most common way to implement that kind control circuit is to set a flip-flop on with a constant frequency pulse and a comparator sets the flip-flop off when the current sense resistor voltage reaches the desired level. This flip-flop controls the buck converter main switch.

The constant frequency causes variable inductor current hysteresis. The hysteresis depends on input and output voltages. The constant frequency is good choice for solutions where frequency has to be same all the time, otherwise there are better choices for control of the buck converter.

2.4 Measurement and Feedbacks

To achieve stable operation and the desired output it is important to design measuring and feedback circuits correctly. This consists of where to measure, how to measure and how to use measurements, like scaling or changing current to voltages. Sometimes measurements need filtering to remove high frequency noise or a leading edge blanking to guarantee a comparators proper operation, by removing leading edge spikes which can trig flip-flops too early.

When implementing a control circuit for a current source, proper inductor current measurement is very important, because the whole control loop is based on the inductor current feedback. The easiest way to measure current is using a current sense resistor which changes current to voltage with Ohm’s law. The place of the current sense resistor is important. (Mohan et al. 1995, pp. 337-340.)

There are several possible locations for the current sense resistor. These locations are shown in Figures 2.10 and 2.11. Every place has pros and cons. The best results can be achieved when the current sense resistor measures pure inductor current over the whole switching cycle. This is the demand for hysteresis control.

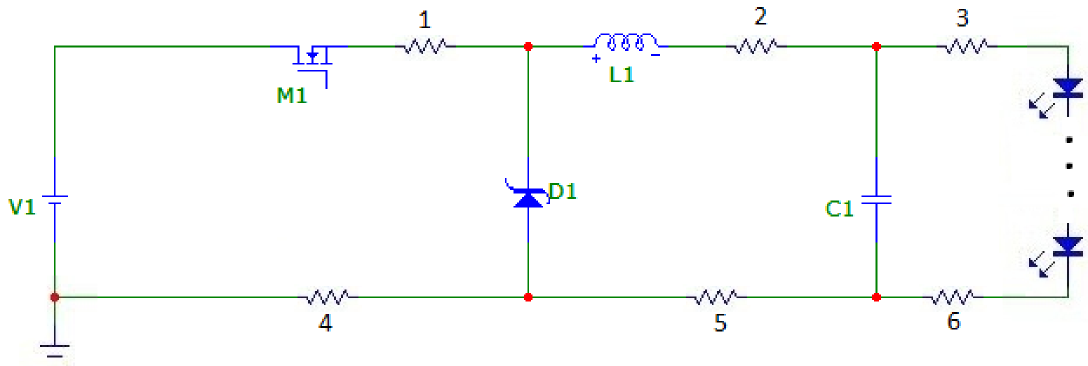

Figure 2.10. *Current sense resistor locations, when using a high-side buck converter.*

In Figure 2.10 resistors number one and four measure only input current, this measurement method is suitable when using the constant OFF-time control method. Resistors number two and five are the best for measuring, because they measure pure inductor current which is demand for hysteresis control. Resistors number three and six measure the LED load current which is the filtered inductor current. These places aren't good for the buck converter feedback because the best results cannot be achieved without precise inductor current measurement.

When using a low-side current sense resistor, the measurement circuit is easier to design and more cost-efficient because one end of the resistor is connected to earth and the current can be calculated by measuring the voltage over this resistor. When high-side current sense resistor is used, the measurement has to be done differentially and it may be expensive, but then the negative supply line is free and it can be common to all LED strings.

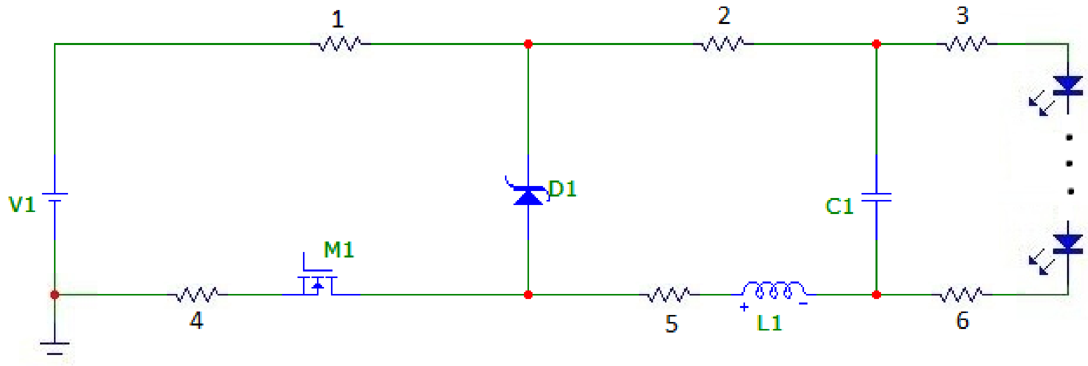

Figure 2.11. *Current sense resistor locations, when using a low-side buck converter.*

In Figure 2.11 resistors number one and four measure input current and they are suitable for the OFF-time control method feedback resistors. When using the resistor number four, the measurement circuit is cost-efficient because one end of the resistor is connected to earth and the current can be calculated by measuring voltage over this resistor, but the hysteresis control cannot be implemented with this resistor. Resistors two and five are suitable for the hysteresis control, but the measurement circuit has to handle high voltages and measurements have to be done differentially. Resistors three and six measure the LED string current and, as said before, they aren't a good choice for the buck converter feedback resistors.

2.5 Power Losses

As told in Chapter 1, energy-efficiency demands are high for lighting products and demands get tighter and tighter. That's why even LED current sources have to be developed the whole time to reach new efficiency demands. Usually the buck converter has efficiency up to 90-95% and highest power losses are generated by the MOSFET, the diode and the inductor: the MOSFET has switching and conduction losses, the diode has switching and conduction losses too, and the inductor has wire and core losses. Also other components generate losses like the current sense resistor, the driver IC and the output filtering capacitor.

Power losses heat components and their temperature rises. Rising temperature increases for example resistive component power losses, which further increases power losses. That's why power losses and temperature have to minimized by good design. If temperature rises too high the current source needs cooling or component lifetime gets shorter. Especially electrolytic capacitors are sensitive to heat.

2.5.1 MOSFET

MOSFET losses can be divided into switching and conduction losses. Conduction losses are generated by a drain to source resistance and switching losses are generated by properties of the MOSFET and the control circuit. Losses during one switching cycle are shown in Figure 2.12. U_{ds} is the voltage across MOSFET drain and source, I_d is the current through MOSFET, U_{gs} is the voltage across MOSFET's gate and source. The total power loss is shown by the green line and it is divided into turn-on, turn-off and conduction losses (P_{on}). (Mohan et al. 1995, pp. 583-589.)

The MOSFET has the resistance (r_{DS}) between drain and source, which produces a voltage drop across the MOSFET. When current flows through the MOSFET, there has to also be power loss, as shown in Figure 2.12. This loss can be calculated with the Ohm's law. Different MOSFETs have a variety of r_{DS} and naturally lower resistance generates lower power loss. According to Ohm's law the relation between power loss and current is quadratic. This means that current doubling generates four times higher power loss. (Mohan et al. 1995, pp. 583-589.)

Switching losses are generated by properties of the MOSFET and the driving circuit: total gate charge and gate driving current are the most important factors when calculating MOSFET switching losses. Load also affects to the losses, like diode and inductor capacitances and diode speed.

The turn-on procedure is shown in Figure 2.12. First the gate driving signal starts rising and when it reaches the threshold level, at the time t_1, the MOSFET starts to conduct. Between times t_1 and t_2 the MOSFET is operating in the linear zone and then current is proportional to the gate voltage and voltage across the MOSFET is maximum. When the gate voltage continues rising then current through the MOSFET rises too. At the end of this period the gate voltage reaches the Miller plateau level, at the time t_2, and at same time I_d reaches its maximum level. In the case of an LED current source I_d reaches the inductor current level and after that I_d starts to grow as told in Section 2.1. Between t_2 and t_3 gate voltage is constant. This is the Miller plateau. Drain to source voltage is falling almost to zero within this period. This plateau is a consequence of discharging the parasitic capacitor between gate and drain, because all the gate driving current discharges C_{gd} and no current charges C_{gs}. After the time t3 the gate voltage continues rising to its maximum level and this decreases r_{DS} and U_{ds}. (Mohan et al. 1995, pp. 583-589.)

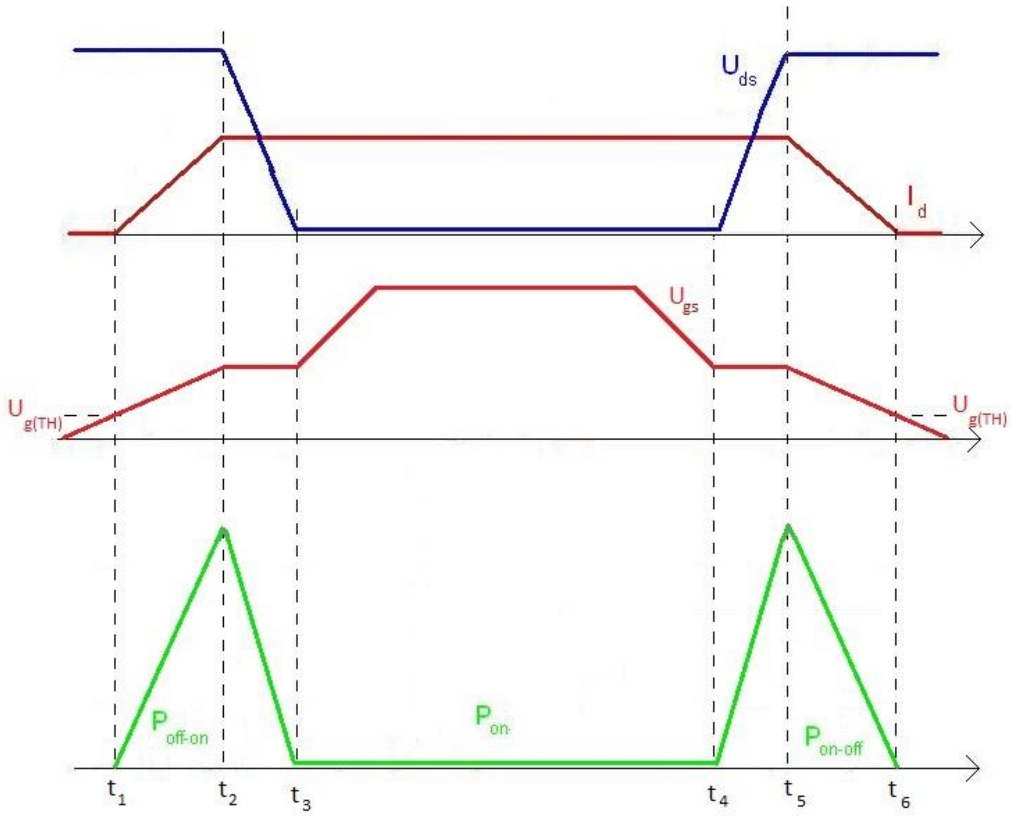

Figure 2.12. *Power losses of a MOSFET.*

The turn-off procedure is also shown in Figure 2.12 and it is similar to the turn-on procedure but backwards. First U_{gs} is falling to the Miller plateau level, voltage reaches that plateau at the time t_4. Between t_4 and t_5 the gate voltage is constant, the parasitic capacitor C_{gd} is being charged and U_{ds} rises from almost zero to maximum. At the time t_5 C_{gd} is charged and U_{ds} reaches its maximum level , after that the MOSFET operates in the linear zone and gate driving current starts to discharge the parasitic capacitor C_{gs} and gate voltage starts falling. The current through the MOSFET starts falling because in the linear zone the current is proportional to gate voltage. When the gate voltage reaches its threshold level, at the time t_6, the current has reached zero and the MOSFET doesn't conduct anymore. After that the driving circuit is fully discharge C_{gs} and the gate voltage reaches zero. (Mohan et al. 1995, pp. 583-589.)

The MOSFET driving circuit properties actual current driving ability is also an important thing when calculating MOSFET power losses. The peak of the MOSFET driving current directly affects to parasitic capacitors charging and discharging times. These times affects to current and voltage slopes and also times when power loss is generated. High peak current allows fast switching between ON- and OFF-times which leads to lower power losses. Some driver circuits also have different abilities to sink and to source current, that's why slopes can be differ between turn-on and turn-off procedures, but in Figure 2.12 the procedures are equal. Different MOSFETs have different parasitic capacitances. Naturally, lower capacitance takes a shorter time to charge and discharge and then the MOSFET produces lower power loss.

The MOSFET operation can be divided into four stages; the OFF-time, the turn-on procedure, the ON-time and the turn-off procedure. At the OFF-time there isn't power loss at all, because current through the MOSFET is zero. Turn-on power losses are generated by current rising when there is still voltage across the MOSFET. This power loss can be calculated by integrating the product of current and voltage over the turn-on-time and multiplying the result by the switching frequency. ON-time power loss is easy to calculate with Ohm's law and multiplying the result by the conversion ratio. Turn-off power losses are generated by rising voltage when there still is current through the MOSFET, and the power loss can be calculated in the same way as the turn-on loss.

2.5.2 Diode

Pn-junction diodes have power losses because, as with MOSFETs, a diode switching isn't ideal and therefore voltage across the diode and current through it generates switching and conduction losses. Diode reverse recovery time increases MOSFET power losses because the diode continues to conduct current the period when it is became reverse biased. When the MOSFET is turned off and diode starts to conduct, the voltage over diode rises high which increases diode losses. (Mohan et al. 1995, pp. 535-539.)

Because of the structure of a diode it has a parasitic capacitance, whose charging current is added to the MOSFET current. This current flows through the MOSFET during its turn-on and turn-off procedures and it can increases MOSFET power losses significantly.

Diode speed is a very important thing when optimizing total power losses of the converter, that's why a Schottky diode is a good alternative. Schottky diodes have much faster switching characteristics than the normal diode, and a lower ON-time voltage, that's why Schottky diodes are a good choice for an energy-efficient buck converter, although its parasitic capacitance is relatively high (Mohan et al. 1995, pp. 539-543).

One solution is replace the diode with another MOSFET, this method is called synchronous rectification. This method decreases power losses especially in high current solutions and when the duty cycle is low. The diode voltage drop then affects to high losses, although MOSFET conduction loss depends on the MOSFET r_{DS}. The another MOSFET also needs a driving circuit and there has to be dead-time in the driving sequence to prevent both MOSFETs conducting simultaneously.

2.5.3 Inductor

Inductor losses can be divided into core and winding losses. Core losses consist of eddy current loss and a hysteresis loss. Copper losses are a DC-resistance, a skin effect and a proximity effect. All these phenomena combined generate significant power losses which have to be noticed when designing an energy-efficient converter. A calculation of power losses is complicated and there are equations and variable values for estimating inductor losses. In many solutions this method is good enough. (Eichhorn 2005.)

Eddy current is induced into the core material because the flux is changing. According to the Lentz's law, direction of these eddy currents is such that their induced flux is opposite to the main flux. (Eichhorn 2005.) Hysteresis losses are caused by inductor core material hysteresis. Magnetic domains affect hysteresis, they are turned by the magnetic flux in same direction and when all the domains are turned the core material is saturated. When magnetic flux is removed, domains don't turn back to their original directions and this means that the material has a remanence flux. This remanence flux needs a coercivity field to drive the magnetic flux to zero. This phenomenon means that magnetizing the inductor needs more energy than demagnetizing and their difference is power loss, and it heats the inductor core. (Eichhorn 2005.) The inductor current rises during every main switch ON-time and decreases during every OFF-time, thus the inductor is magnetized every ON-time and demagnetized every OFF-time and hysteresis generates power losses during that operation.

An inductor wiring DC-resistance causes power loss which can be calculated with Ohm's law and with current DC-component. The DC-resistance is determined by the wire cross-sectional area, length and material resistivity. Usually the wire is made copper. The proximity effect increases the wiring AC-resistance and causes power losses. (Eichhorn 2005.)

The skin effect is affected by eddy currents that generate higher current density on the outer surface of the wire. The skin effect can be estimated by using a penetration depth. It describes how deep part of wire outer surface carries current. (Eichhorn 2005.)

The proximity-effect occurs because conductors induce eddy currents to nearby conductors. This eddy current increases the wire AC-resistance because the effective cross-sectional area for current decreases. The proximity-effect increases when the penetration depth is some amount narrower than the wire diameter and the wire is wound that winding layers are on top of each other. Then current in the upper wire layers induce eddy currents to other wire layers which are beneath them. The phenomenon increases in every layer and this increases the AC-resistance of the wiring. (Mohan et al. 1995, pp. 771-773.)

2.5.4 Other

Other losses are generated by the current sense resistor, gate resistors, ICs, capacitors and the supply voltage generation. The amount of these losses depend on the type of components and the design of the circuit. Power loss of resistors can be calculated with Ohm’s law, the current sense resistor can generate high power loss if the voltage across it is moderately high. ICs, like control- and measurement ICs, dissipate energy too, which generates some power losses. Especially gate driving ICs might dissipate large amounts of energy. If the gate switching speed is high, the gate driving energy is high. Capacitors have a parasitic series resistance which generates power losses too. A capacitors power loss depends on current ripple, and the ESR value depends on the frequency of this ripple. ICs may need a lower supply voltage than the circuit input voltage. This supply voltage has to be generated by its own circuit. Power supply generation isn’t ideal and it generates power losses too. Some designs dissipate more energy than others. That’s why power supply generation has to be designed correctly to minimize losses.

2.6 Light Level Control

There is the need to adjust the light level in many installations. A light level control saves energy and meets demands for different light outputs. Different light levels are needed in theaters, working desks and so on. One solution for light level control is a RGB (Red Green Blue) control which can generate different colors by mixing three basic colors. Demand for light levels may be from 1% to 100% or from 0,1% to 100%. When using the RGB, control to mix the different colors is also setting high demands for the light level control. Levels have to be precise and all three channels have to be working identically to achieve stable and precise color output.

There are two different basic dimming methods; PWM (Pulse Width Modulation) and analog dimming. PWM dimming sets the average LED current to a desired value. At the PWM ON-time, the LED current is at its normal level and at the OFF-time it is at zero By changing the duty ratio the average LED current can be adjusted. The analog dimming method is changing the LED current level linearly and this reduces current flow through the LEDs continuously. (Aimtec 2011.)

There is also a third dimming method called hybrid dimming. It is implemented by mixing PWM and analog dimming methods together. Using hybrid dimming gives some advantages compared to pure PWM or pure analog dimming, but it also has some disadvantages. (Beczkowski & Munk-Nielsen 2010.)

Dimming is difficult because it always affects the wavelength of the LED, and at same time the color has to stay stable, especially with RGB control. Hybrid dimming is a solution to this because PWM and analog dimming methods change the LED wavelength in different ways; the PWM method decreases and the analog method increases the wavelength. Using the right combined amount of PWM and analog dimming the wavelength can stay stable. Other difficulties are caused by the demand for step-less dimming. (Beczkowski & Munk-Nielsen 2010.)

2.6.1 PWM Dimming

PWM dimming is the most common LED dimming method. It is simple and doesn't affect the LED wavelength too much. The idea of PWM dimming is to reduce the time-averaged LED current by blocking the LED current flow at the OFF-time of the switching cycle and at the ON-time the current through LED is normal, as shown in Figure 2.13. In Figure 2.13 the blue line is the PWM-signal which controls the PWM dimming, red is the inductor current and black is the time-averaged inductor current. This reduces the time-averaged current level and dims the LED. There are two ways to implement PWM dimming, one is to prevent the main switch operating during the PWM OFF-time, and the another is by using a shunt-MOSFET or series-MOSFET to prevent current flow to the LED load. (Texas Instruments 2011.)

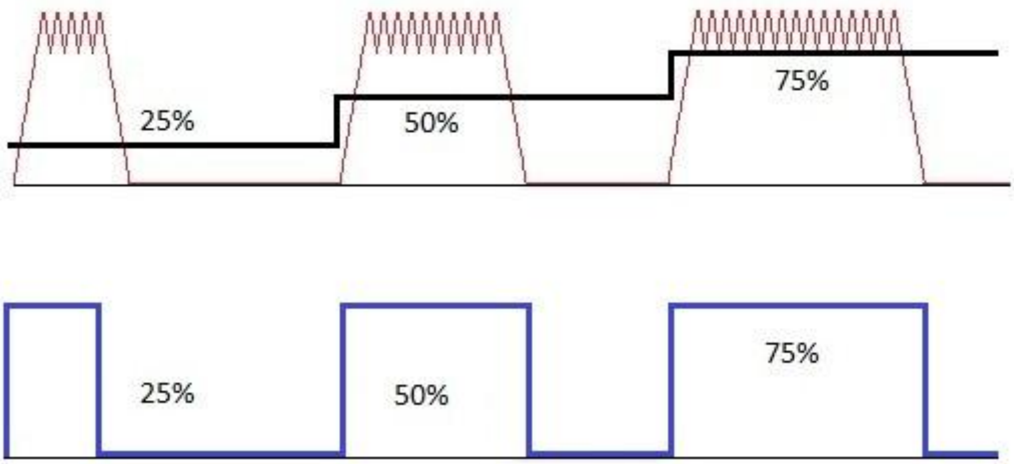

Figure 2.13. *PWM dimming waveforms.*

The PWM dimming frequency is important to be set correctly. It has to be higher than the human eye can detect otherwise the dimming is causing a flickering effect which is disturbing. A too high frequency may cause inaccuracy and other problems at low duty cycles, if at PWM ON-time has too few buck switching cycles. Suitable dimming frequencies depend on the solution properties, but for LED current sources the frequency should be over 200 Hz to avoid flickering and under 1 kHz to avoid inaccuracy. PWM dimming could also produce audible noise during operation if the PWM frequency is in the audio band which is between 20 Hz to 20 kHz (Richardson 2012).

PWM dimming is based on preventing current flow through the LEDs during the PWM OFF-time. Many buck-control ICs have this feature. Inside them is a logic-circuit which prevents the buck main switch from turning on at the PWM-dimming signal OFF-time. Dimming shuts down only the output, and the control IC is running to achieve fast response when the PWM dimming signal goes to a high level and the output current has to rise high too. If the delay is short enough dimming is precise and inaccuracy is minimized. (Texas Instruments 2011.)

The PWM dimming has one big problem; the inductor current rise and fall times which are caused by inductor energy charging and discharging. Inductor energy charging causes inaccuracy because the inductor current rise needs time and during that rise the inductor current is lower than demanded. This reduces the real output current. Inductor energy discharging causes a current tail which can’t cut off. This current flows through the diode and also through the LED string and it increases the LEDs average current during dimming. The amount of these current inaccuracies depends on the inductor current level and the inductor current rising and falling slopes, which depends on input and output voltages and the inductor inductance. In some cases these cause high inaccuracy to the LEDs dimmed current.

Other ways to implement PWM dimming is using a shunt-MOSFET or a series-MOSFET. The shunt-MOSFET is parallel to the LED load and the series-MOSFET is in series with the LED load. The shunt-MOSFET dimming method is based on a dimming MOSFET which short circuits the output at the PWM OFF-time as shown in Figure 2.14. Then the output current flows through the dimming MOSFET and the power loss can be calculated with Ohm's law from the output current and MOSFET r_{DS}. The shunt-MOSFET PWM dimming may cause a huge power loss and an output capacitor can't be used. The output capacitor is prohibited because short circuiting the capacitor causes very high current spike and the MOSFET can't handle that current. (Texas Instruments 2011.)

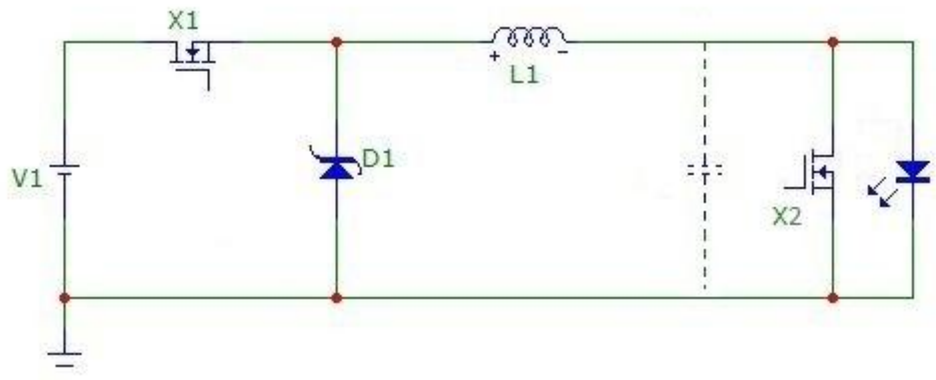

Figure 2.14. Shunt-MOSFET dimming method.

The series-MOSFET PWM dimming based on a series-MOSFET which prevent current flowing to the LED load as shown in Figure 2.15. This method needs an extra diode (D2 in Figure 2.15.) to provide a way for the inductor current to prevent damage to the circuit.

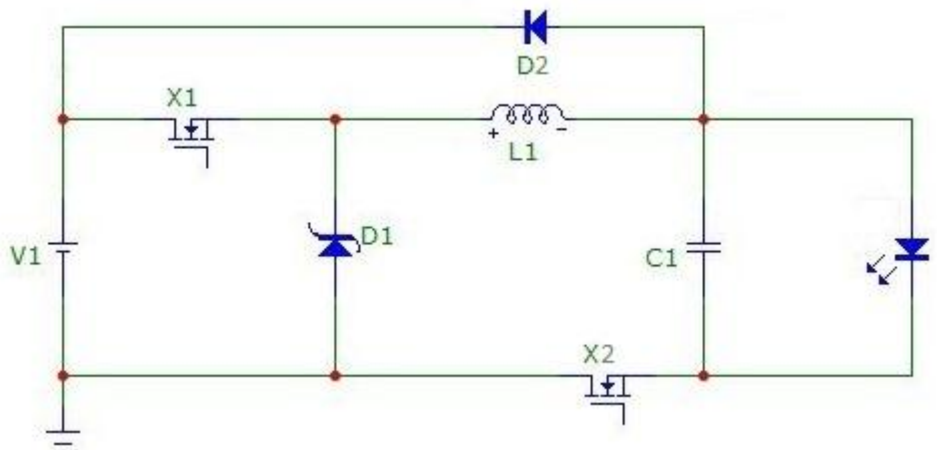

Figure 2.15. Series-MOSFET dimming method.

In some cases PWM dimming methods can be mixed by using the shunt-MOSFET only to cut the inductor current falling time tail off. This allows more accurate dimming because the LED string current flows only at the PWM ON-time, but the output capacitor is prohibited and the inductor current ripple is the same as the LED string current ripple.

Dimming accuracy is important, especially in multi-channel solutions. To add PWM dimming accuracy at short ON-times without using the shunt-MOSFET, the PWM switching frequency can be changed. This means practically an OFF-time control. It is then easier to achieve precise logarithmic dimming levels. The ON-time stays constant and the OFF-time is adjusted and after some steps the ON-time decreases and the OFF-time starts to grow again. This allows more precise dimming without flickering or visible steps.

2.6.2 Analog Dimming

An analog dimming is also called linear dimming. It differs from the PWM dimming in that the LED current is continuous and dimming is based on lowering the current level as shown in Figure 2.16. In Figure 2.16 the red curve is the inductor current and the black line is the time-averaged inductor current. Analog dimming doesn't cause flickering as easily as PWM dimming, but it changes the LED wavelength more and may affect significantly to color. The LED wavelength is inversely proportional to the current through the LED. That's why analog dimming can't be used at lowest light levels, but it is usable in some light output range which depends on the LED properties. (Beczkowski & Munk-Nielsen 2010.) Some LEDs even have a minimum operating current level (Osram 2011).

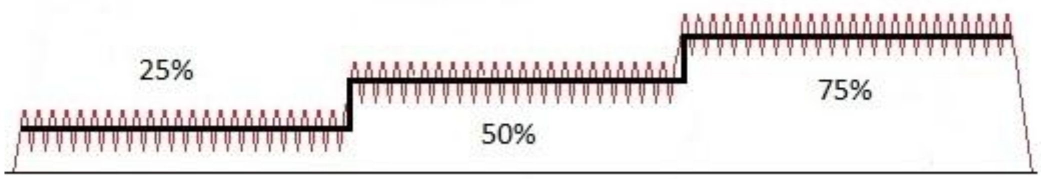

Figure 2.16. *Analog dimming waveforms.*

Analog dimming can be implemented by changing references of comparators. This is the simplest method and many buck control ICs use this method. The second method for implementing analog dimming is by driving extra current through the current sense resistor by using, for example, an operational amplifier. This increases the pin's initial voltage and lowers the LED current which is needed to trigger the comparator to set the mains switch off. Basically both methods are based on the same effect of lowering the comparators' trigger voltages. (Texas Instruments 2011.)

There are three ways to control the inductor current ripple while dimming LEDs analogically; leave the ripple the same as 100% level, decreasing the inductor current ripple in proportion to the LED average current, or try to keep the LED current ripple at the same percentage of the average current the whole time. Acceptable LED current ripple is 5-10% of the average current, and it depends on the LED properties.

When the inductor ripple is the same as the 100% level, the switching frequency doesn't change and the LED current ripple is the same. Then the minimum level is limited because the current ripple has to be lower than the peak current so that the converter stays at CCM. This method can't be used because the LED ripple percentage gets too high at low dimming levels.

When decreasing the inductor current ripple proportional to the average current, the switching frequency is also proportional to the average current. This setting also affects the minimum level, because too high switching frequency generates high switching losses. When current ripple decreases and at the same time the switching frequency increases, the LEDs current ripple decreases faster, because the output capacitor filtering effect is more effective when the switching frequency is higher.

When the LED current ripple percentage of the average LED current is the same at all dimming levels, the connection between the average current and the inductor current ripple isn't linear. The inductor ripple doesn't have to decrease as much as the LEDs ripple decreases because the switching frequency increases and the output capacitor filtering effect gets stronger.

2.6.3 Hybrid Dimming

Hybrid dimming is a method which mixes PWM and analog dimming methods and combines both method's pros together without their cons. Hybrid dimming can be implemented in three ways; first PWM and then analog, first analog and then PWM, or both at the same time. Finally, the dimming curve has to be logarithmic and the minimum level has to be reached. In Figure 2.17 first comes analog dimming 100% to 50% and then PWM, so 50% to 100% can be done by using only the analog method and to reach 25% light output the analog level is then 50% and PWM is 50% too. In Figure 2.17 the red curve is the inductor current and the black line is the time-averaged inductor current.

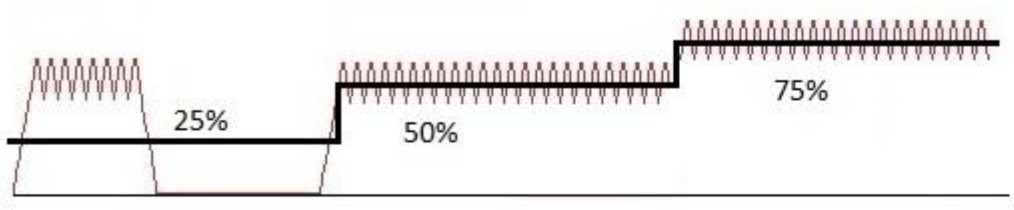

Figure 2.17. *Hybrid dimming waveforms.*

If PWM dimming comes first, then the analog total power loss will be lower because the buck converter is operating only at the PWM ON-time and at the OFF-time it is in stand-by and the power loss is negligible. This method has an inductor current rise and fall time problem when using pure PWM dimming. If analog dimming comes first, then PWM, the inductor current rise and falling time problem is smaller because the inductor peak current is lower when the converter starts to use the PWM dimming method. This method problem is a wavelength change which covers almost the whole dimming range. If using both dimming methods at same time the wavelength change can be minimized.

When comparing hybrid dimming to pure PWM dimming, hybrid dimming reduces flickering and increases accuracy because the inductor current is lower level when the PWM duty ratio is low and that’s why the inductor current rise and fall times are much shorter. Shorter current rise and fall times a significant enhancement because the current rise and fall times are the highest source of inaccuracy when using pure PWM dimming, especially if the output voltage is low. Low output power levels are also easier to be implemented because the narrowest PWM pulse width is much wider than using pure PWM dimming. Then the PWM ON-time is holding more buck converter switching cycles and when the ON-time is widened it doesn’t causes visible steps as easily as pure PWM dimming whose ON-time widening is very narrow.

Lower light output levels can be achieved using hybrid dimming versus pure analog dimming. Pure analog dimming has a minimum level because the converter has to be at CCM or BCM and the switching frequency has to be under the maximum, which depends on the control circuit and main switch properties. Also some LEDs have a minimum operating current, which reduces the analog dimming methods minimum light level (Osram 2011).

Hybrid dimming can also compensate for analog dimming wavelength changes because PWM and analog dimming changes the wavelength in different ways. When mixing these methods with the correct percentages the wavelength can stay exactly same. But different LEDs have different properties and both dimming methods affect their wavelength in different ways. That’s why wavelength can’t stay exactly the same for all LEDs, but it stays more stable than using pure PWM or analog dimming. (Beczkowski & Munk-Nielsen 2010.)

2.7 Control Buses

Dimming also needs control and there are many different dimming control buses which can control dimmable ballasts like LED current sources and fluorescent lamp electronic ballasts. Buses are specified with standards. Buses can be analog or digital. Digital buses provide many other functions than just dimming one ballast, they can control ballast groups and they can have a memory to remember settings. This Section is focused on the digital DALI protocol and an analog voltage signal, because they are used by Helvar. Phase cut dimmers are also described because they are common. There are also many other digital buses like DMX and MIDI. (Simpson 2003.)

2.7.1 DALI

DALI (Digital Addressable Lighting Interface) has been developed, among others, by Helvar, Osram and Philips for standardized lighting control. The main idea is that one control bus can control many ballasts separately. Each device has its own address and commands can be sent to the right ballast by using it. Ballasts can be divided into groups which have one common address and then all ballasts in one group can be controlled at the same time. Ballasts can send answers to requests too, this can be status data like a light level, lamp failure or load data. A DALI-network needs configuration before using it, this consists of addressing, grouping and so on. (Simpson 2003.)

The DALI light control curve is optimized for eye sensitivity and it is logarithmic, increasing to the next level is always the same percentage of a previous step ca. 2,7%. The precise value is shown in equation 2.6, where L_n is a the light level and L_{n+1} is the next light level. This means that at low light output the total light increase between levels is lower than at high light output levels as shown in Figure 2.18 which presents the DALI output curve.

$$L_{n+1} = L_n \cdot 10^{\frac{3}{255}} = L_n \cdot 1{,}027459 \qquad (2.6.)$$

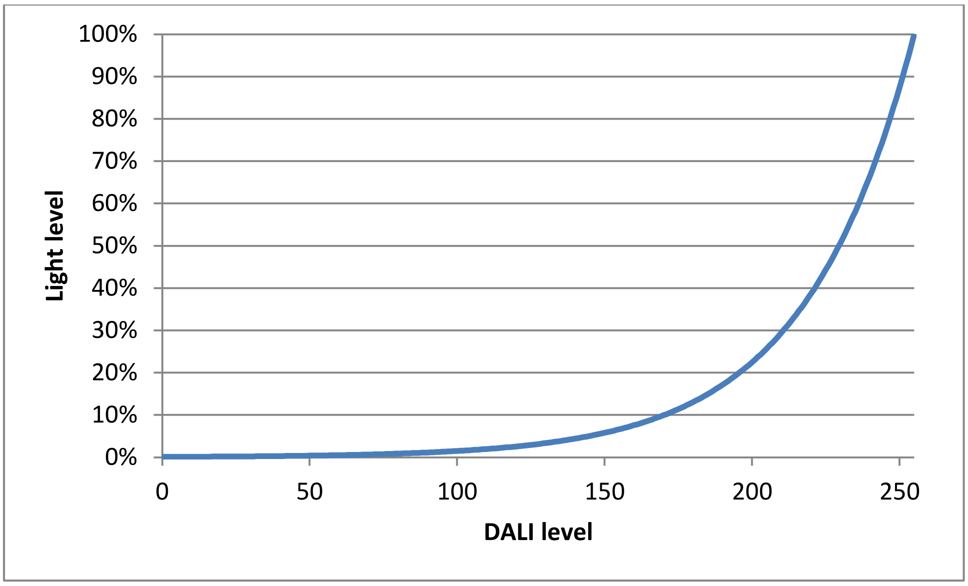

Figure 2.18. *The DALI output curve.*

Properties of DALI are chosen to ensure proper operation in every situation. The signal data rate is low and signal wires can routed with no tight rules: wires can be looped, mixed, and no termination is required. Signal levels have wide tolerances, so the bus is almost immune to noise. The bus has galvanic isolation from the mains voltage. This means that the control signal is safer than without it. Although the control bus is not considered to be SELV, because the bus wiring is too close to the main supply wiring, sometimes wires are run in the same cable shield. The bus uses Manchester coding and the signal DC-level can be blocked and can then be delivered through transformers. (Simpson 2003, pp. 293-295.)

There are many DALI compatible device groups: ballasts, control panels, sensors and gateways. Control panels can be push buttons, sliders, rotary potentiometers, touch screens, remote controllers and so on. Sensors can be used to detect occupancy or light level or both. With gateways DALI-networks can be merged or they can be connected to PCs. DALI is very versatile and it offers a control solution in many installations. (Simpson 2003, pp.295-298.)

2.7.2 Analog

An analog dimming method is very simple. The analog voltage level provides a change to the relative light output level. Naturally all ballasts that are connected to the same analog bus behave in the same way, since all of them receive the same control signal level. There are two ways to implement analog dimming control signal levels: a standardized 1-10 V (IEC60929) whose dimming curve is logarithmic to achieve smoother light output changes for the human eye as seen in Figure 2.19, and 0-10 V whose dimming curve could be linear or logarithmic. (Simpson 2003, pp.280-283.)

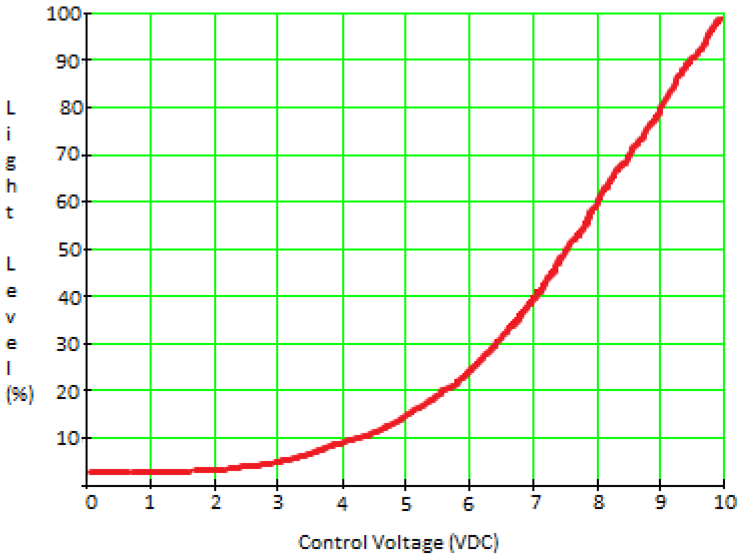

Figure 2.19. *The 1-10 V system dimming curve (Helvar 2012).*

A standardized 1-10 V was developed for discharge lighting. The minimum level is relative and it can vary from 1% to 20% depending on the device and there isn't an OFF-command at all. Whereas a 0-10 V system controls the ballasts to off when the control voltage is something between 0-1 V. A 1-10 V system is more robust than a 0-10 V system because the control line might be noisy, which can cause light level changes at low control signal voltage levels. (Simpson 2003, pp.280-283.)

2.7.3 Phase Cut Dimmers

Phase cut dimmers are a traditional and easy way to dim incandescent light bulbs and that's why they were the most common dimming method earlier. In many installations phase cut dimmers are still used and they can also be used to control LED brightness.

Phase cut dimmers dimming is based on cutting the mains voltage which is reducing the average voltage across the load and when the load is resistive, as an incandescent light bulb, this reduces the output power and the light level linearly in respect to the voltage across the load. There are two different phase cut dimmers: rising (leading) edge dimmers and falling (trailing) edge dimmers. Rising edge cutters are based on a TRIAC or two thyristors. The TRIAC or one of the thyristors is turned on when the mains voltage phase is desired (something between 1 and 179 degrees) and when the load is resistive it conducts until the voltage reaches zero and the phase is 180 degrees. Then the TRIAC or the another thyristor is turned on when the phase is desired (181-359 degrees) and it conducts until the phase reaches 360 degrees when load is resistive and the cycle starts again. Falling edge cutters are based on two transistors, one of them is turned on when the mains voltage is zero and the phase is zero too, and when phase reaches the desired level (1-179 degrees) the transistor is turned off. When the voltage reaches zero level and the phase is 180 degrees the another transistor is turned on, when the phase reaches the desired level (181-359 degrees) the transistor is turned off. When the mains voltage phase reaches 360 degrees the cycle starts again. (iWatt 2012.)

Phase cut dimmers cannot be used directly to dim LEDs because LED current sources need uninterrupted mains voltage. There are special ICs that are designed to handle a phase cut dimmer output voltage and set the dimming level according to that. The control circuit has to measure the mains voltage to detect the dimming level. This information is converted to a normal dimming procedure which dims the LEDs. The control circuit as well as the whole LED current source has to manage situations when the main supply line is zero level at some part of sine wave, usually this means big capacitors which provide supply voltages during main supply cut off. (iWatt 2012.)

3 OPTIMIZING

When optimizing an energy-efficient buck converter many things have to be noticed. The first decision is an alternative, a main switch high-side or low-side. After that is the selection of the main circuit basic components, which affects very much to the efficiency and cost of the whole converter. When the main circuit is chosen, next comes a control method choice and a control circuit design. When the converter and the control circuit works well, the next consideration is designing of a light level control. All sections have to be designed correctly and at the same time other parts have to be kept mind, otherwise results aren't good enough. The aim is to find a compromise which meets the specifications and is the best possible solution.

3.1 Buck Converter Possible Problems

A LED current source designer faces many kinds of problems. Some problems relate to the switched-mode converter itself. This Section describes the most common problems which are related to the buck converter.

3.1.1 Efficiency and Heat Generating

As told in Section 2.5. the converter components generate power losses, which decrease the efficiency and generate heat. The most important components are: a MOSFET, a MOSFET driving circuit, a diode, an inductor and a supply voltage generation circuit. The designer has to find a solution where the total power loss is minimized. This solution is a compromise because increasing efficiency at one place may generate more power losses somewhere else. The chosen frequency is also a significant factor when optimizing buck converter efficiency, because power losses of all components are proportional to the operating frequency, especially the MOSFET and its driving circuit losses.

Supply voltage generation is a very important thing because this circuit can generate high power losses if it isn't designed properly. For example a resistive voltage divider or a resistor and a zener-diode circuit generates high power losses and they generate too much heat and decreases efficiency too much.

3.1.2 Filtering Output Current

Output current filtering is another problem. Is it needed? How high capacitance of the capacitor is enough? The output capacitor reduces output current ripple then the inductor current ripple can be higher when an output capacitor is used. Then an inductor with a lower inductance value can be used and it is also physically smaller. The output capacitor placement in the circuit may affect to the control loop, that's why it is important to place it correctly. A capacitor ESR value changes its filtering ability, because the capacitor current goes through the ESR and it causes a ripple voltage over the ESR. Output capacitor selection depends on the LED strings dynamic resistance and the capacitor's own ESR. (Texas Instruments 2007.)

The selected frequency range is also an important thing when choosing the output capacitor. A capacitors ability to filter current is strongly proportional to frequency; a higher frequency means better filtering and a lower frequency means worse filtering respectively. A capacitor isn't ideal, it has an ESL which means that the capacitor has a resonance frequency and at that frequency the capacitor behaves like a resistor whose value is the ESR value, and above that frequency the capacitor behaves like an inductor and it hasn't the desired filtering effect.

A capacitor tolerates heat badly if it isn't made for high temperatures. Its temperature has to be between specific limits, which depends on the chosen capacitor type and specifications. The capacitor has to tolerate the heat produced by the converter. Different types of capacitors have different temperature limits and other properties, so capacitor type choice is also important.

3.1.3 Output Dynamics Demand

Output dynamics are also an important thing which may cause problems if the dynamics are too wide. Optimizing doesn't give the best results if the output voltage range or output current range are too wide. Then too many compromises have to be done and the converter becomes too expensive and big, because all components have to be chosen to handle the highest output current and the highest input voltage. That's why specifications have to be chosen carefully.

Wide output current dynamics cause a vast operational frequency range. In hysteresis control the frequency is inversely proportional to the output current hysteresis: halving the output current hysteresis doubles the operational frequency. This increases the MOSFET switching losses significantly. Wide output voltage dynamics generate power losses too, especially if the output voltage is much lower than the input voltage, then the diode losses increase when the duty ratio is low.

3.1.4 EMI

A switched-mode converter generates noise to the environment, both radiated and conducted. This noise is at high frequency and it might be powerful. There is also tight limits how much noise electric devices are allowed to emit. Emitted noise could cause other devices malfunction. Conducted noise can be divided two parts: conducted to the input side and conducted to the output side. At the input side, noise could cause problems like heating and input voltage ripple. Therefore there is a need for filtering. (Mohan et al. 1995, pp. 500-502.)

Radiated emissions might also be a big problem. A switched-mode converter emits electric and magnetic fields. Both of them are harmful and must be below the specific limits. An inductor is a significant source of magnetic fields which might also be a problem. Those fields interfere with the converter itself, its control circuit and also other nearby devices. Especially inductors with open magnetic circuit emit magnetic fields because their fields have to go through free air. Inductors with closed cores don’t emit magnetic fields as much because their field goes mainly through the magnetic core. (Mohan et al. 1995, pp. 500-502.)

3.1.5 Inductor Core Saturation

Inductor core saturation is also a problem which has to be noted. The inductor core maximum flux density determined by the core material shall not be exceeded. Magnetic flux through the inductor core is proportional to the current through the inductor winding, if the flux is below maximum flux. This determines the inductor maximum current. If current through the inductor exceeds its maximum current, the core starts to saturate and the relation between inductor current and magnetic flux are not linear anymore. That’s why when the inductor core saturates, the inductor current starts to grow faster because the inductor inductance is reducing and is now dependent on the inductor current. The inductor has to be designed not to saturate in normal operating conditions. That’s why there has to be a margin between the inductor saturation current and converter maximum peak current.

3.1.6 Price

The converter total price is also a significant thing. All earlier choices affect to the price. The goal is to make a cost-efficient and energy-efficient converter. As said before, the converter design is optimized and includes many compromises. The result has to be cost-efficient enough and good enough and it has to meet the specifications.

3.1.7 Physical Dimensions

The converter physical dimensions are also very important because a smaller converter makes it possible to use a smaller LED current source case. A smaller LED current source is easier to mount because usually it is inside the luminaire and it is easier to find a place to fit it. A small size is also saving costs. The PCB is smaller as well as the case costs. When an LED current source is small, logistical costs are also lower because the needed space for storing and transportation are lower.

3.2 Optimizing Buck Converter

The aim is to design a cost-efficient and energy-efficient buck converter whose output power is 30 W and whose output currents are 700 mA and 350 mA. The output voltage dynamics aren't too wide: at 700 mA it is 30-42 V and at 350 mA it is 40-60 V. Multi-channel compatibility is demanded too, as well as small size.

3.2.1 Topology and Main Circuit

In the optimized buck converter the main switch and the inductor are low-side, because then multichannel LED current sources are easier and more cost-efficient to implement. Also MOSFET control is easier to implement, because no high voltages are needed. Other components are in their own places as shown in Figure 3.1. Input voltage is 70 V and all components have to handle that with a safety margin, at least 100 V is a good maximum voltage for components. Maximum output current is 700 mA, 1 A current is suitable when choosing components. Output current ripple can be up to 10% of the LED current and the operational frequency range is 100-400 kHz. All components have to handle these demands, and the whole converter too.

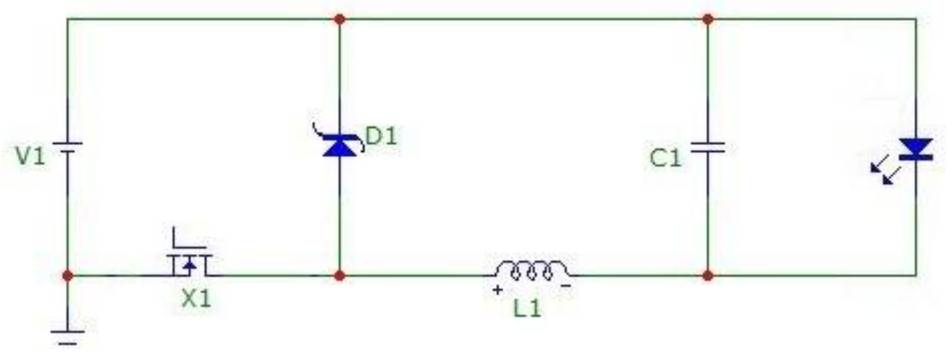

Figure 3.1. *The optimized buck alternative.*

The alternative used is low-side, since then all channels can use the same positive voltage cord as shown in Figure 3.2. This saves space on the PCB and connector costs. The converter in Figure 3.2. could be an RGB-light LED current source, which is mixing different colors from red, green and blue.

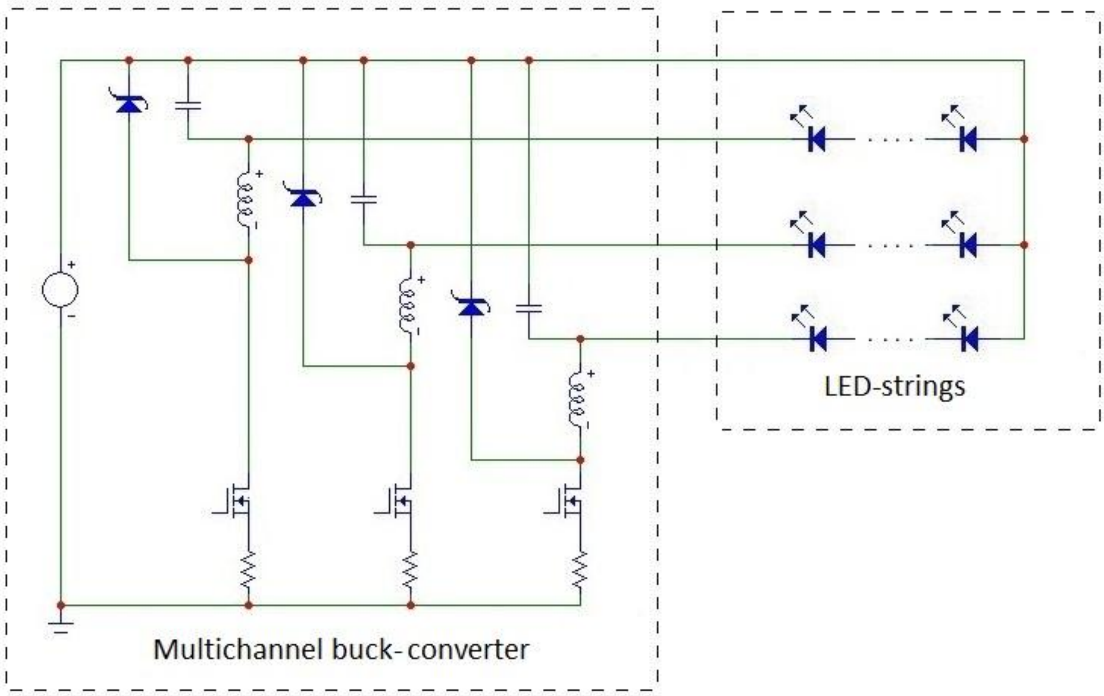

Figure 3.2. *A multichannel LED current source.*

The components of the main circuit have to be chosen next. The chosen components affect very much to the converter properties and how it meets specifications.

3.2.2 Control Method

The best way to control a buck converter is the hysteresis control method, then there isn't need for slope compensation at all, the converter is stable without it. The chosen low-side buck alternative does not allow the use of hysteresis control without a complex and expensive current measuring circuit, thus the hysteresis control cannot be used. The second-best way to control a floating buck converter is the constant OFF-time control method. It can be used with the chosen topology and it does not need slope compensation either. This method only measures the rising inductor current and this can be done by a cost-efficient current sense resistor and a comparator with reference voltage.

Current hysteresis depends on a constant OFF-time. The OFF-time has to be controllable during operation if analog dimming is used or output voltage is changing.

3.2.3 MOSFET and Driving IC

A selection of the main switch isn't easy, because many things have to be noticed; maximum voltage and current, total gate charge, resistance between drain and source, package size and maximum temperature. After maximum voltage and current, the most important factor is the total gate charge when choosing a MOSFET, because the switching frequency is relatively high. Switching losses can be significant if switching is too slow. Also the r_{DS}-resistance and the package size are important specifications when choosing the MOSFET. Main priority is to find at least 100 V and 1 A MOSFET with a very low gate charge to ensure quick enough switching. A suitable package size is DPAK which could dissipate the generated heat.

A MOSFET driving IC has to have good current driving ability to ensure fast switching. The IC also has to be fast enough to reach the desired operational frequency. Also control would be good to be done with this IC. Suitable IC would be a constant OFF-time buck-control IC.

3.2.4 Inductor

Inductor properties that have to be optimized are losses, size, saturation and inductance. First the desired inductance has to be calculated, which depends on the operational frequency and the desired hysteresis. Next it has to be decided what kind of inductor is the best for that solution: core type, material and wire type. A closed core type is better than unclosed because then less wire turns are needed to achieve the same inductance. The inductor DC-resistance is lower and the magnetic field doesn't interfere with the converter itself or other nearby devices. The core has to be, for example, E and I, E and E –shaped, or a pot core to achieve that. The best core material for these operational frequencies is ferrite. The inductor maximum current has to be over 1 A to prevent its core saturating.

The wire type depends on how much space is available for wiring. All this space would be used to optimize copper losses. The thickest possible copper wire has to be used to fill all space and minimize copper wire resistance. It is also possible to use a multi-strand wire to minimize the skin-effect and the proximity effect.

3.2.5 Diode

The diode maximum voltage has to be over 100 V and its maximum current over 1 A to prevent failures. The diode used has to be as fast as possible to prevent switching losses. The parasitic capacitance has to be minimized to prevent too high MOSFET current during switching. For speed a Schottky-diode is a good choice, although its capacitance is relatively high, but it reduces the converter total losses. A Schottky-diode also has a lower forward voltage drop than regular pn-juntion diodes, which decreases the diode losses especially if the output current is relatively high.

The diode features include reverse leakage current which discharges the output capacitor during the PWM dimming OFF-time. This voltage drop affects inaccuracy for the dimmed output current because, when the ON-time starts, the capacitor has to be charged before current flows to the LED load. Schottky-diodes have high reverse leakage current compared to other diodes.

3.2.6 Output Capacitor

An output capacitor value calculation depends on the inductor current ripple, operational frequency and the desired output current ripple. Capacitor value selection according to calculation is important because if it has too high capacitance it costs too much and takes too much space. If it has too low capacitance it doesn't filter the output current enough and it may cause too high ripple which may damage the LED load. The capacitor has to be selected for the worst case situation, when the operating frequency is the lowest, the inductor ripple is the highest and the allowable LED current ripple is the lowest. A well-chosen capacitor reduces space on the PCB because a smaller inductor can be used to achieve the same LED current ripple.

3.3 Light Level Control Possible Problems

Light level control also causes some problems, part of them are new and some of them are same as the buck converter problems. Some of them are very serious, like visible dimming steps and flickering.

3.3.1 Visible Dimming Steps

The biggest problem in light level control is visible dimming steps, especially in low light output levels and when the output current peak level is high and the output voltage is low. Steps are disturbing and they aren't acceptable in high quality lighting devices. Steps are caused by too low buck frequency compared to PWM frequency and inductor current rising and falling times. That's why some dimming levels don't affect the output current and effective levels affect a high light output change between them. Basically these steps are caused by the inductor energy charging at the beginning of the PWM ON-time and the discharging through the diode which can't be stopped. It flows through the LEDs if the circuit does not include a shunt-MOSFET or series-MOSFET. Effective dimming levels are located only at the main switch ON-time, because only then the dimming control can affect the buck converter operation. Also the dimming level after the buck ON-time is effective because it determines the last buck cycle which affects to the output current.

Especially when the output voltage is low compared to the input voltage, the main switch OFF-time is long compared to the ON-time, and then the current goes through the diode for almost the whole time of one buck cycle. This will strengthen the visible steps between effective dimming levels because it reduces the amount of them. If during the main switch OFF-time there are many dimming levels, the amount of effective dimming levels are reduced as shown in Figure 3.2. Figure 3.2. shows the inductor current and PWM dimming levels, the output voltage is low and then the OFF-time is much longer than the ON-time. The buck converter switching cycles are numbered 1-4 as well as PWM dimming levels 1-16. As said, PWM dimming control can only change the output current at the buck converter ON-time, now many PWM dimming levels are located during the buck converter OFF-time. These levels cannot affect the output current, they are useless.

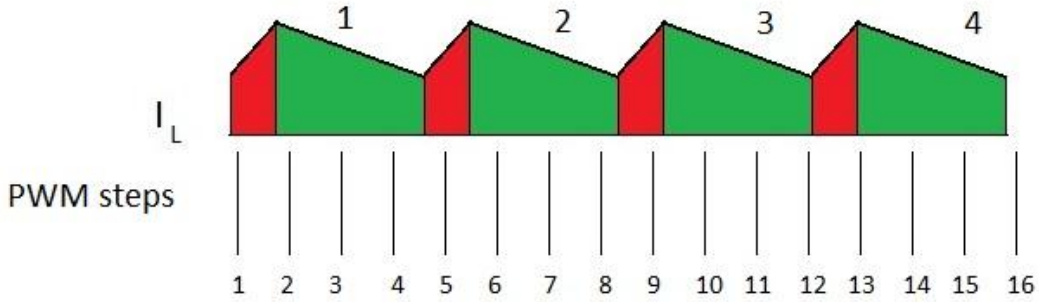

Figure 3.2. *PWM dimming levels.*

Let's calculate how many dimming levels are usable and how many are useless. During the first buck cycle dimming levels #1 and #2 can affect to the output current. In the second buck cycle levels #5 and #6 are effective. In the third buck cycle levels #9 and #10 are counted, but in the 4th buck cycle there isn't effective dimming levels at the ON-time at all. Level #13 is only effective then. Seven levels are usable and the rest is useless. This means that over 50% of levels are useless. This effect is visualized in Figure 3.3. On the left the PWM dimming signal goes down at level #6 and the inductor current is shown above. To the right the PWM dimming signal goes down at level #8 and the inductor current is shown above too. As shown, the inductor current waveforms are identical, and so are the output current waveforms.

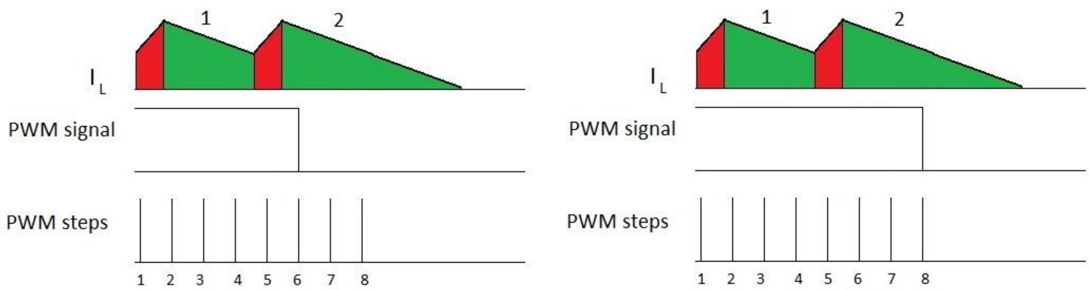

Figure 3.3. *Useless dimming levels.*

So in this example two dimming levels per any buck cycle are useless. This is a significant problem if the buck converter frequency is low and output voltage is low too. Accuracy cannot be added by increasing dimming levels, this increases useless levels much more than usable levels because the main switch ON-time is narrow compared to the OFF-time. Then the cost increases much more than real dimming accuracy.

3.3.2 Flickering and Stroboscopic effect

PWM dimming may cause flickering if the PWM frequency is low enough that a human eye can detect switching. Flickering is very disturbing and may cause many harmful effects to humans, like headache. Flickering isn't acceptable when a high quality LED current source is dimming LEDs. That's why the PWM frequency has to be higher than the human eye can detect.

PWM dimming may also cause stroboscopic effect, which is caused by a light source that is continuous switching ON and OFF like a stroboscopic light. The stroboscopic effect can be dangerous if a rotating object seems to be stopped. This phenomenon occurs when the PWM frequency and the rotating speed of the object are equal. If the PWM frequency is changing during dimming the stroboscopic effect could occur more easily than during constant frequency PWM dimming, because then there is a bigger chance that the PWM frequency is matched to the rotating frequency.

3.3.3 Input Voltage Ripple

Earlier in this thesis the buck converter input voltage is assumed as ripple free DC-voltage. In a real solution the input voltage is generated from the AC-supply, which causes low frequency ripple. The amount of that ripple is dependent on earlier stages, like the PFC, half bridge or fly-back.

An input voltage ripple changes the buck input voltage which affects the inductor current rising slope. This may cause visible light level changes because the length of buck cycles are changing even if the user does not adjust light level. When the length of the buck cycle is changing, this might change the number of buck switching cycles during the PWM dimming ON-time, which changes the buck converter output current. Those changes aren't acceptable in high quality lighting devices and they are disturbing too.

The input voltage ripple depends on the buck converter load, the capacitor filtering ability and input stage dynamics. When the buck converter input current is high, then the input capacitor load is high, and the input voltage ripple is high too. The amount of voltage ripple effects proportionally to inductor current slopes. Higher ripple means the higher the slope changes.

In multichannel devices this phenomenon is the strongest when one channels light output is low and other channels light output is maximum. One channels input current is at the maximum level, the input capacitor load is high, and the input voltage ripple is high too. Because of that, the channel with the low output level may have visible light level changes if the input voltage ripple changes the inductor current slopes too much.

3.3.4 Color Change

A color change is also a problem when controlling an LEDs brightness. Both usable dimming methods change an LED wavelength and at the same time the wavelength has to be stay stable. Even hybrid dimming changes the wavelength because the LED properties vary and the optimal dimming methods mixing curve differ for different LEDs. Especially for RGB-mixing solutions, the wavelength has to be constant to achieve the correct output colors.

3.3.5 Output Dynamics Demand

Output dynamics demands cause problems also with dimming. If the output voltage is too high the inductor current rising slope is very low. This causes long inductor energy charging and causes inaccuracy to dimming as shown in Figure 3.4. In Figure 3.4 the lowest curve is the PWM dimming signal and the top curve is the inductor current waveform when the output voltage is high. T_1 is the error caused by the inductor energy charging, and T_2 is the inductor energy discharging error. As seen, when the output voltage is relatively high, the biggest problem is the inductor energy charging.

If the output voltage is too low the inductor current falling slope is very low and the main switch OFF-time become very long compared to the ON-time. This causes visible dimming steps as told in Subsection 3.3.1. A low inductor current falling slope also causes a huge current tail when the PWM-signal goes down and the inductor current starts to decrease, as seen in Figure 3.4. In Figure 3.4 the middle curve is the inductor current when the output voltage is low, $T_{1'}$ is the current rising error, and $T_{2'}$ is the current tail error. The long inductor current tail causes inaccuracy during dimming, because the inductor current flows through the LEDs when the PWM-signal is down and the control system assumes that the current through the LEDs is zero. This problem is bigger when the output voltage is significantly lower than the input voltage. This problem derives from too wide output voltage dynamics.

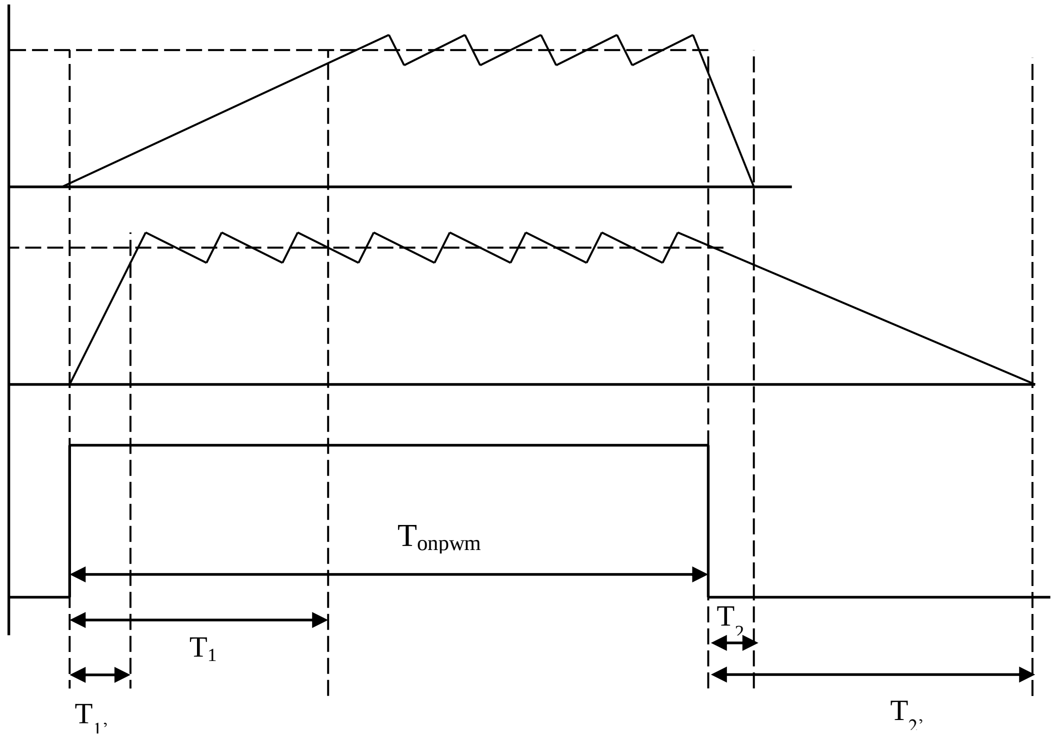

Figure 3.4. *Inductor current inaccuracies, when output voltage is high (the top curve) and when output voltage is low (the middle curve), the lowest curve is the PWM dimming signal.*

The amount of the error caused by the inductor energy charging and discharging can be calculated by using the following equations: in equation 3.1 the inductor current rise time is calculated by using the input and output voltages, the inductance of the inductor and the nominal inductor current level.

$$U_{in} - U_{out} = L \cdot \frac{I_{nom}}{T_1} \rightarrow T_1 = \frac{L \cdot I_{nom}}{U_{in} - U_{out}} \quad (3.1.)$$

Equation 3.2 shows the inductor current fall time, and it can be calculated by using the output voltage, inductor inductance and the nominal inductor current level.

$$U_{out} = L \cdot \frac{I_{nom}}{T_2} \rightarrow T_2 = \frac{L \cdot I_{nom}}{U_{out}} \quad (3.2.)$$

The PWM dimming aims to get the inductor current demanded and it is shown in equation 3.3.

$$I_{target} = \frac{T_{onpwm}}{T_{pwm}} \cdot I_{nom} \quad (3.3.)$$

The real inductor current is shown in equation 3.4 and in equation 3.5 it is simplified. As seen, the inductor current rise time decreases, and fall time increases, real current.

$$I_{real} = \frac{T_{onpwm}-T_1}{T_{pwm}} \cdot I_{nom} + \frac{1}{2} \cdot \frac{T_1}{T_{pwm}} \cdot I_{nom} + \frac{1}{2} \cdot \frac{T_2}{T_{pwm}} \cdot I_{nom} \tag{3.4.}$$

$$I_{real} = \frac{I_{nom}}{T_{pwm}} \cdot \left(T_{onpwm} - \frac{1}{2} \cdot T_1 + \frac{1}{2} \cdot T_2\right) \tag{3.5.}$$

The inductor current is demanded when T_1 equals T_2 and this means that the output voltage has to be half of the input voltage, as shown in equation 3.6. Then the inductor current rise and fall slopes are equal, and their errors cancel each other.

$$T_1 = T_2 \rightarrow \frac{L \cdot I_{nom}}{U_{in}-U_{out}} = \frac{L \cdot I_{nom}}{U_{out}} \rightarrow U_{out} = \frac{1}{2} \cdot U_{in} \tag{3.6.}$$

With equation 3.7 the error compared to the target can be calculated.

$$\frac{1}{2} \cdot \frac{T_2-T_1}{T_{onpwm}} = \frac{1}{2} \cdot \frac{\frac{L \cdot I_{nom}}{U_{out}} - \frac{L \cdot I_{nom}}{U_{in}-U_{out}}}{T_{onpwm}} = \frac{1}{2} \cdot \frac{L \cdot I_{nom}}{T_{onpwm}} \cdot \left(\frac{1}{U_{out}} - \frac{1}{U_{in}-U_{out}}\right) \tag{3.7.}$$

Let’s calculate two examples: input voltage is 70 V, output voltage 1 is 60 V and output voltage 2 is 10 V, nominal current is 700 mA and the inductance of the inductor is 600 µH. Then in equation 3.8 E_1 is the error when using output voltage 1, and in equation 3.9 E_2 is the error when using output voltage 2.

$$E_1 = \frac{1}{T_{onpwm}} \cdot (-17{,}5\mu s) \tag{3.8.}$$

$$E_2 = \frac{1}{T_{onpwm}} \cdot (+17{,}5\mu s) \tag{3.9.}$$

In equations 3.8 and 3.9 T_{onpwm} is dependent on the frequency of PWM signal if the frequency is 200 Hz which means period of 5 ms. If the minimum level is 1%, T_{onpwm} is 50 µs, E_1 is -35% and E_2 is +35%. This means that when the output voltage is 60 V the real output current is 0,65% of nominal, and when the output voltage is 10 V the output current is 1,35% of nominal and the demanded output current is 1% of nominal. This error caused by output dynamics demands can be called static error.

3.3.6 LED Leakage Current

A low current flows through the LED chip before the threshold voltage is exceeded. This current discharges the output capacitor at the PWM signal OFF-time and the voltage of the capacitor decreases. Dimming control assumes that the capacitor voltage is constant during the PWM-signal OFF-time and when the inductor current starts to flow all that current goes to the LED load. However, at the beginning of the PWM-signal ON-time all of the inductor current is charging the output capacitor and no current flows through the LED load. This phenomenon may be the big source of inaccuracy and the only method to prevent that is using a series-MOSFET when dimming LEDs by using PWM dimming.

3.3.7 Audible Noise During PWM Dimming

PWM dimming may also cause audible noise during operation if the PWM frequency is in the audio band which is between 20 Hz to 20 kHz. Two component groups could be the source of this noise; magnetic components and multi-layer ceramic capacitors. The magnetic component core swells and shrinks when the magnetic field in it changes, changing the current and causing magnetic component vibration. When the core vibrates it mechanically touches the windings, shielding or the PCB, and this makes the noise. In a buck converter an inductor might be a significant source of audible noise, especially if the inductor structure is too loose so as the core could vibrate easily. Capacitor noise comes from the piezoacoustic effect when AC current flows through it. Higher current generates higher noise. A buck converter input capacitor faces the highest AC-current and it might be a significant source of audible noise if it is a multi-layer ceramic capacitor. (Richardson 2012.)

3.4 Optimizing Light Level Control

The aim is to design light level control from 100% to 1%. The dimming shouldn’t have visible steps or flickering, and the dimming curve has to be logarithmic to make it appear linear for the human eye.

3.4.1 Visible Dimming Steps

There are many solutions for visible dimming steps; using a shunt-MOSFET or a series-MOSFET, measure how much energy is delivered to the LED string and then calculate the PWM-signal OFF-time length, calculate the PWM-steps only at the main switch ON-time, or by using hybrid dimming. All of these solutions have pros and cons. If using some specific dimming method the output dynamics can be widened because dimming accuracy increases compared to pure PWM dimming.

A shunt-MOSFET is a good way to cut off the inductor current tail, then it doesn’t causes inaccuracy. Current flowing through the LEDs stops when the PWM-signal goes down because the shunt-MOSFET starts to conduct. With the shunt-MOSFET an output capacitor can’t be used because, as told in Subsection 2.6.1., the peak current through the shunt-MOSFET could be too high. The second advantage of the shunt-MOSFET is at really low light output levels. When the minimum light output level should be between 1% and 0.1% the shunt-MOSFET method is probably the only way to achieve the right current through the LEDs, because difference between light output levels are so tiny. At those light output levels it is almost impossible to achieve the right current levels with normal dimming methods.

A series-MOSFET is also a good way to increase the dimming accuracy, because then only the demanded current is flowing through the LEDs. This dimming method eliminates static error which is caused by the inductor energy discharging. The series-MOSFET method also eliminates LED leakage current which can cause inaccuracy if the leakage current is high, as told in Subsection 3.3.6.

The second way is use feedback, measuring the energy delivered to the LEDs at the PWM ON-time and then calculating the needed PWM-signal OFF-time. This method is complicated and it needs a powerful microcontroller to calculate the needed OFF-time, but if microcontroller software works correctly this method gives good dimming accuracy and steps between different light levels are minimized, because by using the OFF-time control a logarithmic dimming curve is easier to achieve. This method uses a variable PWM-signal frequency. This dimming method is also suitable for converters which have high output voltage dynamics, because energy feedback eliminates the influence of long inductor energy charging and discharging.

A PWM-step calculation only at the main switch ON-time removes useless dimming steps,. All PWM-steps could adjust the inductor current and also every step could change the current through the LEDs. As told in Subsection 3.3.1. many dimming steps might be useless and then there are significant light level changes between usable steps. As shown in Figure 3.5. all PWM-steps aren’t useless if PWM-steps are located only at the main switch ON-time. This kind of dimming control is also complicated and needs a high speed microcontroller to achieve a fast enough PWM-step calculation. When this kind of control is working correctly dimming accuracy is good and the human eye cannot detect steps between light output levels. This method is suitable for converters with high output voltage dynamics as energy feedback.

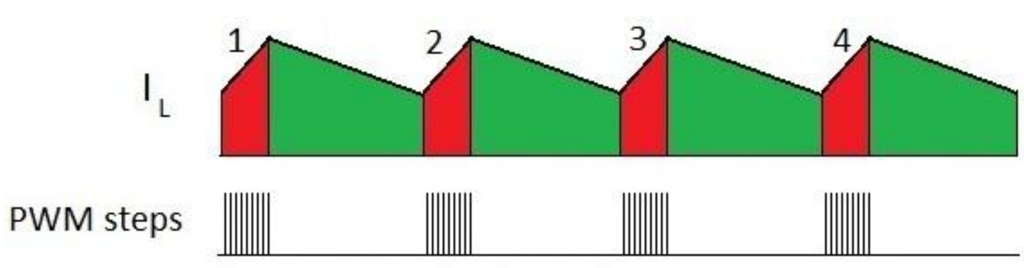

Figure 3.5. *PWM-steps only at the main switch ON-time.*

One way is also use hybrid dimming. When hybrid dimming is used the narrowest PWM-signal level is much wider than using pure PWM-dimming and the absolute increase of PWM-signal length is longer. Also, useless PWM-step counts are reduced when the buck converter frequency is higher if the analog dimming reduces hysteresis. These things will increase dimming accuracy significantly. Also lower inductor current peak values decrease the inductor energy charging and discharging times, and their influence to the inductor current are reduced many times lower. Especially if the output voltage is relatively low or high, the inductor current rising and falling times reduction is significant and it also increases dimming accuracy much more, as shown in Figure 3.6.

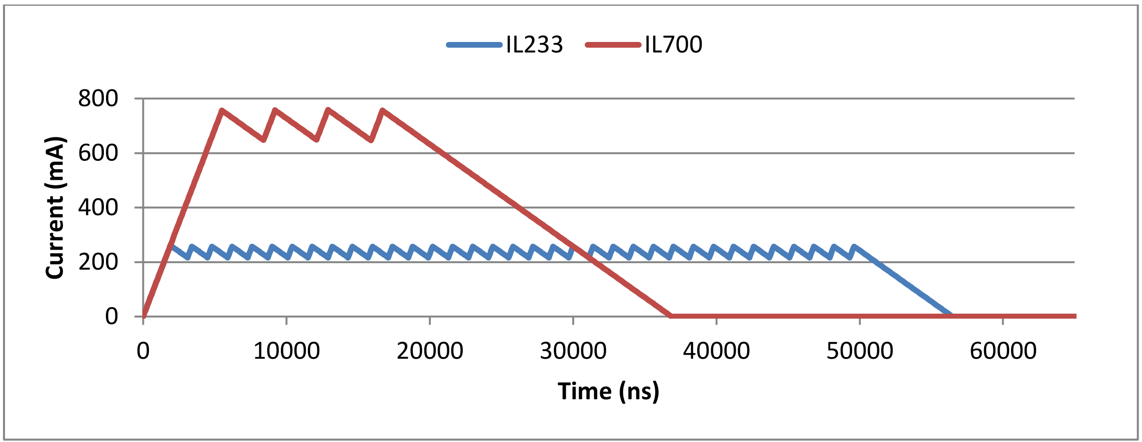

Figure 3.6. *Inductor current rising and falling times.*

In Figure 3.6. the red curve is the inductor current when pure PWM dimming is used. PWM-signal frequency is 200 Hz and light output level is 1 %, input voltage is 70 V and output voltage is 15 V. The blue curve is the inductor current when using hybrid dimming. Hybrid dimming level is one third and then the PWM-signal length is three times wider than when using pure PWM dimming. Input and output voltages are the same as with the red curve. The desired inductor current can be calculated by using equation 3.3 from paragraph 3.3.5:

$$I_{target} = \frac{T_{onpwm}}{T_{pwm}} \cdot I_{nom} = \frac{0{,}5\ ms}{50\ ms} \cdot 700\ mA = 7\ mA \qquad (3.10.)$$

According to this equation the desired current is 7 mA. The difference between the target and real inductor current when using pure PWM dimming can be calculated by using equation 3.7 from paragraph 3.3.5:

$$E_1 = \frac{1}{2} \cdot \frac{L \cdot I_{nom}}{T_{onpwm}} \cdot \left(\frac{1}{U_{out}} - \frac{1}{U_{in} - U_{out}}\right)$$
$$= \frac{1}{2} \cdot \frac{400\ \mu H \cdot 700\ mA}{0{,}5\ ms} \cdot \left(\frac{1}{15\ V} - \frac{1}{70\ V - 15\ V}\right) = 0{,}135757 \qquad (3.11.)$$

The result is about 13,6%. When using hybrid dimming where the analog dimming level is one third of the peak inductor current, then the PWM signal is then three times longer. Now current error can be calculated in the same way as earlier and the error is one ninth compared to pure PWM dimming. The error is 0,015084, which is about 1,5%.

Table 3.1. *The error between desired and real currents.*

	Error percent-age	Absolute cur-rent
Desired current	0 %	7 mA
Pure PWM	13,60 %	7,95 mA
Hybrid	1,50 %	7,11 mA

13,60% error is too much and that's why the minimum output voltage has to be limited if a pure PWM dimming is used. With hybrid dimming the error is much lower and it gives much more accurate inductor current values. Accuracy is dependent on the analog dimming level. Both dimming methods have a static error if the output voltage is something else than half of the input voltage.

3.4.2 Flickering and Stroboscopic effect

As told in Subsection 3.3.2., the only way to prevent flickering when using PWM dimming is to use a high enough PWM frequency. The frequency has to be faster that the human eye can detect. A suitable frequency is over 200 Hz. Above that frequency flickering isn't a problem (Richardson 2012).

To prevent stroboscopic effect one way is use a PWM frequency, or its multiples, which are not located nearby frequently used rotating speeds. Frequently used rotating speeds are 1000, 1500 and 3000 in countries whose mains frequency is 50 Hz. and 1200, 1800 and 3600 where the mains frequency is 60 Hz.

3.4.3 Input Voltage Ripple

A buck converter input voltage ripple has to be under a specific level to prevent flickering and light level changes. The LED current source AC-DC-converter line and load regulation has to be good enough to keep its output steady. The output voltage has to be steady at high loads as well as low loads. The speed of the control has to be fast enough to prevent too high voltages when the load changes to lower, and prevent too low voltages when the load changes to higher.

Input capacitor selection is important when designing a switched-mode converter because input current ripple can be very high and the load changes too. The input capacitor lowers the dynamics demands for the input stage voltage source because the capacitor smooth's the input current ripple and voltage changes during load change situations.

3.4.4 Audible Noise During PWM Dimming

Since the input and output capacitors as well as the inductor may generate audible noise, the selection of them is important when trying to reduce that noise. Suitable capacitors for a switched-mode converter are multi-layer ceramic capacitors or film capacitors, because their ESR values and lifetimes are enough for reliable converters. Richardson's study reveals that film capacitors are more quiet than multi-layer ceramic capacitors and that's why film capacitors are a better choice for a switched-mode converter. (Richardson 2012.)

The Richardson study also reveals that a kool mu toroid and powdered iron core inductors are the best choice for the switched-mode converter inductor. If a ferrite core has to be used then film capacitors are good choice with it. (Richardson 2012.)

A suitable dimming frequency should be as close to 200 Hz as possible, because it is the hardest PWM dimming frequency to hear with an unaided ear. 200 Hz is also the theoretical limit above which the human eye cannot see flickering. 200 Hz is a suitable dimming frequency. While 1 kHz is easy to hear because the human ear is sensitive at a range of 1-5 kHz. (Richardson 2012.)

4 SIMULATIONS

Simulations are an important part of an electric circuit design. They save costs and time, they are safe and usually they give results which are good enough. A modern simulation program allows very complicated simulations. Simulated circuits can be very vast and simulation accuracy can be high and the time step can be very short. However simulations are only simulations and their real accuracy depends on the used component models and used simulation algorithms. They are only estimates of the real behavior of the circuit and they can be used to confirm new circuit behavior. This saves costs and time because simulations can reveal circuit malfunctions before making an expensive prototype.

A simulation program also allows fast circuit, component value or component type changes. They also can draw voltage, current, power and other waveforms which can be difficult to measure from a prototype device. Simulations are also safe because simulation models don't light up, explode or give an electric shock. Different experiments can be done safely with simulation programs.

Simulations are a good tool to estimate a devices power losses and efficiencies, especially MOSFET power losses are easier to simulate than to measure. Power loss simulations need very precise component models. Usually component manufacturers offer component models for their components, but even they aren't perfect. Before simulations models have to be checked and results have to be checked critically too.

4.1 The Simulation Model

At first the simulation model has to be built. Component properties have to be the same as for the real components as well as possible. Simulation model components are editable. All component parameters can be edited to match real life components whose properties are listed in datasheets.

Figure 4.1. shows the used simulation model whose components match well enough to the real components. The MOSFET model is downloaded from the manufacturer and the parameters are checked from the datasheet to achieve proper behavior. The MOSFET driving circuit is modeled by using a voltage controlled voltage source and a resistor, which slows switching down and models a real MOSFET driving circuit. The Schottky-diode model is modified to match the needed diode, especially its parasitic capacitance, which is proportional to the voltage over the diode. The inductor is also

replaced by an inductor model, which includes the inductor parasitic components and which models the inductor power losses too. The output filter capacitor ESR is also included to the simulations. The effect of ESL is tiny at the used frequencies so that it could not be noticed.

The control circuit is modeled by using two comparators with different voltage references to achieve a hysteresis. The comparators are connected to the SR-flip-flop set and reset inputs. This flip-flop controls the MOSFET driving circuit as is described in Subsection 2.3.1. The SR-flip-flop model is ideal and very simple except the delay-blocks which are editable by user. The used comparators are modeled to be realistic.

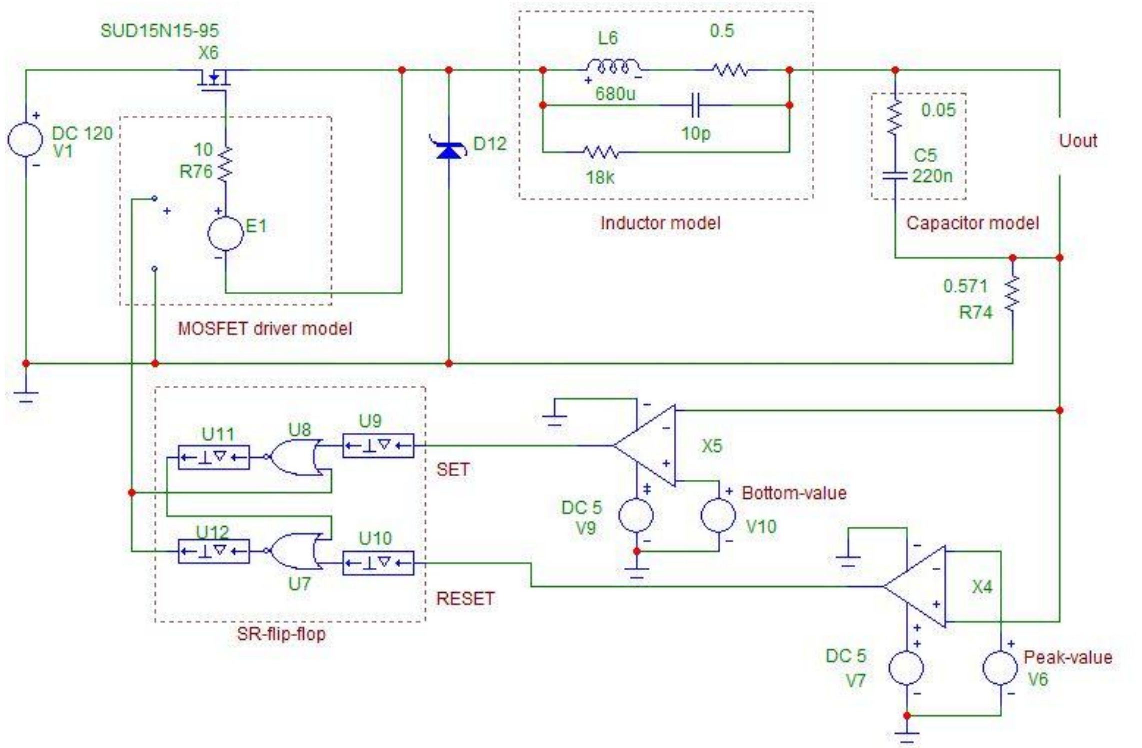

Figure 4.1. *The simulation model.*

The main circuit operation is confirmed and there isn't a need to simulate that. Instead the power losses and output current ripple are important things which have to be simulated before making the prototype. Also, when simulating efficiency and output current ripple, the used alternative isn't important and a high-side MOSFET-circuit can be used although the optimized device uses a low-side MOSFET-circuit. The converter efficiency simulations are done by using a fixed input voltage to the buck converter as well as using variable input voltage and also how power losses change when the converter operates at the BCM. The results are compared to each other and how power losses are divided into the components.

4.2 Fixed Input Voltage Simulations

When the constant hysteresis buck converter input voltage is fixed, the operating frequency changes in respect to the output voltage. The highest frequency is achieved when the output voltage is one half of the input voltage. Below that point the frequency is decreasing, and above that point it is decreasing too. The frequency curve is polynomial as seen in Figure 4.2. It can also be seen how the frequency is doubled when the output current is cut in half. The input voltage is 120 V and the inductor inductance is 1360 µH.

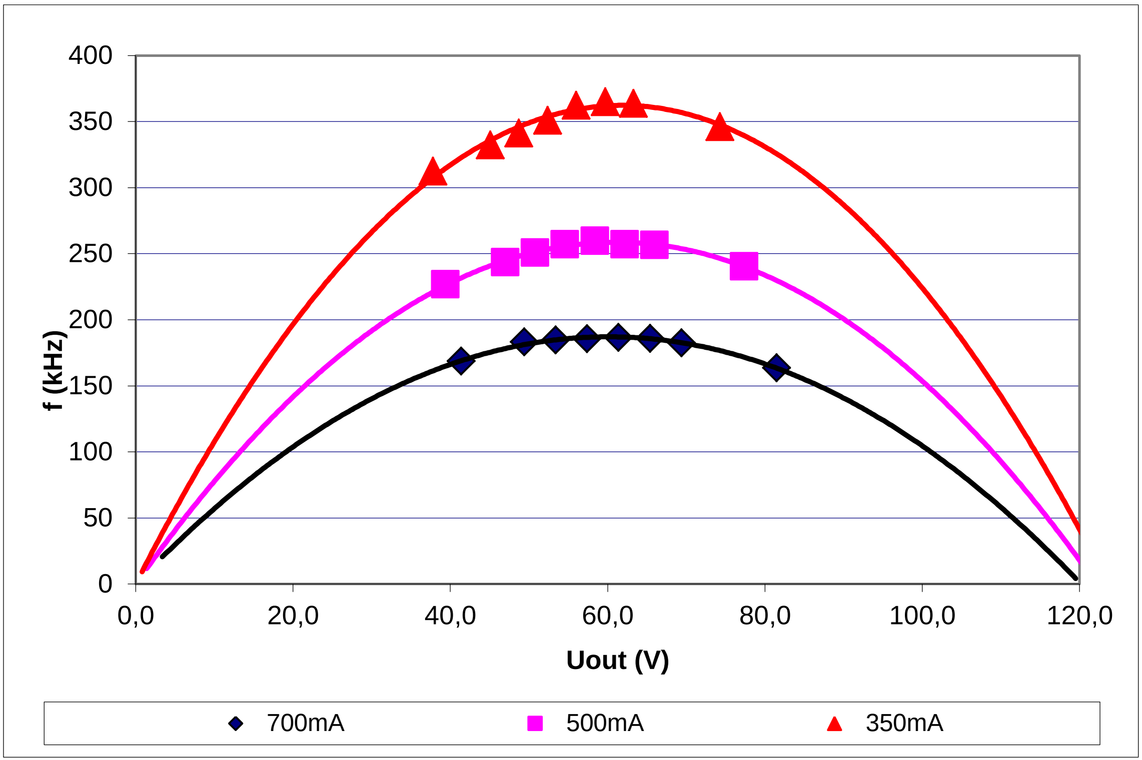

Figure 4.2. *Frequency curves with different output voltages and currents.*

In Figure 4.2. the red triangles are the simulated frequency points when the output current is 350 mA and the red curve is the polynomial fitting for these simulated points. The pink squares are the simulated frequencies using the 500 mA output current, and the pink curve is the polynomial fitting for them. The black squares are drawn by using the 700 mA output current, and the black curve is the polynomial fitting for those points.

Precise simulation results are in Appendix 1 Tables 1. and 2. Table 1 results are simulated with 60 V input voltage and Table 2 results with 120 V. There is also a Figure for frequencies when the input voltage is 60 V, Figure colors and marks are the same as Figure 4.2.

4.3 Variable Input Voltage Simulations

One solution to prevent too high frequencies and too high variance of frequency as a function of the output voltage is to use variable input voltage, then the operational frequency doesn't change as much as in fixed input voltage situation. Simulations are done by using a constant voltage difference between the input and output, in this case a constant 15 V difference is used. Another way to calculate the voltage difference is by using some percentage of the output voltage. The inductor inductance is 680 µH to maintain frequencies about as high as earlier simulations.

The simulation results are shown in Figure 4.3. As can been seen, the frequencies stay more stable, but there is also frequency doubling when the output current is reduced to half. If a constant percentage voltage difference is used the frequency changes would be even lower. Precise simulation results are in Appendix 2 Table 1.

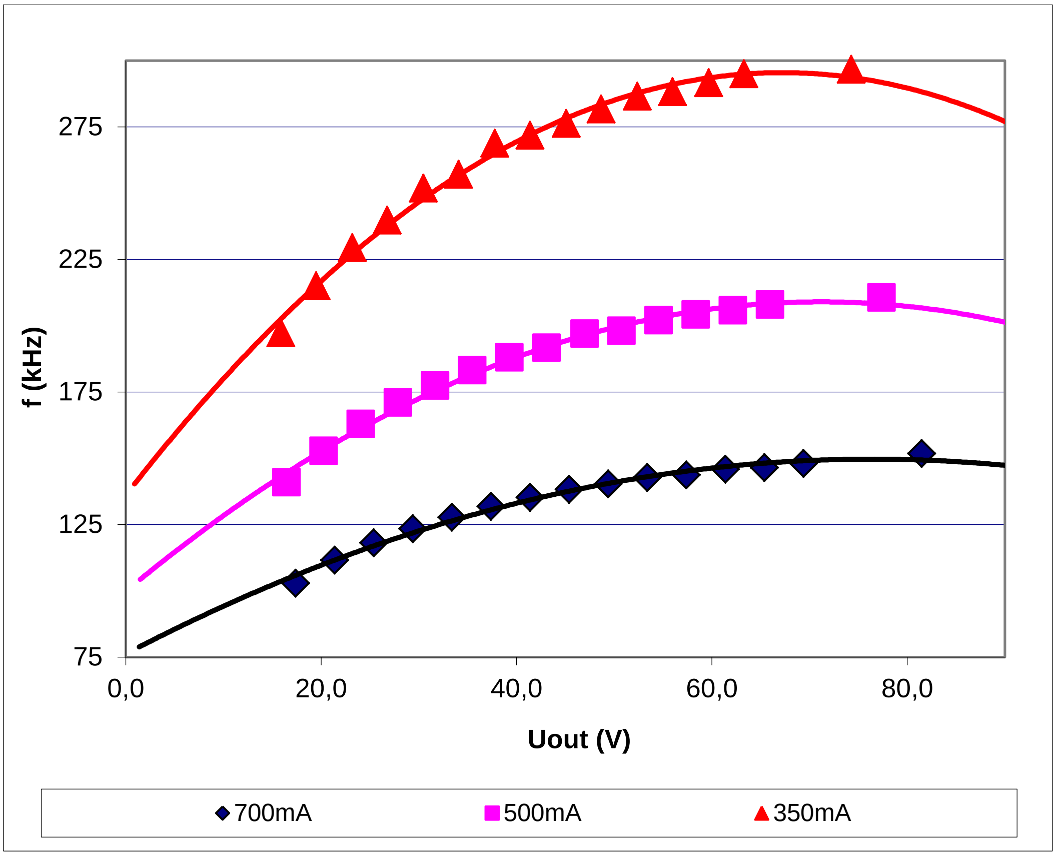

Figure 4.3. *Frequency curves when using variable input voltage.*

In Figure 4.3. the red triangles are the frequency points which are simulated when the output current is 350 mA. The red curve is the polynomial fitting for these points. The pink squares are the frequencies simulated by using the 500 mA output current. The pink curve is the polynomial fitting for them. The black squares are drawn by using the 700 mA output current, and the black curve is the polynomial fitting for those points.

4.4 BCM Simulations

In the BCM the inductor current reaches zero at every buck cycle, therefore the inductor current ripple is the same as the inductor current peak value. That's why the inductor has to be smaller to achieve the same frequency as in earlier simulations. The inductor inductance is only 56 µH. As shown in Figure 4.4. the frequencies stay more stable than using fixed input voltages, and when the output current is reduced the frequencies increase only by a small amount. Precise simulation results are in Appendix 3 Table 1.

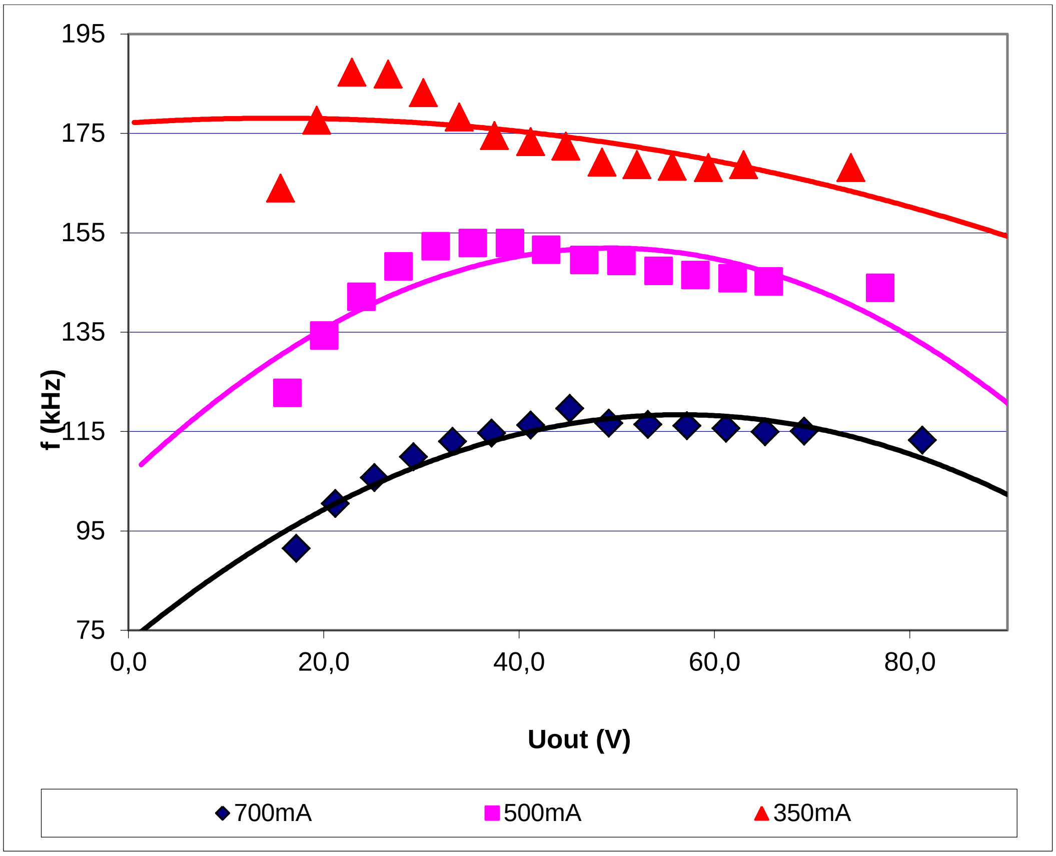

Figure 4.4. *Frequency curves in BCM and when input voltage is variable.*

In Figure 4.4. the red triangles are the simulated frequency points when the output current is 350 mA. The red curve is the polynomial fitting for these simulated points. The pink squares are the simulated frequencies which are simulated by using the 500 mA output current. The pink curve is the polynomial fitting for those points. The black squares are the simulated frequencies which are simulated when the output current was 700 mA. The black curve is the polynomial fitting for them.

4.5 Efficiency Simulations

Efficiency simulations have been done with high load to achieve a high output voltage and optimal operation. The output voltage is about two thirds of the input voltage and the converter operates optimally and with good efficiency. Efficiencies have been simulated with a fixed input voltage where the input voltages were 60 V and 120V, with variable input voltage, and with variable input voltage when the converter operates at BCM. Precise simulation results are in Appendix 4 Tables 1., 2., 3. And 4.

Figures 4.5, 4.6, 4.7 and 4.8 show how power losses are divided between different components and how MOSFET losses are divided into turn on, ON-time and turn off losses. Power losses are simulated with three output current values: 700 mA, 500 mA and 350 mA.

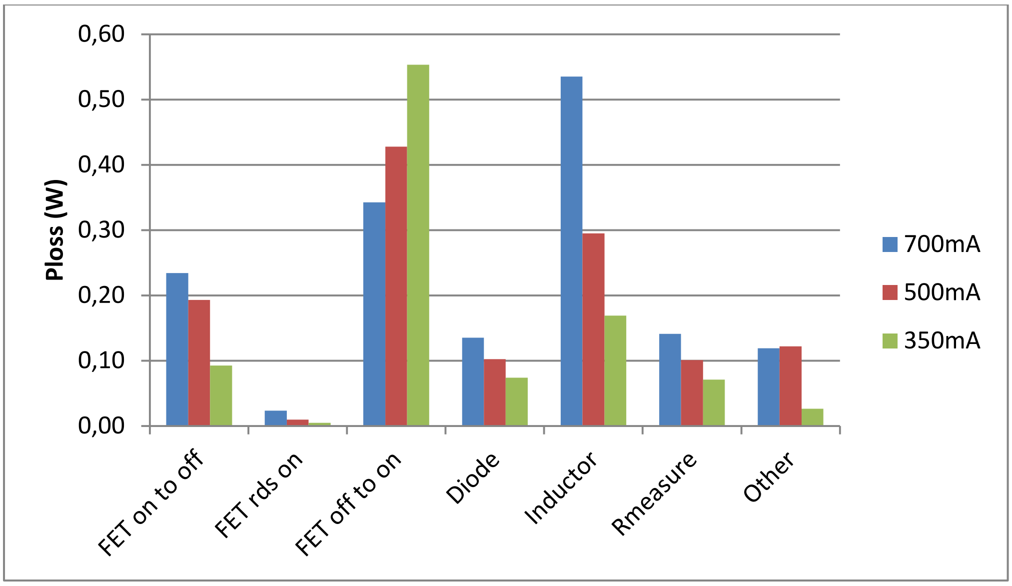

Figure 4.5. *Buck converter losses when U_{in} is 60 V and U_{out} is 40 V.*

As seen in Figure 4.5, a MOSFET turn off loss is dependent on the current through it with proportional relation. An ON-time loss is negligible and a turn on loss is strongly dependent on the operating frequency with inversely proportional relation. Diode losses are proportional to the current through it and inductor losses too, as well as the current sense resistor loss. Other losses are from the MOSFET gate resistor, capacitor and simulation program inaccuracies.

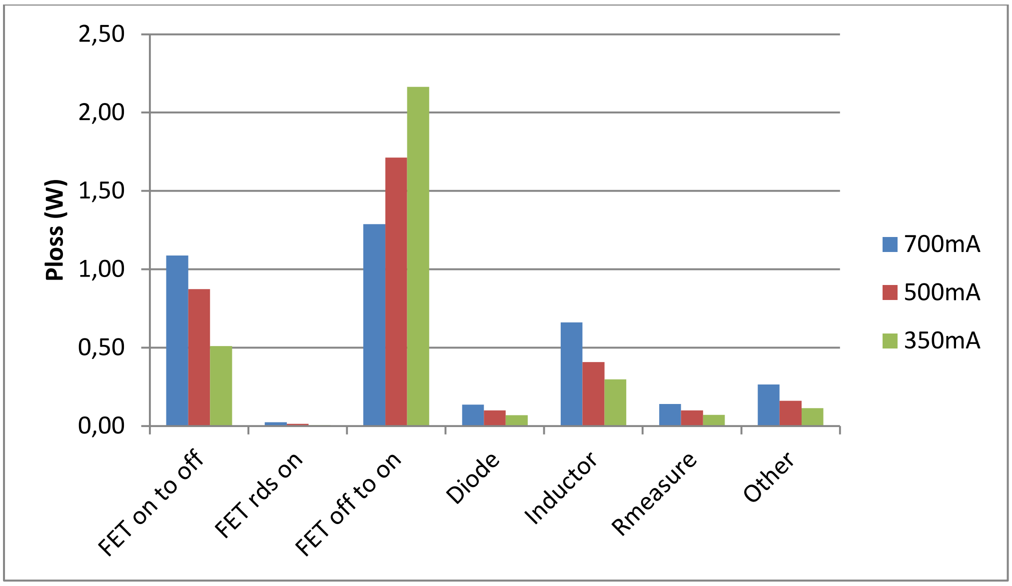

Figure 4.6. *Buck converter losses when U_{in} is 120 V and U_{out} is 80 V.*

When comparing Figures 4.5 and 4.6 it can be seen that the MOSFET turn on and turn off losses are strongly proportional to the voltage across it. The MOSFET turn on and turn off losses are about four times bigger in Figure 4.6 than in Figure 4.5, and other losses are about at the same level as earlier. That's why when using a high voltage buck converter the selection of the MOSFET is important. Another solution is to use a dynamic input voltage, or by using BCM as seen in the next Figures.

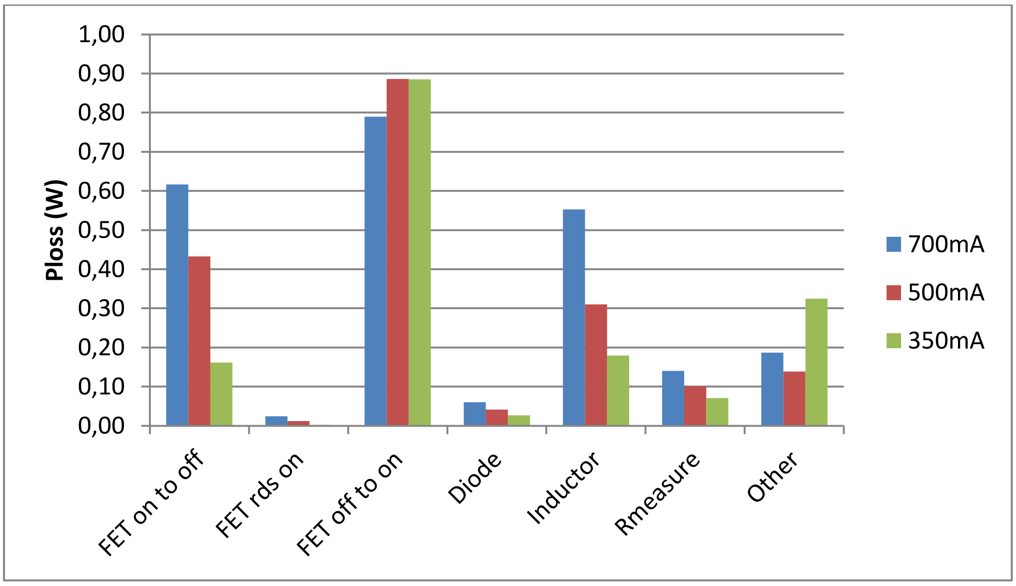

Figure 4.7. *Buck converter losses when U_{in} is dynamic, U_{in}=U_{out} + 15 V.*

As seen in Figure 4.7, the MOSFET turn on and turn of losses are much lower when using a dynamic input voltage. Other losses are at the same level as earlier. The dynamic input voltage also decreases losses if the output voltage dynamics are vast, because lower input voltage increases losses significantly. Especially the diode losses decrease because the duty ratio increases.

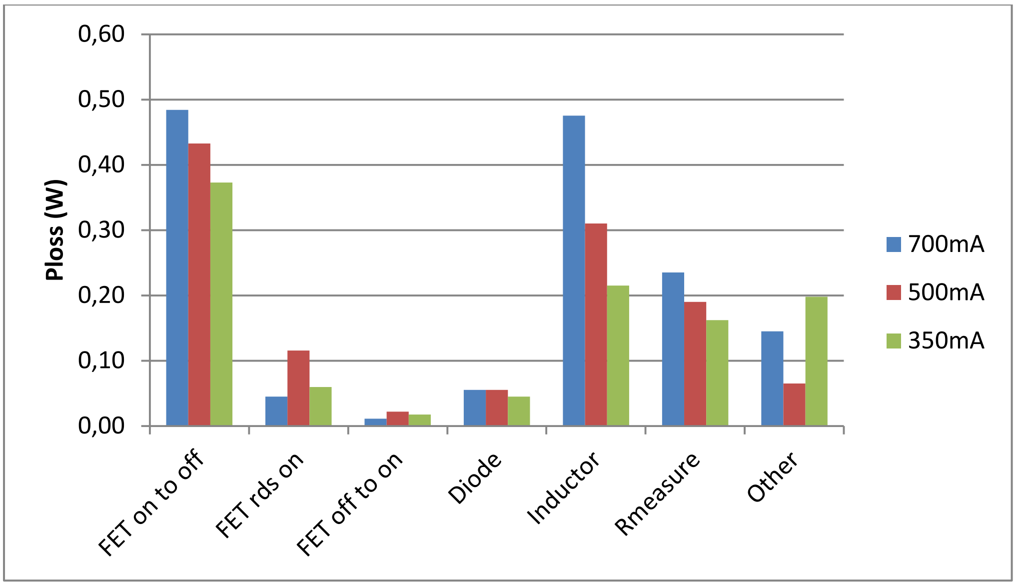

Figure 4.8. *Buck converter losses when U_{in} is dynamic, $U_{in}=U_{out}$ + 15 V and converter operates at the BCM.*

As seen in Figure 4.8, the MOSFET turn on loss become negligible because the MOSFET switching is done when the inductor current is zero. Other losses are about equal to earlier situations.

5 PROTOTYPE DEVICE

The prototype device is designed on the basis of the earlier study. In this Chapter the prototype device is introduced and the most important component choices have been described too. There are two main parts of the prototype device: the main circuit and the control circuit. Both of them are important and they will be introduced.

5.1 Main Circuit

The prototype device is designed according to the demands which were introduced in Chapter 3. The schematic of the prototype device is in Appendix 6. The buck alternative is the inverse buck and its main circuit components are chosen according to the demanded output voltage and current as well as the input voltage. The chosen main circuit and component names are in Figure 5.1.

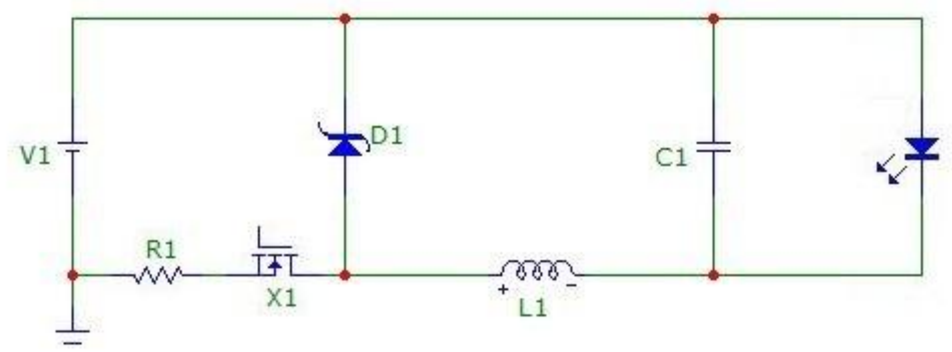

Figure 5.1. *The main circuit and component names of the protype.*

Since the chosen alternative is the inverse buck and the current sense resistor is located after the main switch, a good way to control the buck converter is using a constant OFF-time control method. Now there are two choices: design the control circuit with discrete components, or by using a control IC from the manufacturer. There is a suitable control IC from ON Semiconductor: NCL30105, which meets the specifications. Demands for the control circuit or IC are the ability to dimming the LED load with PWM and analog dimming, good MOSFET driving ability, constant OFF-time control method, and small size. This IC meets these demands.

Then the operating point (operating frequency, and inductor current ripple) have to be selected according to the rated output power. The nominal output power is 30 W and the output current is 700 mA, which means that the nominal output voltage is about 42 V. A suitable operation frequency is 250 kHz and a suitable inductor current ripple is 200 mA. Therefore the inductor peak current is 800 mA and the valley current is 600 mA. According to the desired operation frequency the total switching time has to be 4 µs. If input and output voltages are chosen, the OFF-time has to be 1,9 µs. To achieve 200 mA ripple and 1,9 µs OFF-time the inductance of the inductor has to be 400 uH. The IC current sense voltage reference is 1 V (ON Semiconductor 2011). To achieve 800 mA peak current the current sense resistor has to be 1,25 **Ω**. This can be done by connecting 2,2 **Ω**, 3,3 **Ω** and 22**Ω** in parallel.

The MOSFET has to meet the specifications which are told in the Subsection 3.2.3. A suitable MOSFET is SUD06N10-225L, whose maximum voltage is 100 V, maximum current is 6.5 A, gate charge is 2,7 nC, a r_{DS} is 0,2 **Ω**, package is DPAK, and maximum temperature is 175 ^{o}C. Those properties guarantee very fast and reliable operating as a buck converter main switch. The chosen diode is STPS1H100 whose maximum voltage is 100 V, maximum current is 1 A, and type is Schottky, so switching is fast.

The inductor design is size and loss optimizing but, as told in Subsection 3.2.4, the magnetic core should be closed to decrease EMI and wire turns. The desired inductance is 400 µH and the design calculations and specifications are in the Appendix 5. The output capacitor is chosen according to the LED string dynamic resistance by using equation 5.1 (Texas Instruments 2007).

$$C_{out} = \frac{1}{2\cdot\pi\cdot f_S\cdot(r_C+Z_C)}, Z_C = \frac{\Delta i_{LED}}{\Delta i - \Delta i_{LED}} \cdot r_{LED} \tag{5.1.}$$

In equation 5.1, f_s is the switching frequency 250 kHz, r_C is the capacitor ESR 0,05 **Ω**, Δi_{LED} is the LED string max ripple 70 mA, Δi is the inductor current ripple 200 mA, and r_{LED} is the LED string dynamic resistance 3 **Ω**. According to that, Z_C is calculated in equation 5.2 and the output capacitor value in equation 5.3. A suitable output capacitor value is 220 nF. In Table 5.1 the main circuit components are listed with their most important properties.

$$Z_c = \frac{70\ mA}{200\ mA - 70\ mA} \cdot 6\ \Omega = 3{,}2\ \Omega \tag{5.2.}$$

$$C_{out} = \frac{1}{2\cdot\pi\cdot 250\ kHz\cdot(0{,}05\ \Omega + 3{,}2\ \Omega)} = 196 \cdot 10^{-9} F \tag{5.3.}$$

Table 5.1. *Properties of components of the main circuit.*

	Component type	Value/model	Other info
V1	Input voltage	70 V	Flyback
R1	The current sense resistor	1,25 Ω	Parallel of 3,3Ω, 2,2Ω and 22Ω
X1	The main switch	sud06n10-225l	100 V, 6.5 A 2,7nC
D1	The diode	stps1h100	100 V, 1 A, Schottky
L1	The inductor	400 µH	1 A, 1 Ω, closed core type
C1	The output Capacitor	220 nF	Film capacitor

The input voltage as well as the buck IC supply voltage (15 V) have to be isolated to achieve SELV requirements. A flyback converter is a good method for input voltage generation. There are also other components in the main circuit, like MOSFET gate resistor, which slow the MOSFET switching down to minimize EMI and eliminate trace inductance and the MOSFET parasitic capacitance resonance phenomenon. There is also a resistor from the gate to ground to prevent the MOSFET switching from EMI or miller capacitance leakage current. There is also another diode and ferrite bead which can reduce the MOSFET turn-on loss, because a signal diode has lower capacitance than a Schottky diode and only a signal diode parasitic capacitance discharge current is added to the MOSFET turn-on peak current.

5.2 Control Circuit

The control circuit is implemented by using analog components. The control circuit gets the PWM signal via an opto-isolator. This signal is filtered to a steady voltage level. The opto-isolator is used to achieve SELV requirements. This voltage level is the input signal for the PWM dimming operational amplifier and the analog dimming operational amplifier as seen in the Appendix 6. It is the intention that first comes the analog dimming from 100% to 33% and then comes the PWM dimming with a duty cycle from 100% to 3%. These combine to produce light dimming from 100% to 1%.

The analog dimming operational amplifier scales the input signal to be suitable for the buck IC analog dimming input. There is also a minimum level setting to prevent a too low analog dimming level. The buck IC peak current limit is proportional to the analog dimming pin voltage, and 3 V is the nominal level. Below this the peak current is scaled. The analog dimming circuit scales its input signal voltage level from 3 V to 1 V when the input duty cycle is 100% to 33%.

The PWM dimming operational amplifier is the Sallen-Key filter which controls the PWM-signal generator IC and filters the input signal. The PWM generator IC changes the input voltage levels to a PWM-signal. The input voltage level 0,1 V is the duty cycle of 0%, and 0,9 V is the duty cycle of 100%. The duty cycle difference of 1% is about 0,008 V. The minimum input voltage level has to be 0,124 V to achieve a 3% PWM-signal. The frequency of the PWM signal is set to 260 Hz. The output of the PWM IC is connected directly to the buck IC PWM input pin, because the PWM IC driving ability allows that.

The chosen buck IC generates the constant OFF-time by a current mirror. When the control circuit changes this current mirror current it also changes the OFF-time. The OFF-time has to be changeable since the output voltage changes too. A good method for changing the current mirror current is by using two timing resistors and a variable voltage to another resistor as told in Subsection 2.3.2. A BJT is a good method to generate this voltage, and an operating amplifier is a good way to adjust the voltage over the BJT. The OFF-time has to be inversely proportional to the output voltage. Good feedback is obtained from the MOSFET drain, whose filtered voltage level is the buck input voltage minus the buck output voltage. The OFF-time also has to be changed during analog dimming to achieve the correct hysteresis and ripple. That’s why the OFF-time of the operational amplifier also needs an analog dimming signal level to its input. Those inputs can be implemented by using a summing circuit as seen in Appendix 6.

The output current change is implemented by changing the opto-isolator supply voltage with an operational amplifier. The input voltage can be changed by an external resistor which sets the reference voltage of the operational amplifier. The operational amplifier set for the opto-isolator supply voltage is three times higher than the reference voltage.

6 RESULTS

In this Chapter the prototype device measurements are shown. Measurements include efficiency measurements, operating point measurements, control circuit measurements, analog dimming frequency change measurements and the dimming curve. The results are measured by using Osram 3 W Golden Dragon LEDs. When current through them is about 700 mA, the voltage across them is about 3,45 V. To achieve maximum output power 12 LEDs have to be connected in series when using 700 mA output current.

6.1 Buck Converter

First the buck converter proper operation is confirmed and the operation point is measured. The nominal output voltage is about 42 V. This is also the operation point output voltage. Other values like output current and operation frequency are shown in Table 6.1. As seen, the operating point is suitable for the buck converter.

Table 6.1. *The nominal operation point of the buck converter.*

Uout (V)	Iout (mA)	Operation frequency (kHz)	Inductor current ripple (mA)	Output current ripple (mA)
42,5	708	192	204	77

Efficiency is also an important thing when confirming the buck converter operation. Efficiency is measured with different output voltages. The buck efficiency is drawn by a blue line in Figure 6.1. and, when using a diode and ferrite bead circuit to lower the MOSFET losses, the efficiency is better and is drawn in a green color. In Figure 6.1. the output current is 700 mA and in Figure 6.2. the output current is 350 mA. In this Figure the results are measured with a signal diode and ferrite. Precise measurements are shown in Appendix 7 Tables 1, 2 and 3.

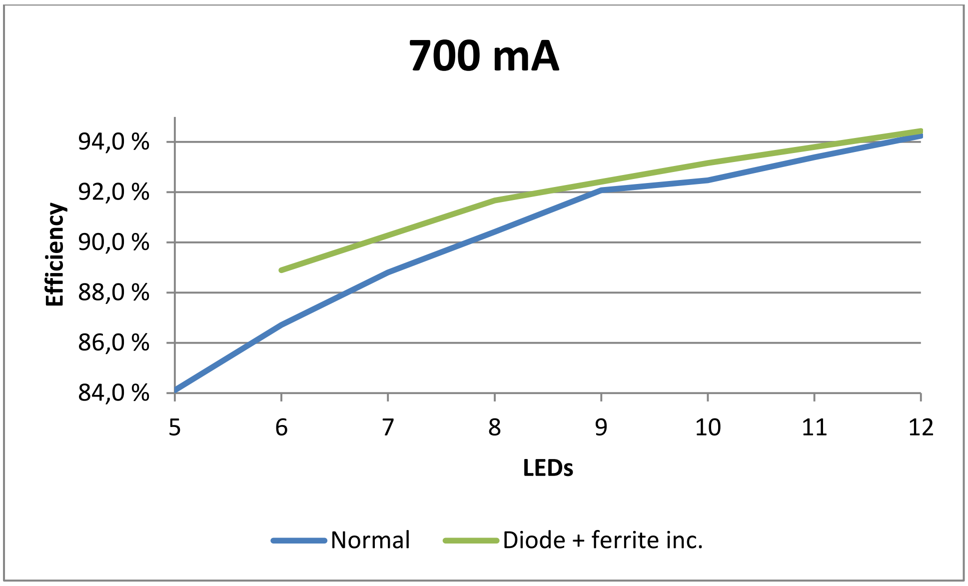

Figure 6.1. *Efficiency of the buck converter when output current is 700 mA.*

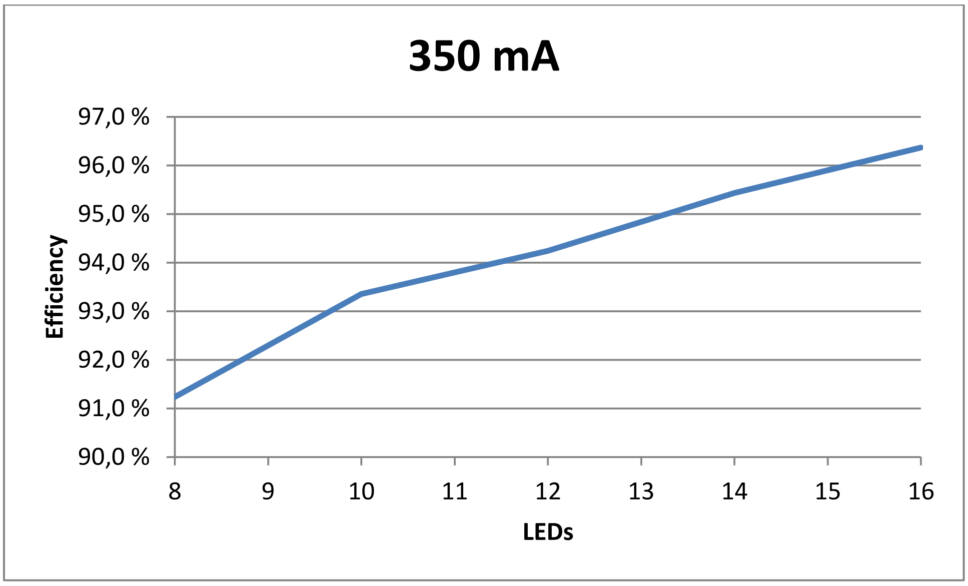

Figure 6.2. *Efficiency of the buck converter when output current is 350 mA.*

As seen in Figures 6.1 and 6.2, the efficiency of the buck converter is high, especially when the output current is 350 mA. When the output voltage decreases, the efficiency decreases too, because the duty ratio is getting lower and the diode losses increase fast. As seen in Figure 6.1, the signal diode and ferrite bead is a good way to increase the buck converter efficiency. This especially lowers the MOSFET losses and MOSFET temperature.

6.2 Control Circuit

The control circuit keeps the output current at the correct value although the output voltage changes. The nominal output current is 700 mA and this is achieved when the output voltage is nominal too. If the output voltage is lower than nominal then the OFF-time has to be longer to achieve the correct output current. Figure 6.3 shows how stable the output current is and, as seen, it is stable enough at the specified output voltage levels. The specified output voltages levels are 30 – 42 V when the output current is 700 mA, and 40 – 60 V when the output current is 350 mA.

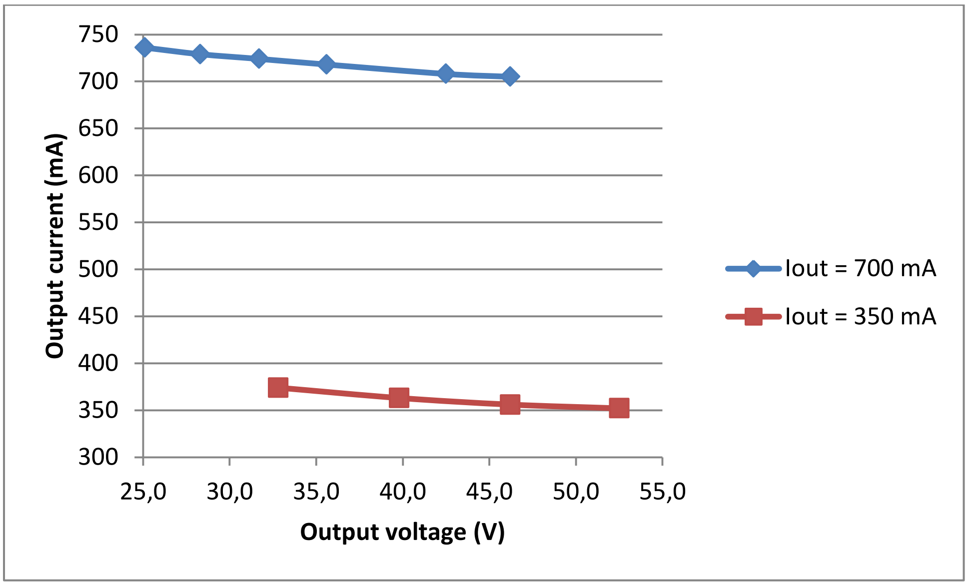

Figure 6.3. *Output current stability respect to output voltage.*

When the output voltage decreases the operating frequency increases, because the ON-time gets shorter. The ON-time gets shorter because the inductor current rising speed gets higher and a shorter time is needed to achieve the needed hysteresis. The operating frequency changes can be seen in Figure 6.4. The blue line is drawn when the output current is 700 mA and red line when the output current is 350 mA.

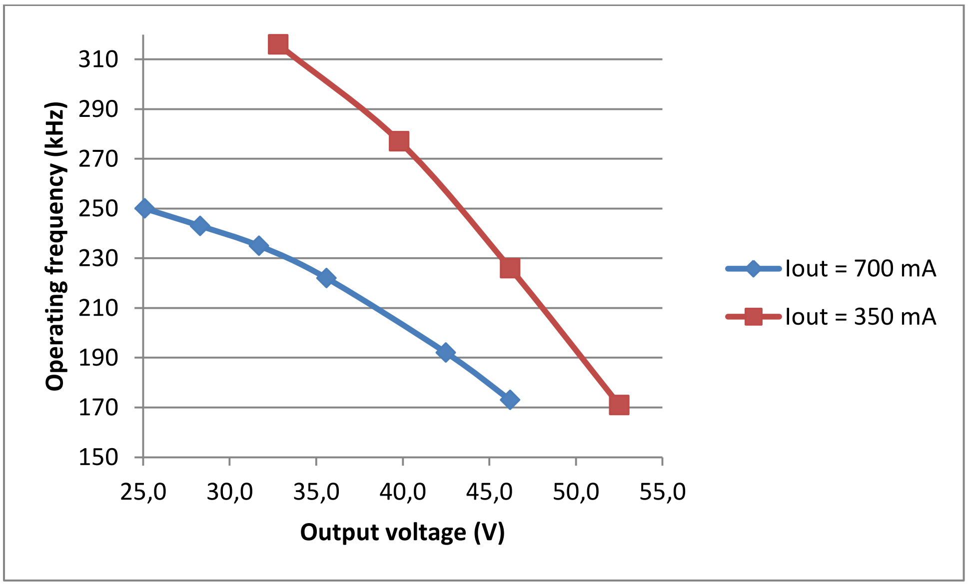

Figure 6.4. *Operating frequency change respect to output voltage.*

Another important thing to measure is the output current ripple. It has to be within specific limits to prevent over stress to LEDs. The measured values for the output current ripple in respect to output voltage are drawn in Figure 6.5. The blue line is the ripple when the output current is 700 mA and red is when the current is 350 mA.

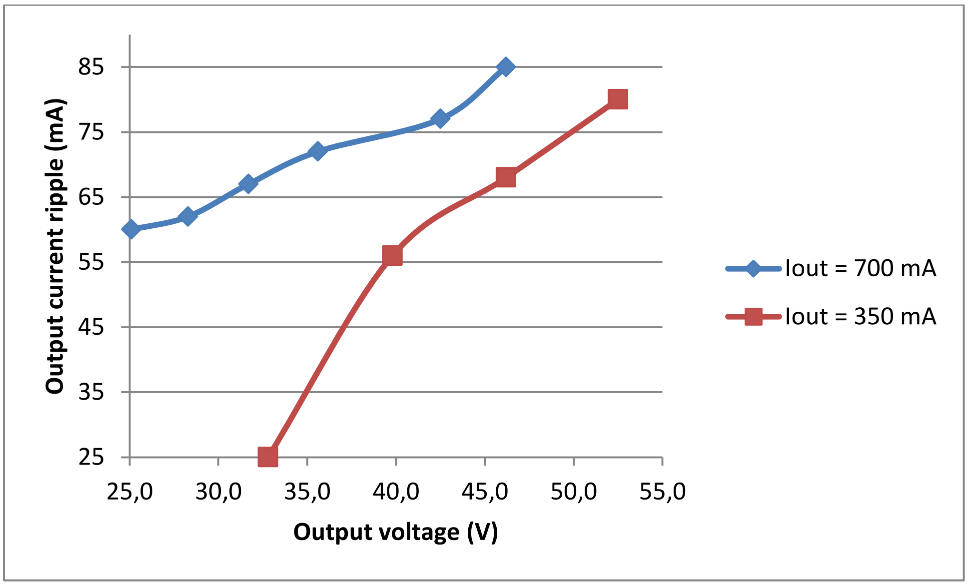

Figure 6.5. *Output current ripple change respect to output voltage.*

The allowable ripple is about 10-15% of the average current when the output voltage level is nominal or lower, and when the output current is 700 mA the ripple is below the limit as calculated in Section 5.1. Respectively, when the output current is 350 mA, the ripple isn't below the limit. This is expected because the output capacitor is only selected for the 700 mA output current. The output capacitor has to be reselected if the current ripple is required to get below the limit when the output current is 350 mA. Precise measurements are shown in Appendix 7 Tables 4 and 5.

6.3 Dimming Performance

The dimming results are very encouraging because dimming seems to be step-less even at the smallest light level. The human eye cannot see steps when the LEDs are dimmed, because the steps are so tiny. The steps are tiny because hybrid dimming is used. Compared to pure PWM dimming the difference is huge.

Analog dimming also increases operational frequency and the change of it is measured. The frequency is increased to add buck cycles before the PWM dimming method starts dimming the LEDs. As seen in Figure 6.6. the increase of the operational frequency is working well. Precise measurements are shown in Appendix 7 Table 6.

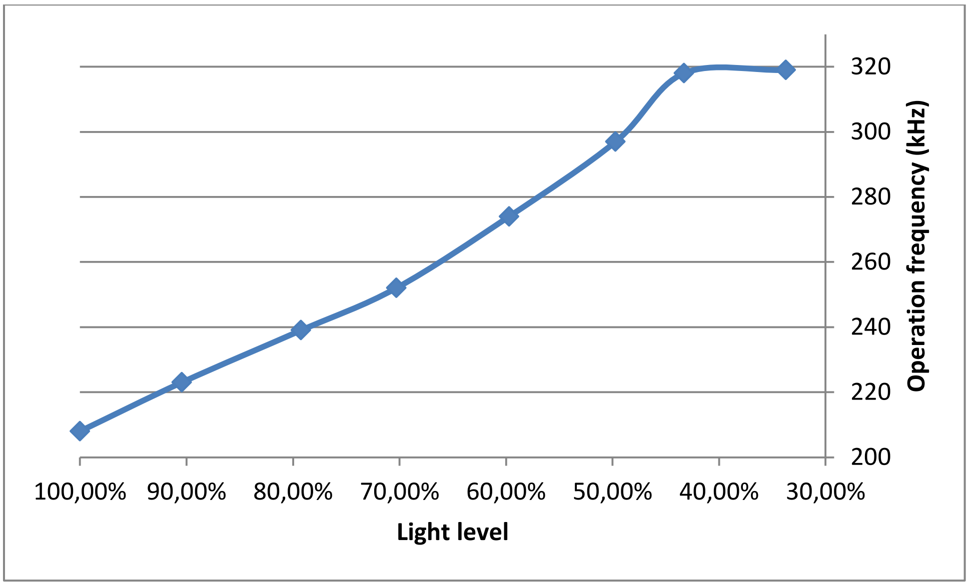

Figure 6.6. *The analog dimming frequency change.*

The dimming curve is also an important thing when measuring dimming performance. The curve is measured by dimming the LEDs and measuring the output current levels. Those measured current levels are converted to relative light output levels. The curve can be seen in the Figure 6.7 and precise measurements are in Appendix 7 Table 7.

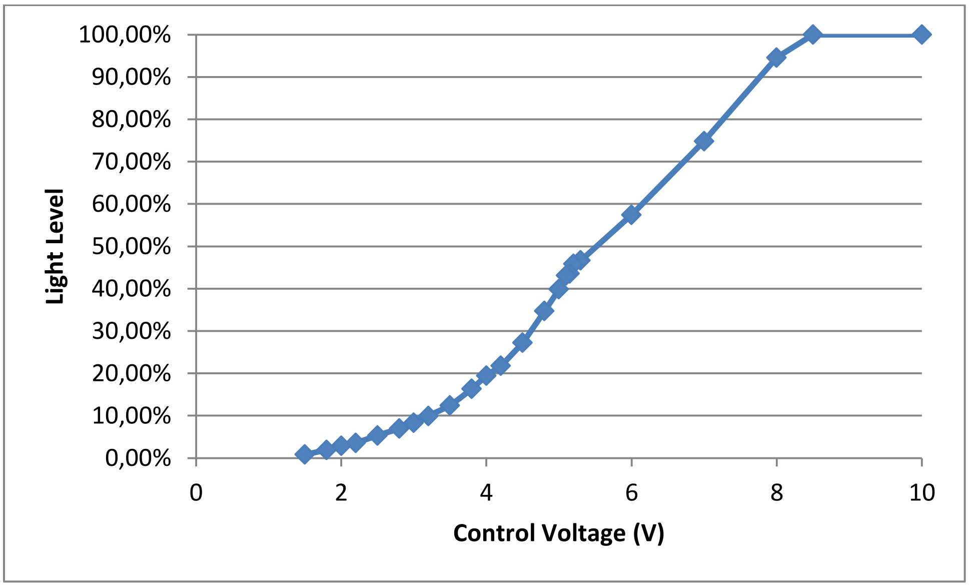

Figure 6.7. *The dimming curve.*

As seen, there is inaccuracy when the dimming method changes from analog dimming to PWM dimming. The change should be sharp to achieve the desired dimming curve, but now the change isn't sharp and both methods are effective at the same time. The result is inaccuracy in the dimming curve, but the curve is good enough because the inaccuracy cannot be detected by the human eye.

7 CONCLUSION

Improving efficiency in lighting solutions is very important and there is a need for high quality dimmable LED current sources in the market. In this thesis the LED current source output stage was optimized for modular and multichannel operation. High quality dimming was also demanded, which means step-less logarithmic dimming. Based on the basic theory a prototype device was designed and the results were encouraging.

The LED current source output stage was designed by using a step down switched-mode converter, also known as a buck converter. The buck converter was implemented by using a low-side main switch and a low-side inductor, then multichannel solutions are easier and more cost-effective to be made. When the main switch is low-side its driver circuit is easier and more cost-effective to implement because no high voltages are needed. This can also be done with a microcontroller which make dimming control easier. The control circuit is very important when designing high quality devices. Also the current sense is important because the whole control loop is based on that. The control method of the main switch was chosen to have a constant OFF-time. The feedback circuit was then easier to be implemented and it is cost-efficient, because no high voltage exist. The main switch, usually a MOSFET, is the highest source of power loss, since MOSFET switching generates high power loss and the switching frequency is relatively high. One solution to decrease the MOSFET switching loss is using an additional signal diode and ferrite, which decreases the peak current through the MOSFET during the turn-on procedure.

In this thesis three dimming methods were introduced. The methods were PWM dimming, analog dimming and hybrid dimming. PWM dimming can cause different problems, but the most severe is visible steps between light levels. In this thesis some solutions for decreasing the effect of these steps were introduced. Usable methods are using a shunt-MOSFET, series-MOSFET, hybrid dimming or calculating PWM-steps only at the main switch ON-time. The shunt-MOSFET and series-MOSFET methods are good methods to increase dimming accuracy because with them the inductor energy charging and discharging times don’t affect to the output current. With a series-MOSFET, LED leakage current also isn’t problem. PWM-step calculation only at the main switch ON-time is also an effective way to increase dimming accuracy, but then there is a need for a very high speed microcontroller. Hybrid dimming combines PWM- and analog dimming methods together, and then the inductor energy charge and discharge times decrease very much and their influence to the output current is acceptable. In the prototype device the hybrid dimming method was in use.

The prototype device was designed and assembled for measurements and tests. Efficiency of the prototype device was measured as well as the control circuit operation. Efficiency was at a good level, especially when noticing the buck control IC high power consumption. If the IC is changed the efficiency could be better. An extra signal diode with ferrite bead increased the efficiency as assumed, especially at low output voltage levels. The control circuit kept the output current and ripple at acceptable levels when the output voltage changes.

The results of hybrid dimming are very encouraging because visible steps were not detected during dimming. This was due to a well-chosen dimming algorithm which used one third analog dimming level and increased the buck converter operational frequency during it. When the minimum analog dimming level was reached the PWM dimming method started to cut current flow through the LEDs. Since the analog dimming level was used, the narrowest PWM-pulse could be wider than using pure PWM dimming. This increased dimming accuracy especially at low light output levels.

Further research topics are to find a better solution for the buck control IC, because now it consumes too much power. The choices are to replace the current IC with a another, to use a microcontroller for control, or to design a control circuit with discrete components. The second research is to confirm the benefits of hybrid dimming and to measure how much analog dimming changes the wavelength of the LED. Also the dimming curve needs to be fixed.

REFERENCES

Aimtec 2011, Application note: Pulse Width Modulation (PWM) vs. Analog Dimming of LEDs [WWW]. [Cited 17.7.2012]. Available at: http://www.aimtec.com/pulse-width-modulation-pwm-vs-analog-dimming-of-leds.

Beczkowski S. & Munk-Nielsen S., 2010. Led Spectral and Power Characteristics Under Hybrid PWM/AM Dimming Strategy [WWW]. [Cited 17.7.2012]. Available at: http://vbn.aau.dk/files/39948127/Led_spectral_and_power_characteristics_.pdf.

Eichhorn T., 2005. Estimate Inductor Losses Easily in Power Supply Designs [WWW]. [Cited 17.7.2012]. Available at: http://powerelectronics.com/mag/504PET20.pdf.

Helvar 2012, EL2x49sc Technical Specifications. [Cited 4.10.2012].

iWatt 2012, iW3612 Datasheet. [WWW]. [Cited 28.9.2012]. Available at: http://www.iwatt.com/pdf/Datasheet/iW3612_Datasheet.pdf.

Maxim 2001, Application Note 2031: DC-DC Converter Tutorial [WWW]. [Cited 21.5.2012]. Available at: http://www.maxim-ic.com/app-notes/index.mvp/id/2031.

Maxim 2002, Application Note 1832: Power Supply Engineer's Guide to Calculate Dissipation for MOSFETs in High-Power Supplies [WWW]. [Cited 25.5.2012]. Available at: http://www.maxim-ic.com/app-notes/index.mvp/id/1832.

Mohan N., Undeland T. & Robbins W. 1995. Power Electronics: Converters, Applications and Design. 2nd edition. John Wiley & Sons Inc. 802 p.

ON Semiconductor 2011, NCL30105 Datasheet. [WWW]. [Cited 21.5.2012]. Available at: http://www.onsemi.com/pub/Collateral/NCL30105-D.PDF.

Osram 2011, Golden Dragon Plus LW W5AM –datasheet. [WWW]. [Cited 2.10.2012]. Available at: http://catalog.osram-os.com/catalogue/catalogue.do;jsessionid=5C2C7235FFBC6A57B123F3D59EC3208C?act=downloadFile&favOid=02000000000260e1000200b6.

Richardson, C. 2012. Cutting The Buzz in LED Drivers, Part II. EDN Europe, 3, pp. 18-21.

Simpson R., 2003. Lighting Control: Technology and Applications Focal Press 564p.

Texas Instruments 2006. Application note 1487: Current Mode Hysteretic Buck Regulators [WWW]. [Cited 21.5.2012]. Available at: http://www.ti.com/lit/an/snva170a/snva170a.pdf.

Texas Instruments 2007. Driving LEDs: To Cap or Not to Cap [WWW]. [Cited 15.8.2012]. Available at: http://www.ti.com/lit/an/snva598/snva598.pdf.

Texas Instruments 2011. Dimming Techniques for Switched-Mode LED Drivers [WWW]. [Cited 18.7.2012]. Available at: http://www.ti.com/lit/an/snva605/snva605.pdf.

APPENDIX 1: Fixed Input Voltage Simulation Results

Simulation results are in Table 1. Input voltage is a constant 60V and the inductor inductance is 680 µH.

Table 1. *Results when buck input voltage is 60 V.*

Uin (V)	LEDs	Uout (V)	Iout (mA)	Δi (mA)	t (us)	f (kHz)	Pin (W)	η (%)	Pfet,avg (W)	Pfet,on (mW)
60	10	41,5	700	105	6,13	163	30,5	95	0,60	31
60	9	37,4	700	105	5,54	181	27,6	94	0,65	31
60	8	33,4	700	105	5,24	191	25,0	93	0,69	31
60	7	29,3	700	105	5,14	195	21,9	92	0,71	31
60	6	25,4	700	105	5,23	191	19,5	91	0,70	31
60	5	21,4	700	105	5,51	181	16,4	90	0,67	31
60	4	17,4	700	105	6,05	165	13,7	88	0,60	31
60	10	39,3	500	77	4,17	240	20,8	94	0,63	18
60	9	35,5	500	77	3,86	259	19,0	93	0,66	18
60	8	31,7	500	77	3,74	267	17,0	93	0,69	18
60	7	28,0	500	77	3,72	269	15,2	91	0,67	18
60	6	24,1	500	77	3,86	259	13,2	90	0,65	18
60	5	20,3	500	77	4,06	246	11,3	89	0,61	18
60	4	16,5	500	77	4,57	219	9,4	87	0,55	18
60	10	37,8	350	54	2,83	353	14,2	93	0,65	9
60	9	34,1	350	54	2,69	372	13,0	91	0,70	9
60	8	30,5	350	54	2,64	379	11,7	91	0,70	9
60	7	26,8	350	54	2,69	372	10,5	89	0,69	9
60	6	23,2	350	54	2,75	364	9,2	88	0,69	9
60	5	19,5	350	54	2,96	338	7,8	87	0,64	9
60	4	15,9	350	54	3,37	297	6,5	84	0,56	9

In Figure 1. the red triangles are the simulated frequency points when the output current is 350 mA, and the red curve is the polynomial fitting for these simulated points. The pink squares are the simulated frequencies using 500 mA output current, and the pink curve is the polynomial fitting for them. The black squares are drawn using the 700 mA output current, and the black curve is the polynomial fitting for those points.

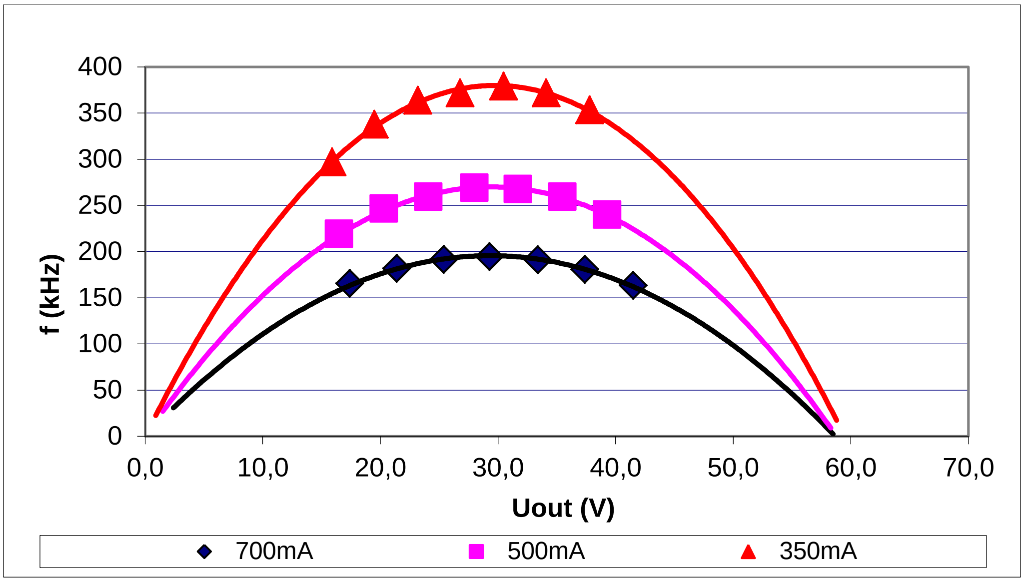

Figure 1. *Frequency curves with different output voltages.*

Simulation results are in Table 2. Input voltage is a constant 120V. The inductor inductance is 1360 µH. Its value has been doubled so that the frequencies would be the same as earlier simulations.

Table 2. *Results when buck input voltage is 120 V.*

Uin (V)	LEDs	Uout (V)	Iout (mA)	Δi (mA)	t (us)	f (kHz)	Pin (W)	η (%)	Pfet,avg (W)	Pfet,on (mW)
120	20	81,5	700	105	6,11	164	60,0	94	2,40	32
120	17	69,4	700	105	5,48	182	52,0	93	2,65	32
120	16	65,4	700	105	5,38	186	49,4	92	2,75	32
120	15	61,4	700	105	5,36	187	47,0	91	2,75	32
120	14	57,4	700	105	5,39	186	44,3	90	2,75	32
120	13	53,4	700	105	5,42	185	41,2	90	2,70	32
120	12	49,4	700	105	5,46	183	38,5	89	2,65	32
120	10	41,4	700	105	5,93	169	32,3	89	2,45	32
120	20	77,4	500	77	4,16	240	42,1	92	2,60	17
120	17	66,0	500	77	3,90	256	36,6	90	2,75	17
120	16	62,2	500	77	3,89	257	34,7	89	2,77	17
120	15	58,4	500	77	3,85	260	32,9	88	2,77	17
120	14	54,6	500	77	3,89	257	30,8	88	2,73	17
120	13	50,8	500	77	3,99	251	28,9	87	2,68	17
120	12	47,0	500	77	4,11	243	27,0	87	2,62	17
120	10	39,4	500	77	4,41	227	23,0	85	2,40	17
120	20	74,3	350	54	2,89	346	29,4	89	2,68	8
120	17	63,3	350	54	2,75	364	25,5	87	2,85	8
120	16	59,7	350	54	2,74	365	24,4	86	2,86	8
120	15	56,0	350	54	2,76	362	23,2	85	2,85	8
120	14	52,4	350	54	2,85	351	21,8	84	2,82	8
120	13	48,7	350	54	2,93	341	20,5	83	2,80	8
120	12	45,1	350	54	3,01	332	19,2	82	2,72	8
120	10	37,8	350	54	3,20	313	16,3	81	2,50	8

APPENDIX 2: Variable Input Voltage Simulation Results

Table 1. *Results when buck input voltage is output voltage plus 15 V and* the inductor inductance is 680 μH.

Uin (V)	LEDs	Uout (V)	Iout (mA)	Δi (mA)	t (us)	f (kHz)	Pin (W)	η (%)	Pfet,avg (W)	Pfet,on (mW)
96,5	20	81,5	700	105	6,59	152	59,2	96	1,43	32
84,4	17	69,4	700	105	6,76	148	50,4	96	1,06	32
80,4	16	65,4	700	105	6,83	146	47,5	96	0,95	32
76,4	15	61,4	700	105	6,86	146	44,4	97	0,87	32
72,4	14	57,4	700	105	6,96	144	41,6	97	0,77	32
68,4	13	53,4	700	105	7,01	143	38,8	97	0,67	32
64,4	12	49,4	700	105	7,13	140	35,7	97	0,60	32
60,4	11	45,4	700	105	7,24	138	33,0	97	0,51	32
56,4	10	41,4	700	105	7,40	135	30,1	97	0,46	32
52,4	9	37,4	700	105	7,59	132	27,3	96	0,40	32
48,4	8	33,4	700	105	7,83	128	24,2	96	0,30	32
44,4	7	29,4	700	105	8,11	123	21,4	96	0,27	32
40,4	6	25,4	700	105	8,47	118	18,8	95	0,21	32
36,4	5	21,4	700	105	8,97	111	16,0	94	0,17	32
32,4	4	17,4	700	105	9,73	103	13,0	93	0,13	32
92,4	20	77,4	500	77	4,75	211	40,4	96	1,33	17
81,0	17	66,0	500	77	4,81	208	34,3	96	0,98	17
77,2	16	62,2	500	77	4,86	206	32,4	96	0,88	17
73,4	15	58,4	500	77	4,90	204	30,5	96	0,78	17
69,6	14	54,6	500	77	4,95	202	28,3	96	0,71	17
65,8	13	50,8	500	77	5,05	198	26,5	96	0,62	17
62,0	12	47,0	500	77	5,08	197	24,5	96	0,54	17
58,1	11	43,1	500	77	5,22	192	22,5	96	0,47	17
54,3	10	39,3	500	77	5,32	188	20,5	96	0,40	17
50,5	9	35,5	500	77	5,46	183	18,4	96	0,34	17
46,7	8	31,7	500	77	5,64	177	16,5	96	0,27	17
42,9	7	27,9	500	77	5,85	171	14,5	96	0,23	17
39,1	6	24,1	500	77	6,14	163	12,6	95	0,19	17
35,3	5	20,3	500	77	6,55	153	10,8	94	0,14	17
31,5	4	16,5	500	77	7,10	141	8,9	93	0,11	17
89,3	20	74,3	350	54	3,37	297	27,6	94	1,23	8
78,3	17	63,3	350	54	3,39	295	23,5	95	0,92	8
74,7	16	59,7	350	54	3,43	292	22,1	95	0,83	8
71,0	15	56,0	350	54	3,47	288	20,7	95	0,74	8
67,4	14	52,4	350	54	3,49	287	19,3	95	0,67	8
63,7	13	48,7	350	54	3,55	282	18,0	95	0,57	8
60,1	12	45,1	350	54	3,62	276	16,6	95	0,51	8
56,4	11	41,4	350	54	3,68	272	15,2	96	0,44	8
52,8	10	37,8	350	54	3,72	269	13,8	96	0,38	8
49,1	9	34,1	350	54	3,89	257	12,6	96	0,32	8
45,5	8	30,5	350	54	3,97	252	11,2	95	0,27	8
41,8	7	26,8	350	54	4,17	240	9,8	95	0,21	8
38,2	6	23,2	350	54	4,36	229	8,6	95	0,18	8
34,5	5	19,5	350	54	4,65	215	7,2	95	0,13	8
30,9	4	15,9	350	54	5,07	197	5,9	94	0,10	8

APPENDIX 3: Variable Input Voltage Simulation Results in BCM

Table 1. *Results when buck input voltage is output voltage plus 15 V in BCM and* the inductor inductance is 56 µH to maintain frequencies the same as earlier simulations.

Uin (V)	LEDs	Uout (V)	Iout (mA)	t (us)	f (kHz)	Pin (W)	η (%)	Pfet,avg (W)	Pfet,on (mW)
96,5	20	81,3	700	8,83	113	58,3	97	0,54	90
84,4	17	69,2	700	8,69	115	49,9	97	0,42	90
80,4	16	65,2	700	8,70	115	47,0	97	0,38	90
76,4	15	61,2	700	8,65	116	44,0	97	0,34	90
72,4	14	57,2	700	8,61	116	41,2	97	0,29	90
68,4	13	53,2	700	8,59	116	38,5	97	0,28	90
64,4	12	49,2	700	8,57	117	35,7	97	0,23	90
60,4	11	45,2	700	8,36	120	32,9	97	0,22	90
56,4	10	41,2	700	8,60	116	30,1	97	0,17	70
52,4	9	37,2	700	8,72	115	27,2	97	0,15	70
48,4	8	33,2	700	8,85	113	24,6	96	0,12	70
44,4	7	29,2	700	9,10	110	21,8	96	0,11	70
40,4	6	25,2	700	9,46	106	19,0	95	0,10	70
36,4	5	21,2	700	9,95	101	16,2	95	0,07	70
32,4	4	17,2	700	10,93	91	13,4	94	0,06	70
92,4	20	77,0	500	6,95	144	39,6	97	0,57	50
81,0	17	65,6	500	6,89	145	33,8	97	0,42	50
77,2	16	61,9	500	6,86	146	31,9	97	0,39	50
73,4	15	58,1	500	6,83	146	29,9	97	0,33	50
69,6	14	54,3	500	6,79	147	28,1	97	0,31	50
65,8	13	50,5	500	6,70	149	26,1	97	0,27	50
62,0	12	46,7	500	6,69	149	24,1	97	0,23	50
58,1	11	42,8	500	6,60	152	22,2	97	0,23	50
54,3	10	39,1	500	6,54	153	20,3	97	0,16	40
50,5	9	35,3	500	6,54	153	18,4	97	0,12	40
46,7	8	31,5	500	6,57	152	16,4	96	0,09	40
42,9	7	27,7	500	6,75	148	14,6	96	0,07	40
39,1	6	23,9	500	7,04	142	12,7	95	0,07	40
35,3	5	20,1	500	7,45	134	10,8	95	0,05	40
31,5	4	16,3	500	8,15	123	8,9	94	0,04	40
89,3	20	74,0	350	5,95	168	26,8	96	0,45	30
78,3	17	63,0	350	5,93	169	22,7	96	0,31	30
74,7	16	59,4	350	5,95	168	21,5	96	0,29	30
71,0	15	55,7	350	5,94	168	20,2	96	0,26	30
67,4	14	52,1	350	5,93	169	18,8	97	0,24	30
63,7	13	48,5	350	5,91	169	17,6	96	0,23	30
60,1	12	44,8	350	5,80	172	16,2	96	0,19	30
56,4	11	41,2	350	5,77	173	15,0	96	0,30	30
52,8	10	37,5	350	5,73	175	13,7	96	0,16	20
49,1	9	33,9	350	5,61	178	12,4	96	0,15	20
45,5	8	30,2	350	5,46	183	11,1	96	0,13	20
41,8	7	26,6	350	5,35	187	9,7	96	0,10	20
38,2	6	22,9	350	5,34	187	8,4	96	0,04	20
34,5	5	19,3	350	5,63	178	7,2	95	0,02	20
30,9	4	15,6	350	6,10	164	5,9	94	0,02	20

APPENDIX 4: Efficiency Simulation Results

Efficiency simulation results are in Tables 1., 2., 3. and 4.

Table 1. *Efficiency results when buck converter input voltage is 60 V.*

Uout	41,5 V			39,3 V			37,8 V		
Iout	700mA			500mA			350mA		
LEDs	10			10			10		
Frequency	163 kHz			240 kHz			353 kHz		
η tot	0,95			0,94			0,93		
Ploss, total	1,53 W			1,25 W			0,99 W		
FET on to off	15,29 %	0,23	W	15,42 %	0,19	W	9,32 %	0,09	W
FET rds on	1,53 %	0,02	W	0,76 %	0,01	W	0,46 %	0,00	W
FET off to on	22,39 %	0,34	W	34,22 %	0,43	W	55,87 %	0,55	W
Diode	8,82 %	0,13	W	8,16 %	0,10	W	7,47 %	0,07	W
Inductor	34,97 %	0,53	W	23,60 %	0,30	W	17,07 %	0,17	W
Rmeasure	9,22 %	0,14	W	8,08 %	0,10	W	7,17 %	0,07	W
Other	7,79 %	0,12	W	9,76 %	0,12	W	2,63 %	0,03	W

Table 2. *Efficiency results when buck converter input voltage is 120 V.*

Uout	81,5 V			77,4 V			74,3 V		
Iout	700mA			500mA			350mA		
LEDs	20			20			20		
Frequency	164 kHz			240 kHz			346 kHz		
η tot	94			92			89		
Ploss, total (W)	3,60 W			3,37 W			3,23 W		
FET on to off	30,20 %	1,09	W	25,92 %	0,87	W	15,76 %	0,51	W
FET rds on	0,66 %	0,02	W	0,39 %	0,01	W	0,17 %	0,01	W
FET off to on	35,80 %	1,29	W	50,85 %	1,71	W	67,01 %	2,16	W
Diode	3,75 %	0,14	W	2,97 %	0,10	W	2,14 %	0,07	W
Inductor	18,33 %	0,66	W	12,11 %	0,41	W	9,20 %	0,30	W
Rmeasure	3,92 %	0,14	W	2,97 %	0,10	W	2,20 %	0,07	W
Other	7,34 %	0,26	W	4,79 %	0,16	W	3,53 %	0,11	W

Table 3. *Efficiency results when buck input voltage is output voltage plus 15 V.*

Uout	81,5 V			77,4 V			74,3 V		
Iout	700mA			500mA			350mA		
LEDs	20			20			20		
Frequency	152 kHz			211 kHz			297 kHz		
η tot	0,96			0,96			0,94		
Ploss, total (W)	2,37 W			1,92 W			1,65 W		
FET on to off	26,01 %	0,62	W	22,51 %	0,43	W	9,77 %	0,16	W
FET rds on	1,03 %	0,02	W	0,62 %	0,01	W	0,11 %	0,00	W
FET off to on	33,31 %	0,79	W	46,13 %	0,89	W	53,67 %	0,89	W
Diode	2,53 %	0,06	W	2,14 %	0,04	W	1,64 %	0,03	W
Inductor	23,33 %	0,55	W	16,15 %	0,31	W	10,85 %	0,18	W
Rmeasure	5,91 %	0,14	W	5,26 %	0,10	W	4,30 %	0,07	W
Other	7,89 %	0,19	W	7,20 %	0,14	W	19,66 %	0,32	W

Table 4. *Efficiency results when buck input voltage is output voltage plus 15 V in BCM.*

Uout	81,3 V			77,0 V			74,0 V		
Iout	700mA			500mA			350mA		
LEDs	20			20			20		
Frequency	113 kHz			144 kHz			168 kHz		
η tot	97			97			96		
Ploss, total (W)	1,45 W			1,19 W			1,07 W		
FET on to off	33,37 %	0,48	W	36,36 %	0,43	W	34,86 %	0,37	W
FET rds on	3,09 %	0,04	W	9,72 %	0,12	W	5,55 %	0,06	W
FET off to on	0,78 %	0,01	W	1,82 %	0,02	W	1,64 %	0,02	W
Diode	3,79 %	0,05	W	4,62 %	0,05	W	4,21 %	0,05	W
Inductor	32,76 %	0,48	W	26,05 %	0,31	W	20,09 %	0,21	W
Rmeasure	16,21 %	0,24	W	15,97 %	0,19	W	15,14 %	0,16	W
Other	10,00 %	0,15	W	5,46 %	0,06	W	18,51 %	0,20	W

APPENDIX 5: The Design of The Inductor

Ae = the ferrite effective area (EF12) $A_e := 12.4 \cdot 10^{-6}$

hw = the height for the wiring $h_w := 2.1 \cdot 10^{-3}$

ww = the width for the wiring $w_w := 7.2 \cdot 10^{-3}$

Aw = the area for the wiring $A_w := h_w \cdot w_w$ $\quad A_w = 1.512 \times 10^{-5}$

Bmax = maximum magnetic flux density $B_{max} := 350 \cdot 10^{-3}$

Imax = maximum current $I_{max} := 1000 \cdot 10^{-3}$

L = the inductance of the inductor $L := 400 \cdot 10^{-6}$

N = number of turns $N := \dfrac{I_{max} \cdot L}{B_{max} \cdot A_e}$ $\quad N = 92.166$

$N_r := \mathrm{ceil}(N)$ $\quad N_r = 93$

η = filling factor $\eta := 0.5$

$A_{wire} := \eta \cdot \dfrac{A_w}{N_r}$ $\quad A_{wire} = 8.129 \times 10^{-8}$

ϕwire = the maximum diameter of the wire $\phi_{wire} := 2 \cdot \sqrt{\dfrac{A_{wire}}{\pi}}$

$\phi_{wire} = 3.217 \times 10^{-4}$

ϕwirer = the chosen wire diameter $\phi_{wirer} := 0.3 \cdot 10^{-3}$

Awr = the area of the wire $A_{wr} := \pi \cdot \left(\dfrac{\phi_{wirer}}{2}\right)^2$

Iturnave = average length of the one turn $l_{turnave} := 2 \cdot \left(6 \cdot 10^{-3} + 6 \cdot 10^{-3}\right) + 2 \cdot h_w$

$l_{turnave} = 0.028$

ρcu = the resistivity of the copper $\rho_{cu} := 1.68 \cdot 10^{-8}$

Rw = the DC-resistance of the inductor $R_w := \rho_{cu} \cdot \dfrac{N_r \cdot l_{turnave}}{A_{wr}}$ $\quad R_w = 0.623$

Λ = permeance $\Lambda := \dfrac{L}{N_r^{\,2}}$ $\quad \Lambda = 4.625 \times 10^{-8}$

The calculated DC-resistance of the inductor is 0,623 Ω, and this is confirmed by measurements. The measurement is done by applying a voltage across the inductor and measuring the current through it. The resistance can then be calculated with Ohm's law. Measurements and result are in the Table 1. As seen, the calculated and measured results are close to the each other.

Table 1. *The DC-resistance of the inductor.*

DC-current (A)	DC-voltage (V)	Rs (Ω)
0,2538	0,1693	0,667061

Inductor core saturation measurements are in the Table 2. As seen, the choice of the inductor maximum current is correct and the inductor is suitable for the buck converter whose output current is 700 mA. In Figure 1 is the saturation curve.

Table 2. *The inductor core saturation measurements.*

Idc[A]	Ltot[mH]		Idc[A]	Ltot[mH]		Idc[A]	Ltot[mH]
0	0,396		0,7	0,393		1,4	0,184
0,099	0,396		0,8	0,39		1,5	0,135
0,199	0,396		0,9	0,386		1,6	0,106
0,299	0,396		1	0,379		1,7	0,089
0,4	0,396		1,1	0,366		1,8	0,078
0,5	0,395		1,2	0,336		1,9	0,07
0,6	0,394		1,3	0,271		2	0,065

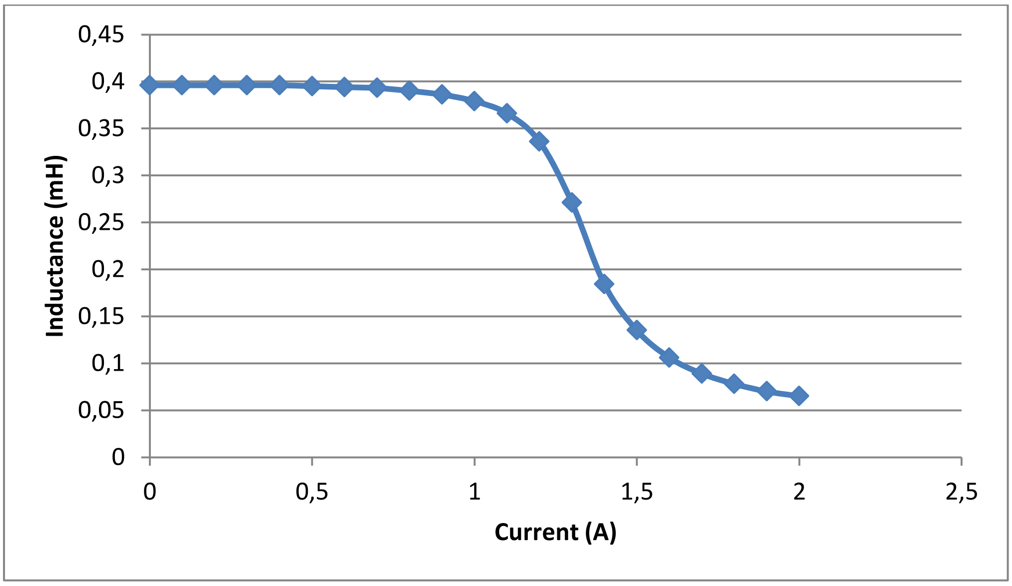

Figure 1. *The inductor saturation curve.*

The parasitic capacitance is also measured. This measurement is done without air gap. Therefore the inductance of the inductor is higher and the parallel resonance frequency is lower. Measurements and result are in the Table 2. 15 pF seems to be correct.

Table 2. *The parasitic capacitance of the inductor*.

Inductance (mH)	Parallel resonance frequency (kHz)	Parasitic capacitance (pF)
5,5	550	15

APPENDIX 6: The Schematic of The Prototype Device

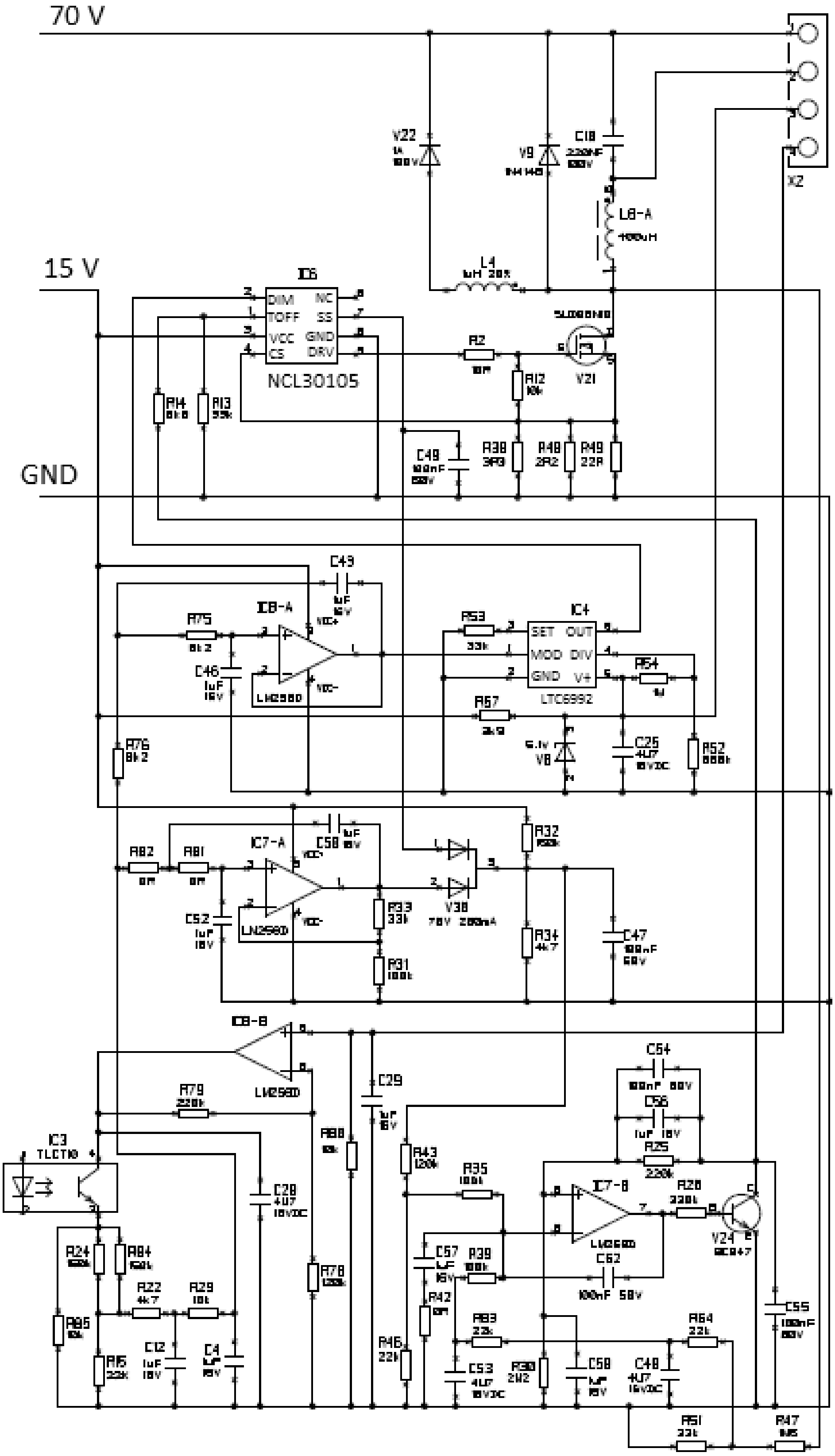

Figure 1. *The Schematic*

APPENDIX 7: Results of The Measurements

Table 1. *Efficiency measurements when output current is 700 mA.*

LEDs	Pin 15V (W)	Pin (W)	Pin tot (W)	Pout (W)		η	Ploss
12	0,36	33,7	34,06	32,1		94,2 %	1,6
11	0,38	31,1	31,48	29,4		93,4 %	1,7
10	0,39	28,7	29,09	26,9		92,5 %	1,8
9	0,4	26,1	26,50	24,4		92,1 %	1,7
8	0,44	24,0	24,44	22,1		90,4 %	1,9
7	0,46	21,5	21,96	19,5		88,8 %	2,0
6	0,49	19,0	19,49	16,9		86,7 %	2,1
5	0,5	16,5	17,00	14,3		84,1 %	2,2

Table 2. *Efficiency measurements when signal diode and ferrite is used.*

LEDs	Pin 15V (W)	Pin (W)	Pin tot (W)	Pout (W)		η	Ploss
12	0,29	33,7	33,99	32,1		94,4 %	1,6
10	0,29	28,8	29,09	27,1		93,2 %	1,7
8	0,3	23,7	24,00	22,0		91,7 %	1,7
6	0,3	18,6	18,90	16,8		88,9 %	1,8

Table 3. *Efficiency measurements when output current is 350 mA.*

LEDs	Pin 15V (W)	Pin (W)	Pin tot (W)	Pout (W)		η	Ploss
16	0,29	18,7	18,99	18,3		96,4 %	0,7
14	0,28	16,8	17,08	16,3		95,4 %	0,8
12	0,28	15	15,28	14,4		94,2 %	0,9
10	0,29	13,1	13,39	12,5		93,4 %	0,9
8	0,29	11	11,29	10,3		91,2 %	1,0

Table 4. *Operating point measurement when output current is 700 mA.*

Uout (V)	Iout (mA)	Operation frequency (kHz)	Inductor current ripple (mA)	Output current ripple (mA)
46,2	705	173	220	85
42,5	708	192	204	77
35,6	718	222	184	72
31,7	724	235	160	67
28,3	729	243	150	62
25,1	736	250	141	60

Table 5. *Operating point measurements when output current is 350 mA.*

Uout (V)	Iout (mA)	Operation frequency (kHz)	Inductor current ripple (mA)	Output current ripple (mA)
52,5	352	171	173	80
46,2	356	226	157	68
39,8	363	277	145	56
32,8	374	316	121	25

Table 6. *Frequency increase during analog dimming.*

Iout (%)	Operation frequency (kHz)
100,00 %	208
90,45 %	223
79,25 %	239
70,30 %	252
59,70 %	274
49,70 %	297
43,28 %	318
33,73 %	319

Table 7. *The dimming curve measurements.*

Control voltage (V)	Relative Iout		Control voltage (V)	Relative Iout		Control voltage (V)	Relative Iout
10	100,00 %		5,1	43,13 %		3,2	9,93 %
8,5	100,00 %		5	39,86 %		3	8,30 %
8	94,56 %		4,8	34,69 %		2,8	6,94 %
7	74,83 %		4,5	27,21 %		2,5	5,31 %
6	57,41 %		4,2	21,77 %		2,2	3,54 %
5,3	46,67 %		4	19,46 %		2	2,86 %
5,2	45,85 %		3,8	16,33 %		1,8	1,90 %
5,15	43,54 %		3,5	12,38 %		1,5	0,82 %