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Design of a Dynamic Hybrid LED Driver

MASTER'S THESIS

to achieve the university degree of

Diplom-Ingenieur

Master's degree programme: Electrical Engineering

submitted to

Graz University of Technology

Supervisor

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Institute of Electronics

Electronic systems

Graz, February 2021

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Abstract

This thesis covers the development of an improved mid-power 48V DC LED driver. The improvements over traditional switch-mode LED drivers aim towards improved current stability and load regulation for deep dimming scenarios for a wide range of load conditions while maintaining an acceptable efficiency.

The approach is based on hybrid designs, which comprise of a switch-mode element as well a linear element. SPICE simulations are performed prior to measurements on the hardware prototype to verify the function and to improve its properties.

After a literature study and an analysis of the traditional reference driver, a parallel and series configuration was simulated. Since the series configuration showed significant advantages in terms of dimming and flexibility, this topology was selected for further optimization. The linear regulation stage as the most inefficient part of the system was improved for efficiency, stability, dimming behaviour. The dynamic voltage control of the device was then implemented by a link between linear stage and switch-mode stage.

The findings are a highly improved load regulation traded for an expected decrease in efficiency. Furthermore, the developed solution offers flexibility in terms of an adaptable number of LED channels while eliminating the need of additional switch-mode converters.

Abstract

Diese Masterarbeit behandelt die Entwicklung eines verbesserten 48V mid-power LED–Treibers. Die Verbesserungen gegenüber traditionellen LED-Treibern auf Basis eines Schaltreglers zielen auf eine verbesserte Stabilität des Ausgangsstroms und Lastregulation in weit heruntergedimmten Situationen ab, wobei eine akzeptable Effizienz erhalten bleibt.

Der Ansatz beruht auf hybriden Konfigurationen, welche sowohl Schaltelemente als auch linear regelnde Elemente beinhalten. Um die Funktion zu verifizieren und Optimierungen durchzuführen, wurden SPICE-Simulationen und anschließend Messungen an Hardware-Prototypen vorgenommen.

Nach einer Literaturrecherche und Analyse eines Referenztreibers mit traditioneller Technologie, wurde eine parallele und eine serielle hybride Konfiguration simuliert. Da die serielle Konfiguration Vorteile bei der Dimmbarkeit und Flexibilität gezeigt hat, wurde dieser Ansatz für eine weitergehende Optimierung gewählt. Die Linearstufe, als ineffizientester Teil, wurde in seiner Effizienz, Stabilität und Dimmverhalten verbessert. Die dynamische Spannungsanpassung wurde als Koppelung der Linearstufe mit der Schaltstufe realisiert.

Das Ergebnis ist ein LED-Treiber mit stark verbesserter Lastregulation, was durch erwartete Zugeständnisse bei der Effizienz erreicht wurde. Weiters bietet die entwickelte Lösung zusätzliche Flexibilität bei der Anzahl der LED-Ausgangskanäle ohne weitere Schaltregler zu benötigen.

Contents

Chapter 1

Introduction

1.1 Motivation

The lighting business has been growing a lot for the last decades. Especially the evolution towards LED lighting has been an accelerator of this business. Advertising the improved energy efficiency, governments all over the world have started to ban old fashioned incandescent light bulbs, which increased the need for a modern replacement. [1]

LED lighting is a huge industry and the need for energy efficient light is as high as never before. Recent developments like human centric lighting (HCL) and the integration in building management systems (BMS) have made smart lighting become reality.

Also the latest acquisitions like OSRAM being bought by ams AG show that there is a high potential in the market [2]. This is not limited to the residential light bulb segment but to a whole range of business segments as well.

Together with the rise of the efficient LED technology came the need of more sophisticated supply to power this type of device. Unlike incandescent bulbs, LED's cannot be directly supplied from the mains grid. LED's behave like diodes while their luminous flux is proportional to the current through them. Therefore, a current regulation is needed, which is done by an electronic device, the driver.

Although the field of energy conversion and current regulation for LED's has been researched for many years already, introducing new approaches and sophisticated circuits can have a massive impact on the market. Nowadays, not just the efficiency but rather a high quality of the output current is desired to create the best possible light output.

The thesis is a result of the collaboration of the Institute for Electronics (IFE) at the Graz Technical University and the XAL GmbH, a world-leading manufacturer of high quality luminaires.

The goal is to design an LED driver for their $48V$ DC track system that is efficient, capable of precise dimming and dynamically adapting to different load configurations. This includes analyzing the currently used DC-DC LED driver and to introduce a new generation of hybrid converters that unites the advantages of two different conversion principles. It must comply with all regulations and has to be small, reliable and reasonably cheap.

It is already known, that the current solution has issues with the load regulation, which results in a dependency of the luminous flux to the forward voltage of the LED string. This behaviour results in adapting the output for each configuration. To avoid this, a lower efficiency can be tolerated.

The project is part of a product development, which means that the design may have the chance to be integrated in high quality luminaires.

1.2 Prior Art

Electronic circuits, intended to regulated the current through an LED and thus regulate their emitted luminous flux have been around for many years. The devices, known as LED drivers have a stage for regulating the current and various circuitry to ensure safe operation or advanced control mechanisms. When LED drivers are supposed to be connected to the AC grid, rectification and power factor correction (PFC) stages are implemented as well.

For regulating the current, two main principles are dominating. The *linear regulation*, using the resistive properties of a series pass device, and the *switch-mode* conversion, relying on the averaging of a pulsed maximum current. Both methods have their own advantages and disadvantages. Linear regulators provide a ripple-free output and very good load regulation while being rather inefficient. Switch-mode converters are very efficient but suffer from bad regulation and can have stability issues.

In order to combine the advantages while cancelling out the disadvantages, it has been tried to combine linear and switch-mode operation into a hybrid driver. A first patent using this technique was applied in 2001 [3]. In this concept, the linear regulator is connected in series to the switch-mode converter to control the current, while the switch-mode converter regulated the voltage. A few years later, this technique has become more popular. A paper from 2006 suggests the use of a Darlington device as the series pass element [4].

A very sophisticated hybrid converter was presented by Hu et. al in 2008, where a circuit was described that regulates the output voltage of the switch-mode converter to the minimum required voltages, needed by the LED and the linear regulation stage. This dynamic voltage output reduces unnecessary losses of the linear regulator and thus, improves the efficiency. In addition to this, the circuit is capable of multiple LED strings, also called channels, by selecting the string with the highest voltage drop accross the LED. This ensures that the voltage is high enough for all LED strings although decreasing the efficiency of the channels with lower forward voltage [5] [6].

A new approach on hybrid LED drivers was given by Garcia, who was researching linear-assisted DC/DC converters. The idea is based of connecting switch-mode and linear regulator in a parallel manner. The switch-mode regulator is set up to provide most of the current while the linear regulator is providing the remaining current. Since linear regulators can have a much faster regulation, the provided current will vary, depending on the ripple coming from the switch-mode converter and thus, actively filter the current. Different control theorems have been provided for current control as well as for voltage control [7].

1.3 Thesis Structure

The scope of this thesis is the investigation if a hybrid LED driver, comprised of switching and linear elements is able to provide significant improvements to a pure switch-mode solution.

The first chapter is providing a background of LED driver technologies and is based on literature research. The chapter is giving a good overview on the properties and characteristics especially for DC to DC LED drivers.

The second chapter goes deeper into the technological aspects of DC to DC regulation, which is the heart of an LED driver. Different methods for this regulation are discussed and a first introduction about the theoretical approach and the publications on hybrid DC to DC regulators are explained.

The third chapter is an analysis of an existing switch-mode LED driver. The chosen LED driver is currently used in the market which makes it a good reference point for the later comparison with a hybrid solution. This chapter contains schematics, measurements and their interpretation of the mentioned driver.

The fourth chapter contains the whole design phase of the hybrid LED driver. Two hybrid concepts are simulated and the results as well as difficulties with the designs are discussed. The best concept is chosen and optimized. This optimization is done in steps, where the first step is an optimization of the separate stages. Here, the main focus lies on the linear stage, as this was found to be a major factor for the efficiency. In the next step, both stages are connected and the overall efficiency is optimized by a sophisticated feedback network. The optimizations made are based on simulations and measurements. As a last step, possible fault conditions are discussed and detection mechanisms are presented.

Chapter 2

LED Driver Technologies

This chapter will give the reader an overview of LED drivers as a system. First, different types and configurations are presented, which cover different needs of input and output signals. Then, common criteria for the quality of LED drivers are discussed as well as the legal requirements for these devices in the European union. Last, smart lighting is discussed, since it is playing a more and more important role for modern devices.

2.1 Applications

LED lighting involves more than just home applications and each of them have their own challenges and developments. Common market segments for LED lighting are: [8, pp. 52–68]

- Residential
- Office
- Retail and Shops
- Industrial
- Automotive Headlights
- Outdoor

- Horticulture
- Backlighting
- Medical

It should be noted that this is only one way of segmenting the market. Furthermore it can be divided in geographical segments, in power demand or in customer groups (eg. consumer or public).

The *residential* market is probably the most known market since everyone in the modern world has artificial light sources installed at home. The challenge for this market are extremely low costs, since the competition is huge. But also design and quality of the light is becoming more and more important.

The market around *offices* is dominated by a high quality of light to ensure a good working atmosphere for the employees. Also a bus connection to manage the fixtures remotely is common here. In this market, *human centric lighting* is becoming more important as well. Its purpose is the emotional influence on humans, which can be done by controling the color of the light to resemble the natural daylight or boosting cold white after lunch to wake employees up. There are also special regulations in place that determine a minimum illumination on the desk as well as maximum glare ratings to reduce unpleasant direct or reflected glare (UGR) [9].

The *retail or commercial* market is extremely focused on the quality and reliability of light. This includes the best color representation, indicated by the color rendering index (CRI) which goes from 0 to 100. Special colors have been developed for specific applications like food lighting, which boosts the colors of vegetables or fish and creates a look of freshness. Additionally this market requires special form factors so that it can fit in a glass cabinet or can easily be installed in a track system. Since these fixture usually run all year, the efficiency and reliability are considered as well.

In an *industrial* market, it is all about light output and ruggedness. The quality of the light as well as design are playing a minor role. Improved ingress protection, marked by the IP rating are often required.

The *automotive* market has its own rules. Long periods of certification and testing are a market barrier for small businesses. This market has the highest requirements, where everything is a tight fit.

Outdoor as well as *horticulture* fixtures are characterized by their high IP ratings. With expansion and compression of the enclosed air by heating and cooling of the LED and the driver, it is hard to keep moisture from creeping in.

A market segment that has developed to a huge volume in tha last years is *backlighting*. Since every LCD display has to be illuminated from behind, the sales in this segments are rising fast. Challenging is the quality of light, which has to provide a well defined color, even illumination and no unpleasant flicker. Also the form factor is tiny to fit inside the shrinking display thicknesses. Efficiency is especially for portable devices a concern.

A segment that has become more popular with the latest pandemic of SARS-CoV2 is the *medical* segment. This includes sterilization devices based on UV as well as lighting for hospitals and surgical lights. Like the automotive market, it is dominated by long certification processes and high quality of the light.

2.2 Quality of Light

Since the trend of new artificial light sources is still ongoing, there are several measures to check how the light quality of the next generation compares [10]. The *spectral quality* of light implies the perceived color of the light. There are different systems in place to describe the colorimetry of light, such as xy coordinates, u'v' coordinates or correlated color temperature (CCT) for white light. Furthermore, there are systems for measuring how the color of enlighted objects differs. The most popular system is the color rendering index (CRI) as well as more modern systems like CIE 224 and IES TM-30 which rate the fidelity in more detail. The lack of red light in the LED spectrum, measure by the R_9 value of the CRI system, was often criticized. But a study has shown that the CRI value does not affect the attractiveness of objects, but can affect the impressions like the freshness of food [11] [12]

[13] [14] [15].

Another quality of light is the *temporal quality*, which describes time related properties of light. Flicker, which is a change in illumination in the range of $1Hz$ to $24kHz$ can cause discomfort in humans and animals. This includes power oscillations in fluorescent tubes, which usually have $2f_{mains} = 100Hz$ as well as higher frequency switching of modern LED light sources. Flicker becomes more dominant in dimming operation when pulse width modulated (PWM) dimming for LEDs or phase-cut dimming for conventional light sources is used. Since LEDs are very fast devices, PWM dimming is common and can be done with high frequencies that are not detected by the human eye but may be detected by animals or electronic cameras [11].

IEEE 1789 rates the health concerns of flicker based on the frequency and modulation and divides it into 3 categories:

- Visually perceptible
- Flicker perceived unconsciously
- Strobe effects

Visually perceptible flicker is in the range of $3Hz$ to $70Hz$. In this frequency region, flicker can cause epileptic seizures and headache. Unconscious flicker with up to $200Hz$ is not immediately recognized but can have effects like headache, fatigue, problems in reading and general. The effects of strobe lights with up to $2.5kHz$ becomes dangerous when rotating tools are illuminated so that the actual speed can not be determined by humans anymore [16].

Figure 2.1 shows the recommended flicker modulation by IEEE 1789. The modulation is defined as the difference between minimum and maximum intensity. A PWM operation with 100% modulation depth therefore must be at a frequency higher than $12kHz$ [16].

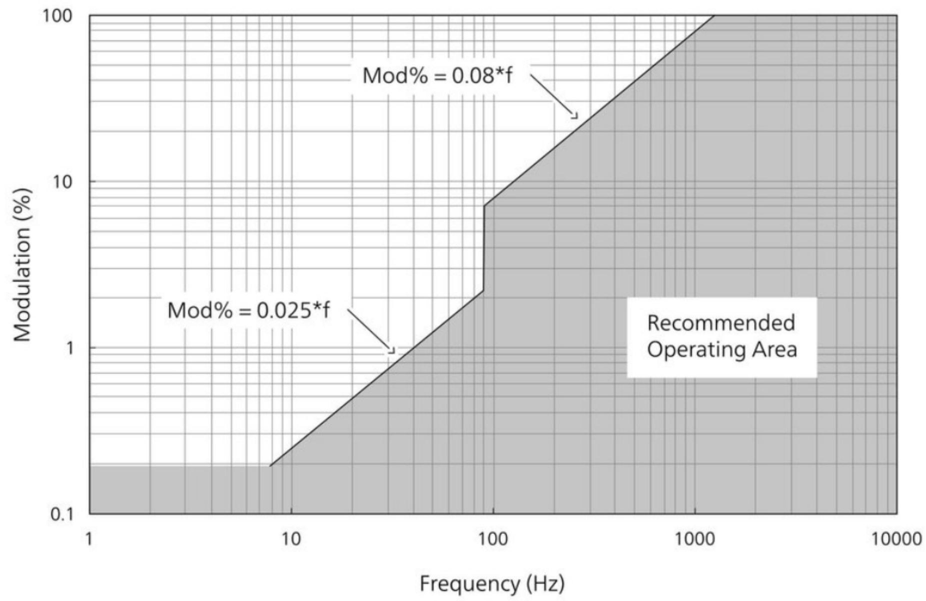


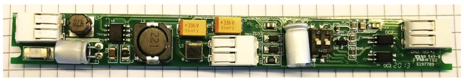
Figure 2.1: Allowable flicker according to IEEE 1789 [16]

2.3 Categorization of LED drivers

The term LED driver is widely used for any current source or regulator. This sometimes causes problems, since the applications and capabilities can be very different. In figure 2.2, two examples of different LED driver types can be seen. Figure 2.2a shows an LED driver for $230V$ while figure 2.2b shows an LED driver that is made for a 48V DC safety extra low voltage (SELV) input. It is easy to see, that DC to DC driver can be much smaller. That is due to the fact that a lot of EMI filtering happens at the AC to DC conversion stage. This means, a DC to DC driver will always operate together with a DC power supply. Furthermore, the SELV voltage requires very small creepage distances compared to 230V grid voltages.

A greatly simplified diagram of an AC LED driver can be seen in figure 2.3. The $230V$ input is first routed through filters to block interferences coupling into the line voltage. Then the voltage is rectified and passed through a power factor correction, which smoothes out the high current peaks drawn from the switch-mode power supply behind it. Parallel to the power converter

(a) Tridonic LED Driver



(b) HEP 48V DALI

Figure 2.2: Examples of different LED driver types

itself, the control circuit is powered with a low voltage.

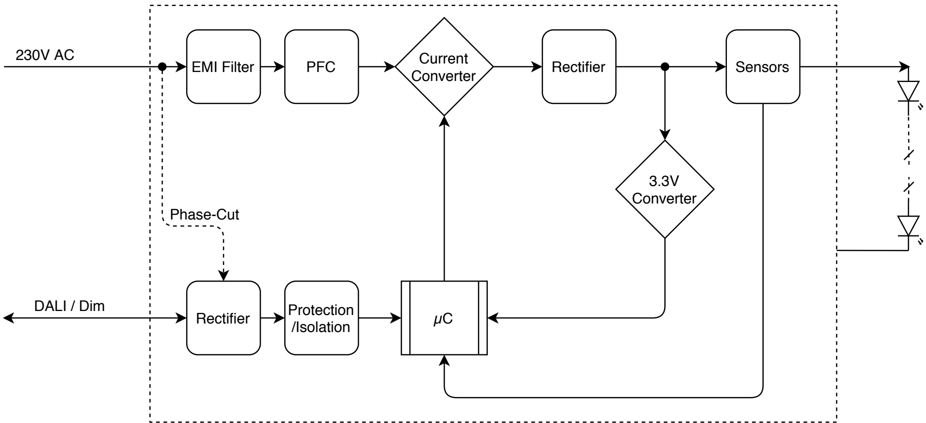


Figure 2.3: Simplified block diagram of a typical 230V LED converter

2.4 Properties of LED drivers

This chapter will discuss the features and characteristics that can define an LED driver. This will also act as an overview of what to look for in an LED driver, which can be helpful for choosing or defining a suitable driver for a specific application.

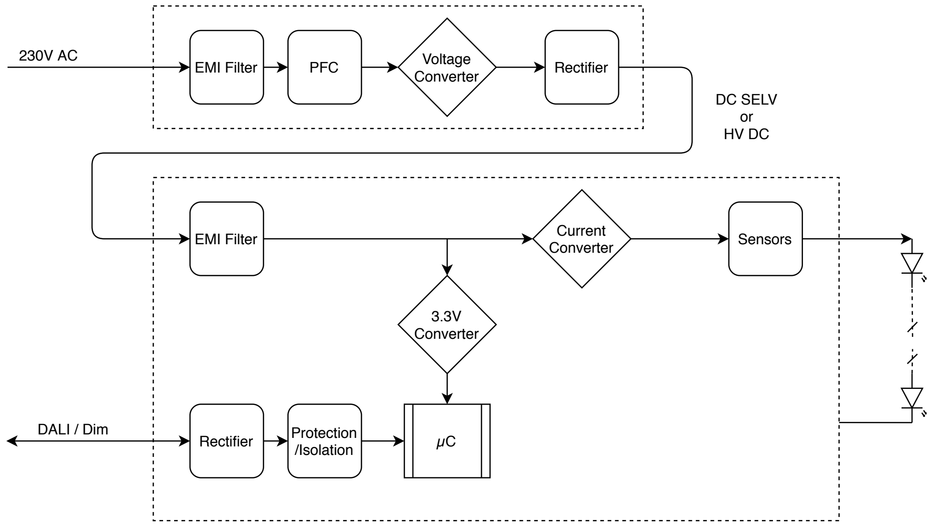


Figure 2.4: Simplified block diagram of a DC-DC LED converter

2.4.1 Characteristics

In the beginning of this chapter, it was shown that LED drivers can come in various forms. The input signal has a large influence on the requirements of the LED driver. Between AC and DC input signals, the difference is so large, that there is little literature covering both topologies in detail.

Based on a survey, Li et al have segmented commercially available AC LED driver into power segments. The low-power segment with up to $10W$ is usually considered for interior lighting such as retrofit LED bulbs. The mid-power drivers are considered up to $50W$ and have most applications in indoor as well as outdoor fixtures. High-power drivers above $100W$ are typically chosen for outdoor, streetlight or floodlight applications [17].

Since this thesis focuses on DC drivers, the characteristics in table 2.1 represent a summary of many sources Most of the characteristics for voltage regulators can be adapted and used for current regulators as well. It has to be noted, that the implementation or improvement of any feature will always be contrasted to size and cost [18] [19] [20].

The efficiency and signal integrity section represent the current regulation itself, while the other sections are related to periphery. The input and output

General characteristics	Electrical characteristics
Voltage ranges	Input voltage range Output voltage range Output current range
Efficiency	Dropout Quiescent current
Signal integrity	Load regulation Line regulation Power supply rejection ration (PSRR) Transient signals Noise Ripple
Internal protection	Overvoltage protection Undervoltage protection Short circuit protection Overtemperature protection
External protection	SELV isolation ESD protection EMI protection and filter HV protection Hotplug protection

Table 2.1: Typical characteristics of a DC-DC converter

voltage range is an important factor for an appropriate selection of any DC-DC regulator. Most DC systems are designed for a specific input voltage rail. Common voltage levels include $3.3V$, $5V$, $12V$, $24V$ and $48V$. These voltage rails have their origin in active components, automotive applications and batteries and are usually limited by the SELV regulation which limits the voltage to $60V$. Nowadays, the popular $12V$ systems are replaced by $48V$ since the resistive losses, given by $P_{loss} = I^2R$ are reduced.

Efficiency is important when it comes to high powers, where the heat produced by the losses is limiting the available power, but also in new energy regulations that have the objective to keep the standby power below a certain limit [21].

The considerations about the signal depends on the requirements of the load. Highly sensitive measurement circuits will require a much better supply signal. LED's require a low ripple current because flicker, as discussed in chapter 2.2, can be a health concern. The load regulation for current sources is defined as the change of the output current for a given change of the output voltage as described in equation 2.1. A good load regulation is needed in an LED driver, especially when two strings of LED's with different lengths are visible at the same time. A bad load regulation will cause one string to be brighter than the other.

$$\%LoadRegulation = \frac{\Delta I_{out}}{\Delta V_{out}} \tag{2.1}$$

The line regulation is the static change of the output current for a given change of the input voltage, according to equation 2.2. For a single DC rail, this is not as critical but may have effects when multiple LED drivers are connected to different voltage rails.

$$\%LineRegulation = \frac{\Delta I_{out}}{\Delta V_{in}} \tag{2.2}$$

The power supply rejection ration (PSRR) is a dynamic dimension and describes the attenuation of an input ripple to the output signal, which is plotted for a range of different frequencies. The PSRR can be calculated with equation 2.3. For voltage to voltage converters, this is not difficult to define, whereas for voltage to current conversion, the input voltage ripple will still be a voltage ripple which partially influences the current based on the transconductive properties of the converter an the load.

$$PSRR = 20log\frac{V_{Ripple,in}}{I_{Ripple,out}} \tag{2.3}$$

Transient signals are referring to overshot, undershoot as well as the rise time and fall time of the output signal when the target current is suddenly changed. The rise and fall times are hereby usually defined between 10% and 90% of the signal. For LED drivers, the rise times and overshoot at startup are especially important because sudden spikes can damage other components in the system.

Noise and ripple are disturbances on the output signal introduced by the converter. Especially switching converter, that are described later, can create spikes with every switching operation. These disturbances can create flicker, glowing, and EMI issues.

Internal protection mechanisms, which are not necessarily caused by an outside event are measure to protect the hardware in case of a short circuit of the load, a temperature that can cause failures or shortens the lifetime as well as non-critical over- or undervoltage events.

External events, such as electrostatic discharge (ESD) or hotplug events are usually much more energetic and can lead to the failure of the device or even cause danger to the environment. An ESD event is cause by a charged human or animal, touching the device. The resulting arc and voltage spikes can be above $8kV$ and can easily damage integrated components. Typical countermeasures are transient voltage suppressor (TVS) diodes. Hotplug is the connection to an active power rail. In this moment, all capacitances of the circuit are sinking current, generating a high current and voltage peak, caused by the inductance of the wires. This energetic power surge can also harm components and cause failures. Countermeasures are dedicated input circuits, designed to slowly start conducting and choosing components with high voltage capabilities.

2.4.2 Legal Regulations

Lighting products have to fulfill a number of legal requirements. These usually depend on the application, as automotive requirements might be completely different than the requirements for office fixtures. Most of these norms are in place to guarantee the safety of the user but also the safety of the environment. These regulations can be differentiated in mechanical, optical and electrical norms. This section will focus on the electrical norms that are currently in place within the European Union for residential and office luminaires. Other countries like the United States have their own, sometimes unique, regulations.

The electrical regulations for lighting devices are focusing on safety classes with different creepage distances, EMI and HV tests, as well as protection against ESD.

The first important regulation is the EN60598. This is the lighting specific standard that covers most electrical safety and protection requirements. The following list is a summary of the most important electrical regulations in EN60598-1 [22].

Chapter 3

DC Current Regulation

This chapter will review the basics of current regulation. First, the principle of linear current regulation is explained. Then switching principles with different topologies and control methods are discussed. In the last part, approaches for hybrid designs which are a combination of switching and linear current regulation are analyzed.

3.1 Linear Current Regulation

The most basic way of setting the current in a DC system is by placing a resistor

The basic idea of limiting the current by a resistor can be improved by using an adjustable resistor which can be implemented by a transistor in its ohmic region. This is a simple linear regulator, regulating the voltage to a reference. The typical circuit can be seen in figure 3.1. Here, a reference voltage is applied to the non-inverting input of an operational amplifier. The inverting input connected to a shunt resistor, referenced to ground. This voltage is proportional to the current through the shunt resistor. As long as this voltage is lower than the reference voltage, the operational amplifier will increase the output voltage. Since the output is connected to a transistor, together with a resistor to limit the base current, the base current and thus the collector-emitter current will rise, increasing the current through

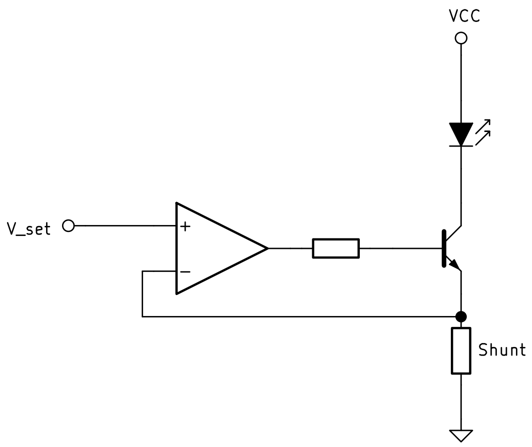


Figure 3.1: Simplified schematic of a linear current regulator

the shunt resistor. This closed loop will regulate so that the voltage over the shunt resistor is equal to the reference voltage. In a steady-state condition, the current can be calculated by equation 3.2. This circuit is a simple current sink, linked to ground potential. The ground reference has a huge advantage in the current measurement. This is because the shunt as well as the operational amplifier and the reference voltage have ground as the same reference potential[23, pp. 228–229] [24, pp. 13 – 31].

Of course this is an idealized circuit and many other parameters have to be considered.

The important parameters that we analysed in chapter 2.4.1 can be studied here. The voltage and current range is determined by the transistor that is used. Since the regulator is on the low-side of the load, the voltage drop across the load has to be considered. The safe operating area of the transistor which is typically defined in the datasheet has to be applied. Especially for bipolar transistors, a second breakdown can damage the device. In most cases, the maximum current is not limited by the current capability of the transistor but by the heat generated by the losses. To determine the maximum current, a temperature test of the device has to be performed in its final housing and condition. The efficiency can be calculated with equation 3.1. The linear current regulator has losses that are proportional to the current. Therefore, a high load voltage has the advantage of a larger power at lower currents. The signal integrity parameters of an idealized linear current regulator are perfect. In reality, it highly depends on the characteristics of the operational amplifier and series pass device.

$$\eta = \frac{P_{out}}{P_{in}} = \frac{P_{out}}{P_{out} + P_{loss}} = \frac{P_{out}}{P_{out} + I * U_{CE} + I^2 * R_{shunt}} \tag{3.1}$$

$$I_{LED} = I_{shunt} = \frac{U_{shunt}}{R_{shunt}} = \frac{U_{ref}}{R_{shunt}} \tag{3.2}$$

$$P_{shunt} = U_{shunt} * I_{shunt} = R_{shunt} * I^2_{shunt} \tag{3.3}$$

$$\beta_T = \frac{I_{ges}}{I_b} = \frac{I_{ges}}{\frac{(V_b - V_{be})}{R_{shunt}}} \tag{3.4}$$

3.2 Switched Current Regulation

The second, very important conversion principle for current regulation is the switch-mode conversion. The extraordinary number of published papers and books on this topic shows the huge interest. As the name suggests, this principle relies on a combination of switching elements and energy storage together with a sophisticated control method. There is a huge variety of switch-mode converter topologies. This chapter will only take a glimpse into a single category of switch-mode converter topologies, namely non-isolated, single inductor topologies. These topologies include buck, boost and buck-boost converters typically found as voltage converters.

Buck Topology

The buck topology that you can see in figure 3.2 is able to convert to lower voltages. While the switch S1 is closed, current starts flowing through the inductor to the load. The inductor is limiting the initial current flowing. When the switch S1 is opened, the magnetic field of the inductor collapses, creating a current with the same direction as the current before. The voltage at the inductor reverses, making it a source and enabling the diode D1 to

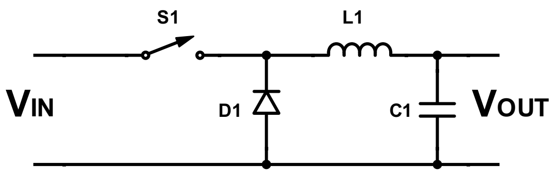


Figure 3.2: Simplified schematic of a buck type converter

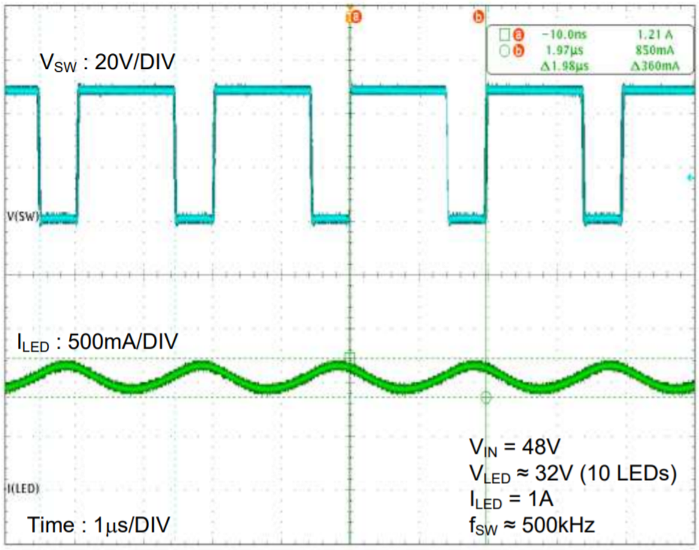


Figure 3.3: Switching waveforms of the TPS92640 LED buck converter [25]

conduct, closing the loop. Key of this topology is the exact control technique [23].

The buck principle can be applied to voltage and current regulation. Figure 3.3 shows the waveform of the LED buck converter $TPS92640$. The switched current is effectively smoothed out by the inductor. For even better control and efficiency, the $TPS92640$ is a synchronous buck converter, which means that the diode D1 is replace by a MOSFET device. This removes the voltage drop V_f across the diode, thus improving efficiency [25].

3.3 Hybrid Designs

As discussed in chapter 1, linear regulators struggle with their efficiency at high ratios of $\frac{V_{in}}{V_{out}}$. The voltage difference is dropped across the transistor resulting in high losses generating heat. On the other hand, linear regulation, if designed properly, provides a ripple-free output signal with high stability and no EMI issues. Furthermore, less and cheaper parts can be used compared to a switch-mode regulator. Therefore, linear regulators are typically used in low current applications or in arrays of LED strings, especially for large surface LED PCBs.

Switch-mode technology can be very efficient but struggles at low currents because it can have severe ripple and EMI issues. More complex switch-mode solutions, like multi inductor topologies or resonant converters can make up for these flaws but inherit more parts that come with considerably higher costs.

One might notice that some weaknesses of one conversion principle are the strengths of the other one. Combining a switch-mode converter and a linear regulator could lead to an efficient converter with a precise current regulation and ripple-free output. This combination is here called hybrid converter.

Combining the two principles is possible in two possible ways, the parallel and the serial configuration. The parallel approach (Figure 3.4) will be explained and simulated and the results are compared to the existing solutions from chapter 6. Afterwards, the serial approach is split into two parts as there are two possible configurations. The first part will consist the linear regulation on the high-side of the load, where as the second part will implement it on the low side of the load. The advantages and flaws of these concepts be will discussed afterwards.

3.3.1 Parallel Hybrid Conversion

One possibility to connect the two regulator concept is the parallel configuration that can be seen in figure 3.4. The idea behind this approach is that the efficient switch-mode converter provides most of the current, while the

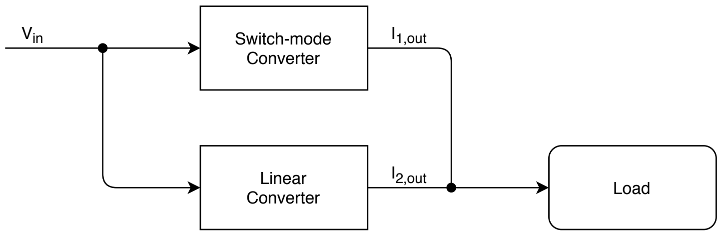


Figure 3.4: Block diagram of a hybrid converter in parallel configuration

linear current regulator is providing a small current to cancel out the current ripple of the switch-mode part.

H. Garcia has published a series of papers focused on "Linear-Assisted DC-DC Voltage Regulators". In these papers, he describes a voltage regulating circuit with a linear regulator interacting with a switch-mode converter in buck configuration. In theory the output will have no ripple as long as the linear device is faster than the slope of the switching ripple. One paper in particular is looking on this principle in the application of a LED driver. Figure 3.5a shows the circuit of a linear-assisted regulator. The linear regulator is wired in a configuration, as described in chapter 6. It comprises a feedback path to the output that in order to match a given reference voltage. The output current of the linear regulator is measured and fed back into the buck converter which adapts according to the output current of the linear regulator. The waveform in figure 3.5b shows what is happening at the startup of the circuit and helps to understand the working principle. It is crucial to begin with the linear regulator as this is the main device. It's reference voltage is set to regulate the output voltage to the desired potential. The buck converter is implemented in such a way that it wants to keep the output current of the linear device at a certain level. The speed of the regulating device plays an important role here, since the linear regulator must always be faster than the buck regulator. When the voltage of the output is lower than specified by the linear regulator circuit, the linear device will provide the necessary current to reach this output voltage. In many cases this current will be much higher than the buck regulator wants the current to be and starts to increase its duty cycle to provide current itself until the current through the linear device decreased. Since the buck converter will introduce a certain ripple to the output voltage, the linear regulator will try to cancel this ripple. The result is a very fast regulator where most of the current is provided by the buck converter and that regulates the output voltage with a linear regulator. For large load transients, the linear regulator will provide most of the current for a short amount of time until the buck converter can adjust it's output. This behaviour can be observed in the theoretical waveform in figure 3.5b [7].

But there a some flaws to this system. Firstly, the linear regulator will always source some current while being inefficient compared to the buck converter. As the linear regulator will only source current, the steady state current must be high enough to cancel out the upper half of the switching ripple. For system with a high ripple, the average current of the linear regulator I_γ can be high, which reduces the advantage of higher efficiency of the parallel topology. A strategy to to tackle this problem is called zero-average control. It features a linear regulation with a push-pull stage that is able to sink and source current. In steady state operation, the average current of the linear device can therefore become zero [26].

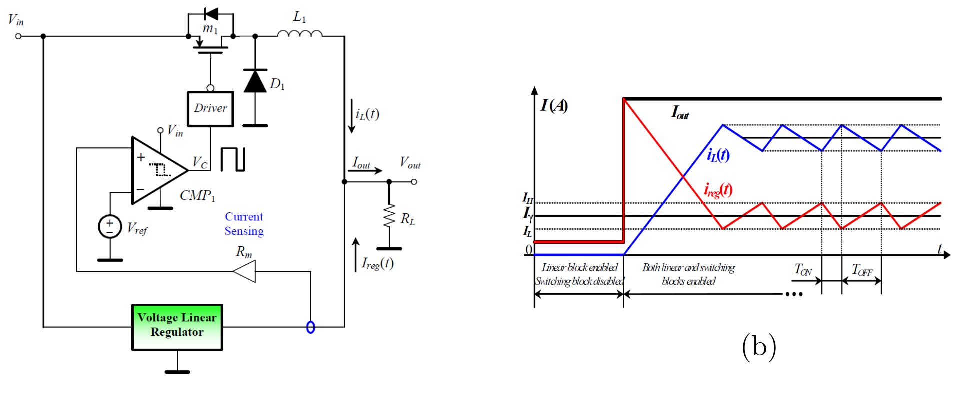
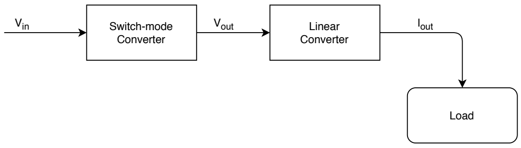
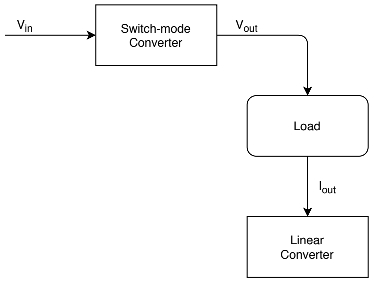


Figure 3.5: Schematic and simplified signals of a linear-assisted converter [7]



(a) High side linear stage



(b) Low side linear stage

Figure 3.6: Block diagram of a hybrid converter in serial configuration

3.3.2 Serial Hybrid Conversion

The second type of hybrid conversion is the serial combination of a switch-mode conversion stage and a linear regulator. This principle is well known and often referred to as *preregulated Linear regulator*. The system can be built in two ways in which the linear regulator is on the high side or the low side of the load, as you can see in figure 3.6.

The idea of these circuits is simple. The efficient switch-mode converter is set up to create an output voltage $V_{out,buck} = V_{load} + V_{dropout}$. This way, the voltage across the linear stage is minimized, thus minimizing the losses. Another advantage is that a single switch-mode converter can be used for many LED strings with individual linear current regulation. Drawback of this system is an efficiency loss compared to a pure switch-mode solution because of the losses in the linear regulator. Furthermore, this circuit requires to adjust the output voltage of the switching converter to the load voltage. Unfortunately, the forward voltage of an LED depends on many factors and can vary with current and temperature. This leads to a voltage headroom that has to be added on top of the dropout voltage $V_{out,buck} = V_{load,expected} + V_{dropout} + V_{headroom}$ which leads to additional losses.

To tackle this important drawback, tracking preregulation was introduced. The idea is to dynamically adapt the output voltage of the switching converter, so that it is always $V_{out,buck} = V_{load} + V_{dropout}$. Hu and Jovanovic prensent in their paper a tracking preregulator solution for LED driver with an improved feedback compared to other topologies and introduced control loop considerations. A major difficulty of these circuits is the tracking implementation. Especially for more than one current regulated LED string, the output voltage must be adapted to meet the requirements of the highest load voltage $V_{out,buck} = V_{load,max} + V_{dropout} + V_{headroom}$. This was previously done by an minimum voltage detection comprised of sensing diodes as you can see in figure 3.7. While this method always selects the lowest voltage drop, it is mentioned that it is highly dependant on the diodes, their temperature and the reference voltage of the switching converter. The lowest voltage drop over a linear regulator that can be achieved is $V_{dropout,min} = V_{ref} - V_{f,sense}$ [5].

In figure 3.8, the block diagram of the small signal control loop is presented. It can be noticed that the system consists of two control loops. The inner loop T_I is regulating the current, while the outer loop T_V only regulates the voltage. The inner loop is determined by the gain of the error amplifier $-G_{EA}$ which amplifies the voltage drop across ΔV_{RS} across the current sense or shunt resistor R_S. The amplified error signal then changes the current ΔI_{DS} through the MOSFET by the gain K_{GS} of the MOSFET. In outer loop T_V, the amplified error voltage is fed into a modulator, which is part of the switch-mode regulator and translates it into a changed duty cycle. Along with the transfer function G_{VD} of the switch-mode regulator, the voltage in the loop is regulated and errors from changing input voltage and changing forward voltage of the LED's is being compensated.

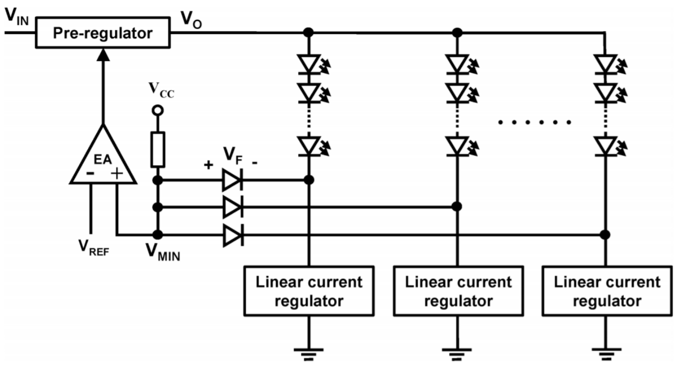


Figure 3.7: Schematic of a hybrid converter in serial configuration with load voltage detection [5]

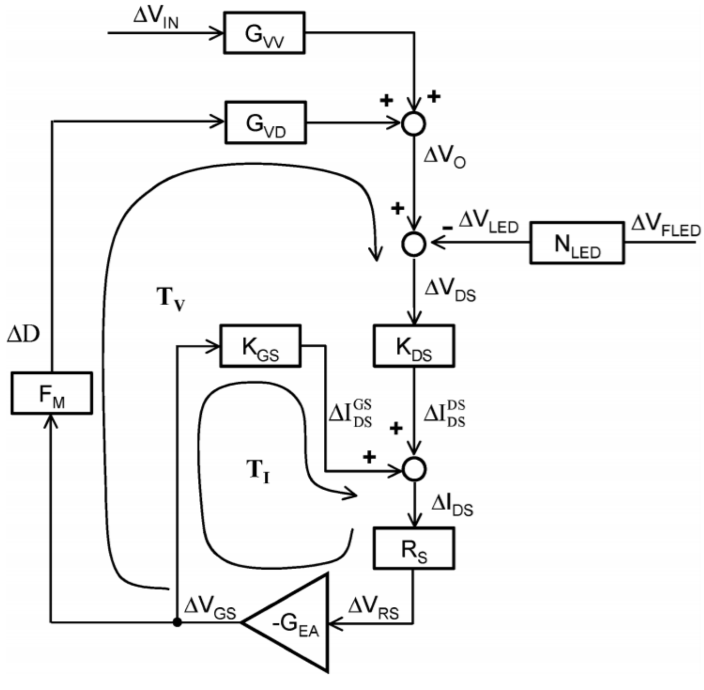


Figure 3.8: Block diagram of a hybrid converter in serial configuration [5]

Chapter 4

Reference Solution

When starting a design, it is a good idea to define a base line where to start. It would not make sense to start right out of the blue. Therefore a reference solution based on a switch-mode current regulator is used as example of a state-of-the art LED driver that is currently used in the industry. The design focuses on the driving part, so that the front end which communicates with the bus system will stay the same.

The schematic can be seen in figure 6.4. The reference solution is using the TPS92640 current regulator made by Texas Instruments. This DC-to-DC controller contains MOSFET drivers to operate with external switching devices. This gives the designer more freedom and can help to spread the heat which mainly occur in these switching elements. It operates as synchronous buck converter in current-mode operation. This, simplifies the required compensation circuit as we will see in chapter chapter 5 [25].

4.1 Characterization

To characterize the current solution, all important parameters that were discussed in chapter 6 are tested and measured. The parameter of the maximum current highly depends on the cooling and ways of thermal dissipation of the generated heat losses. Therefore, a temperature test is conducted to identify critical components but defining a maximum current can't be done without the heatsink and has to be defined for each application individually. An absolute maximum can only be defined by the maximum currents given in the datasheets of the individual components.

The maximum input voltage depends on the weakest component as well. Possible weak links can be the switch-mode controller TPS92640 and the linear regulator generating the $3.3V$ supply for the microcontroller.

The quiescent current of the driver was measured to be $4.1mA$ under no load.

In figure 4.1, the efficiency of the reference driver is plotted versus the load voltage for different load currents. The load used for these measurements is a series string of CREE LED's. Up to 12 LED's in series have been tested to determine the properties of the driver for different load situations. As expected, the efficiency increases with rising load voltage because the voltage difference that has to be converted decreases. The efficiency for higher currents is very good and reaches up to 92% at $34.2V$ with $467mA$. This implies a resulting loss of $1.34W$. For $995mA$, the efficiency also rises above 90% but reaches a loss of $3.43W$, which is already difficult to dissipate with the PCB as passive heatsink.

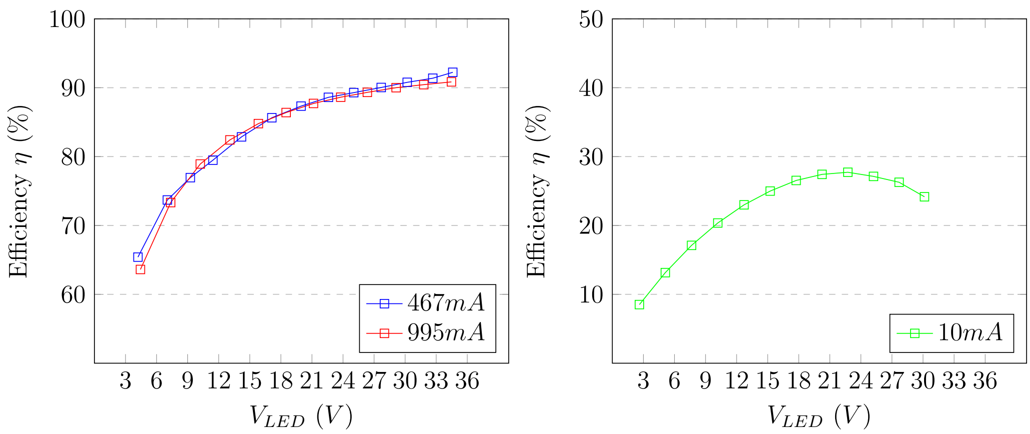


Figure 4.1: Efficiency across the output voltage range for different currents

As mentioned in chapter 2.4.1, the load regulation for a constant current converter is the current change in relation to a change in the output voltage.

In figure 4.2, the current for different output voltages was measured for a device set to $995mA$ and set to $10mA$, which is 1% of the maximum current. The graph shows that the current is dropping for rising output voltage. For $995mA$, the current drops a total of $15.4mA$, equivalent to 1.5% load regulation. For the low current of $10mA$, the current drops by $9.5mA$, which is -85.6%. Since LED current is very critical for low currents, because of the logarithmic sensitivity of the human eye, the bad load regulation is a severe disadvantage of the reference solution. The TPS92640 IC provides connections for a compensation circuit that can shift the current slightly for different voltages, but needs to be adjusted for every voltage individually.

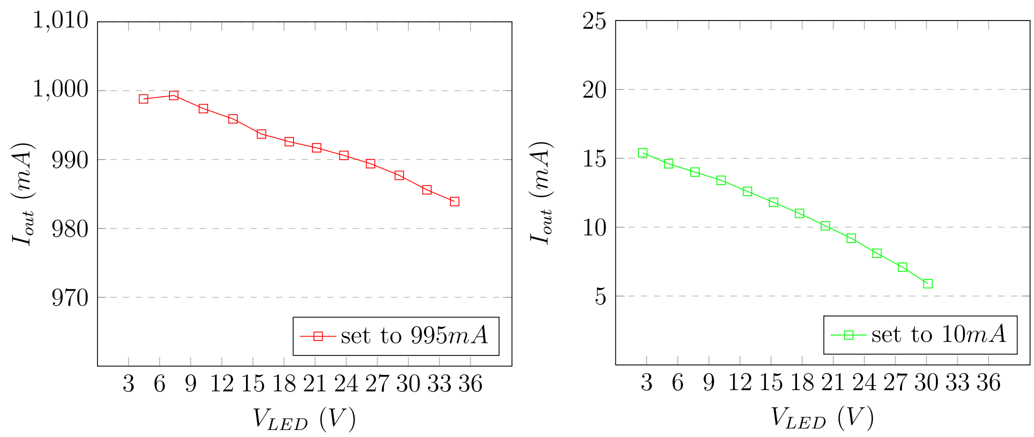


Figure 4.2: Load regulation across the output voltage range for different current settings

4.2 Loss Analysis

A dedicated loss analysis can be helpful to understand where the reference circuit can be optimized in terms of efficiency.

As it can be seen in the schematic, the current before and after the switch-mode converter is constant, neglecting the current ripples. This means, the primary source of losses can be found in voltage drops. Such drops usually occur on diodes and other bipolar semiconductors as well as resistances. The first diodes with voltage drops can be found in the full-bridge-rectifier at the input. While there are already Schottky diodes used, they each drop around $0.5V$ at a current of $1A$. Since always two diodes are conducting, there is a voltage drop of $1V$ present, resulting in a power loss of up to $1W$ at $1A$.

The next losses can appear in the common-mode choke, which has a resistance, as any other coil. From the datasheet of this COILMASTER part, the maximum resistance of 0.4Ω can be taken. The current is again passing twice through the device. The associated power loss is then $P = I^2 * R$ and such $P_{loss,max} = 0.8W$.

On the output side of the switch-mode converter, we have losses in the coil, as well as in the Schottky diode of the buck converter. Calculating these losses is more difficult as the current is actively switched and a constant current can no longer be assumed.

Chapter 5

Hybrid Design

This chapter is going to present the actual design and development process of a hybrid LED driver. First, the parallel approach, that was explained in chapter 3.3.1 is simulated. The results and difficulties are explained afterwards. Second, the serial approach, also explained in chapter 3.3.2, is simulated and the results are explained. Third, additional improvements are made on the linear and the switch-mode stage individually to improve stability, efficiency and other features. Then the dynamic feedback is explained in detail which adds voltage tracking to the device.

Afterwards, the whole driver is characterized and an internal datasheet is generated. This includes all features, as well as important insights about the behaviour during different load conditions.

Last, the PCB layout itself will be analysed and EMI tests, according to the appropriate regulation from chapter 2.4.2, are conducted.

5.1 Parallel Approach

While most of the papers with a parallel hybrid topology are used for voltage control, the properties of a current regulated device is yet to be tested. The first approach is based on the idea of a parallel topology. The basic structure is derived from the proposed driver of Garcia [7].

As proof of concept, a circuit is composed using a standard switching

LED driver and a 3-terminal linear regulator.

5.1.1 Simulation

The schematic in figure 5.1 shows the parallel structure with the switching LED driver LT3763 and an LM317 as linear regulator. In this first simulation the target current was set for both devices individually. As seen in figure 3.5a, the linear regulator is regulating the total current I_{out} although only contributing a small current.

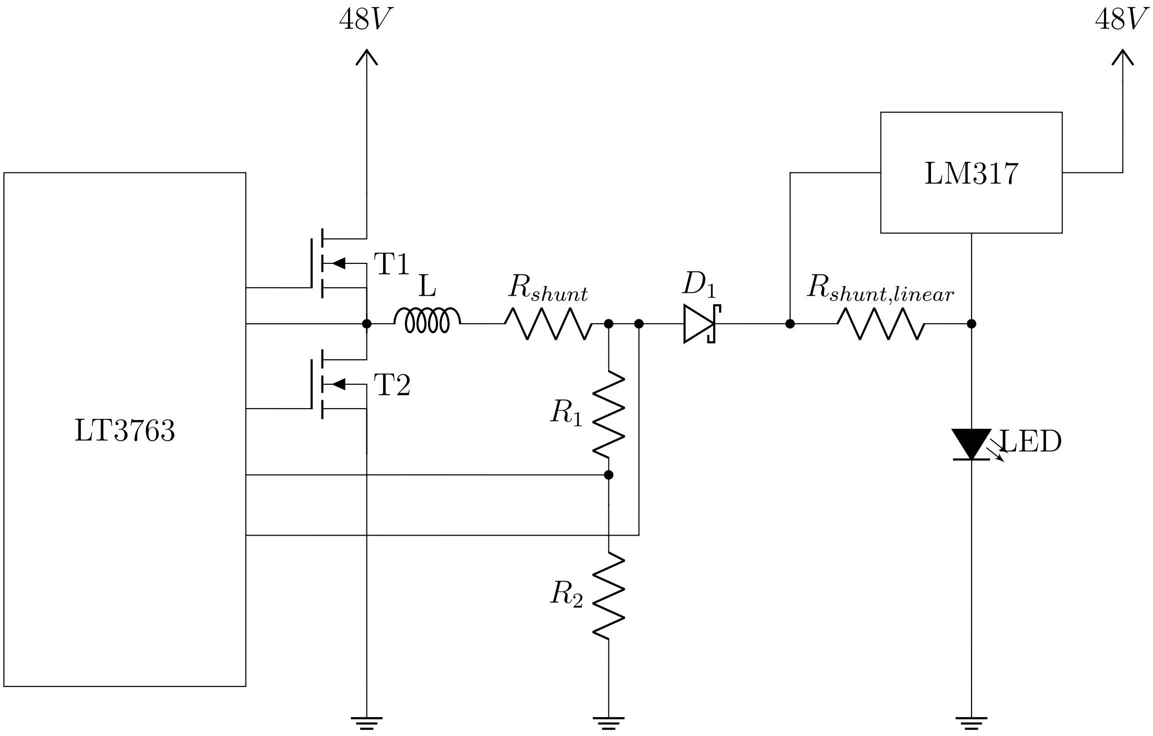


Figure 5.1: Simplified schematic of a parallel topology

Figure 5.2 shows the output current of the linear regulator as $I(R17)$ in red, the output current of the switching converter as $I(D3)$ in pink and the total LED current as $I(D2)$ in green. The total current for the regulation of the linear regulator is set to $0.6A$, while the switch-mode converter is set to $0.45A$. The closeup in figure 5.2b shows the ripple of the switching device with $I_{ripple,switching} < 15mA$ with a frequency of $500MHz$. The linear regulator reacts to the switching ripple to achieve a stable total current I_{LED} with a ripple of $I_{ripple,LED} < 1mA$.

It can be also observed in figure 5.2 that the LED current is relatively stable the whole time and that the linear regulators response is very fast. This is due to the very fast turn-on time of the linear regulator. This behaviour can be advantageous in many applications. These applications can include laser drivers for communication applications or measurement setups, which have to be turned on very fast. The linear regulator will provide the majority of the current until the switch-mode converter is stable.

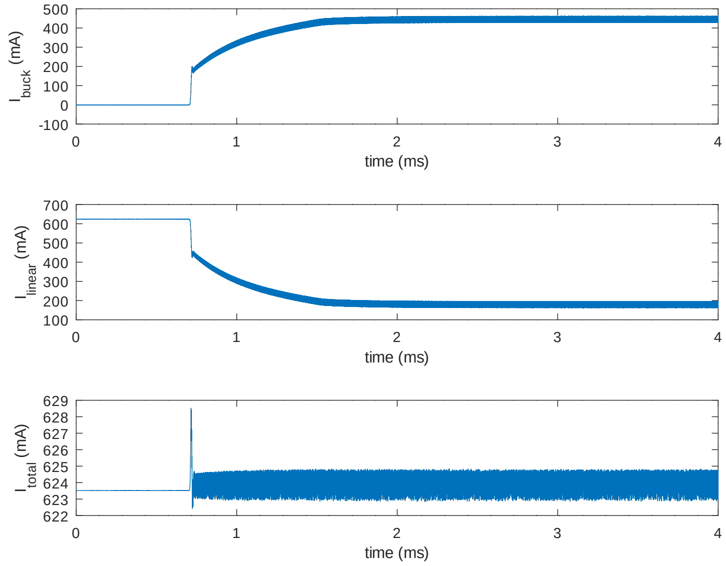
5.1.2 Lessons Learned

The first simulation already showed the difficulties associated with this working principle.

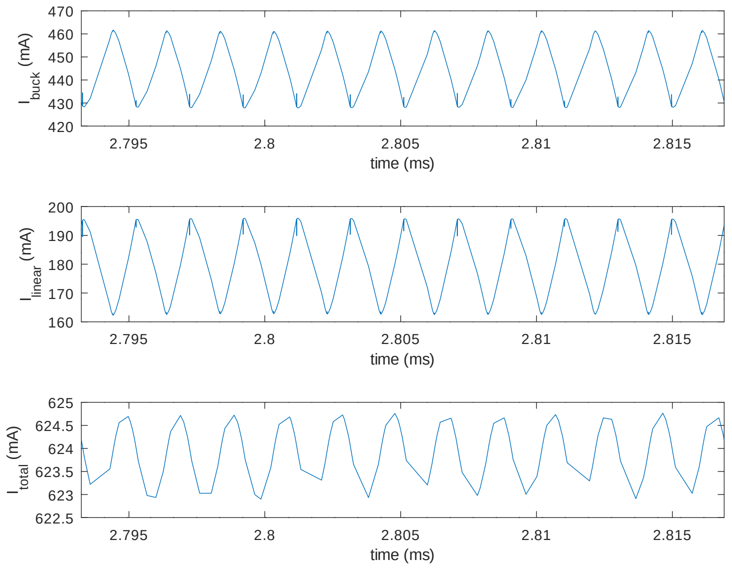
The linear regulator as well as the switch-mode converter have to be connected to the same side of the load. Since most of the commonly available switch-mode converters are high-side devices, the linear regulator has to be on the high-side of the load as well. Many linear regulators have a limited voltage rating of $V_{max} < 30V$. This limits the voltage differential in case of a 3-terminal device or the maximum input voltage in case of a ground connected device.

Another difficulty is the setting of the current. While the circuit works for a set static current, dimming functionality is more difficult. A 3-terminal regulator references its target voltage to its own output voltage whereas the dimming signal, typically generated by a microcontroller, is referenced to ground. Additionally, the output voltage which is the reference, is constantly changing in order to remove the ripple from the output signal.

Another problem will become more obvious in chapter 5.2.2 is the limited bandwidth of linear regulators. In this transient simulation, the bandwidth of the linear regulator is not modelled, leading to a different result in reality. Modern switch-mode converters feature very high frequency switching to reduce the size of the inductor. These switching frequencies are usually



(a) Startup



(b) Closeup of steady-state

Figure 5.2: Transient simulation of a hybrid LED driver in parallel topology

between $200kHz$ and $2MHz$. Unfortunately, linear regulators have a bandwidth that is often below $100kHz$. In chapter 5.2.2, we will see a stable, linear regulator with a bandwidth of $86kHz$, which is still below the switching frequency. This means that the linear regulator is not able to effectively cancel the switching noise.

The very fast turn-on time is advantageous for some application, but not required for common lighting systems. In fact, slower and fading turn-on behavior creates a more natural impression for consumers.

For the above reasons, the parallel approach was not pursuit further in this thesis, since first simulations of the series approach, which can be seen in the following chapter, were showing advantages such as lower complexity and improved robustness. The advantages of parallel solutions can still be considered for voltage output signals.

5.2 Serial Approach

A first Simulation of the series approach was based on the switch-mode voltage converter $LTC3630$ and the common 3-terminal linear regulator $LM317$. The circuit, which can be seen in figure 5.3, first converts the input voltage to an intermediate voltage defined by R_1 and R_2. The $LM317$ is configured as constant current regulator by regulating the voltage across a shunt resistor R_{shunt}. This is a first proof of concept, since the voltage of the switch-mode converter as well as the current regulator are static and can't be adjusted without replacing components.

5.2.1 Simulations

By simulating the circuit in figure 5.3 in LTSpice, it can be seen in figure 5.4 that the intermediate voltage is stable although it shows a high ripple from the switching of the buck converter. The LED current, generated by the $LM317$ is stable.

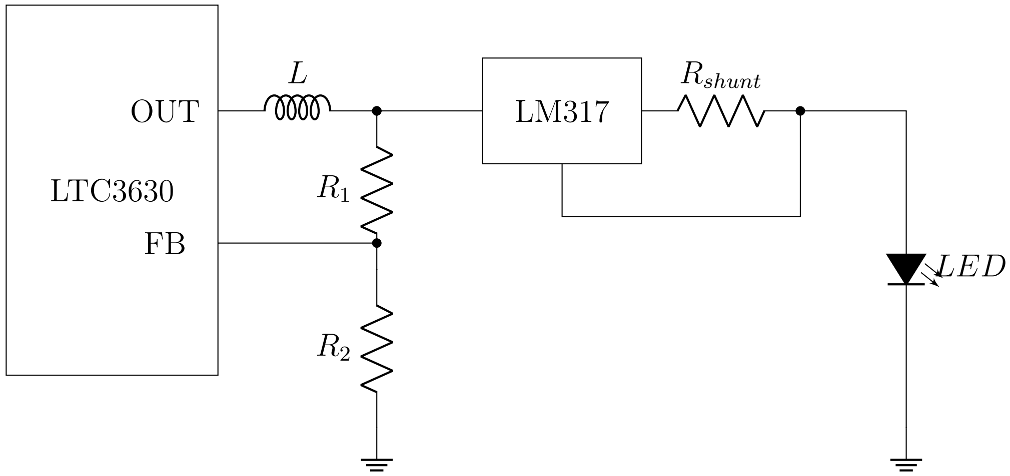


Figure 5.3: Simplified schematic of a serial topology with high-side current regulation

5.2.2 Improving Linear Stages

It was shown in chapter 3.1 how a linear current regulating device can be built. In this chapter, we will focus on the optimization of these circuits.The changes made will be compared in terms of accuracy, dropout voltage, stability, input voltage capabilities, turn-on delays, load regulation, line regulation, dimming behaviour and thermal considerations.

Darlington Stages

The circuit in chapter 3.1 was very basic. One major disadvantage is the inaccurate current through the load. The amplifier senses and stabilizes the emitter current $I_E = I_C + I_B$ while the load only sees the collector current. Bipolar NPN transistor usually have a β of 50 to 250, which results in a load current that is $1/\beta$ lower than desired. This effect can be reduced by the use of a transistor configuration with a much higher β such as a darlington transistor. darlington transistors are using two bipolar transistors to boost their gain to $\beta_{darlington}$, which is calculated in equation 5.1. The circuit can be seen in figure 5.5. Another, error in the accuracy that has to be considered is the offset voltage of the operational amplifier. For devices like

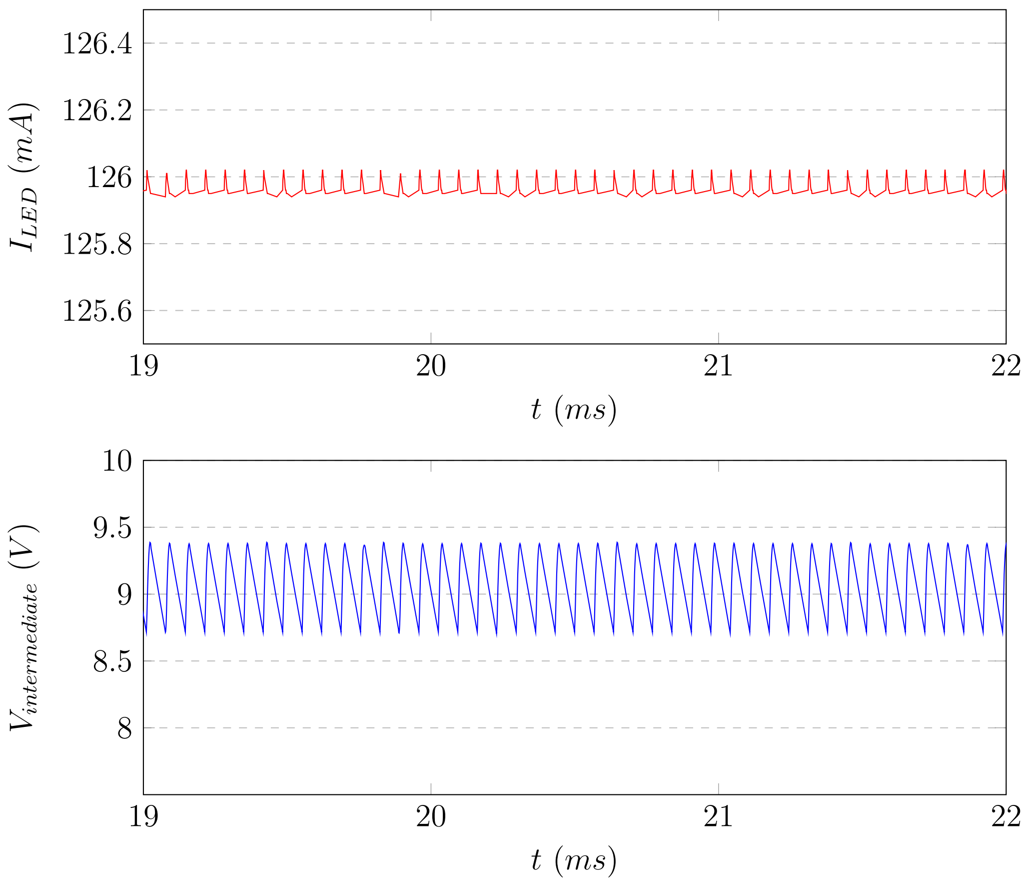


Figure 5.4: Simulated LED current and intermediate voltage of the serial hybrid converter

the $LM358$ it can be as large as $7mV$ and depends on the production process. Especially for small shunt resistors and low currents, the offset voltage can be significant. There are amplifiers on the market which have a factory trimmed offset voltage and even amplifiers with integrated offset correction. A very similar effect on the accuracy has the input bias current. In figure 5.6, the OP284 has a very low offset voltage, but a much higher input bias current than the LM358 in figure 5.7. This bias current causes a voltage drop in the feedback resistor, resulting in an absolute error that is visible at low output currents.

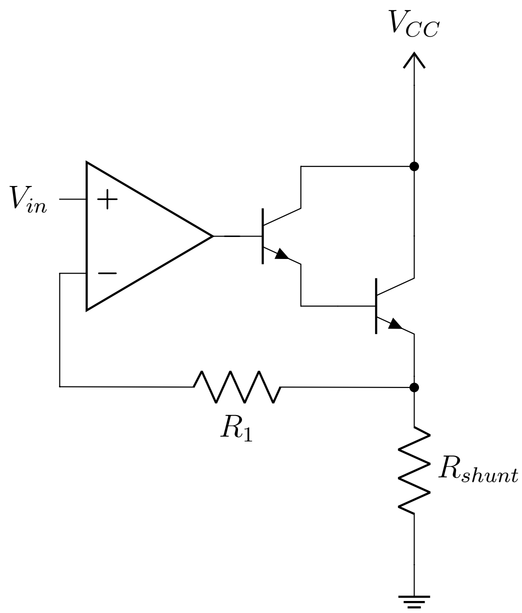


Figure 5.5: Schematic of a linear current regulator with NPN darlington transistor

$$\beta_{darlington} = \beta_1 * \beta_2 + \beta_1 + \beta_2 \tag{5.1}$$

Also the temperature is a factor that has to be considered for accuracy. In figure 5.8 the effects of the temperature on the accuracy of the current can be seen. Two different currents were selected to determine if the error is absolute or is amplified. Since the error is about the same, or gets even bigger with low currents, it is an absolute error.

An essential drawback with darlington transistors is the higher dropout voltage. The required dropout voltage of the linear stage is the saturation voltage of the series pass device added to the voltage drop across the shunt resistor $V_{dropout,min} = V_{CE,sat} + V_{R_{shunt}}$. While a single NPN transistor has a saturation voltage of $V_{CE,sat} = 0.1 - 0.2V$, the darlington configuration needs the first transistor to be in saturation, resulting in the total voltage drop that is higher than a single transistor (see equation 5.2). Since a higher voltage drop means higher losses, this can have a big impact on the maximum current capability. Furthermore, the Voltage $V_{2,CE,active}$ increases with the

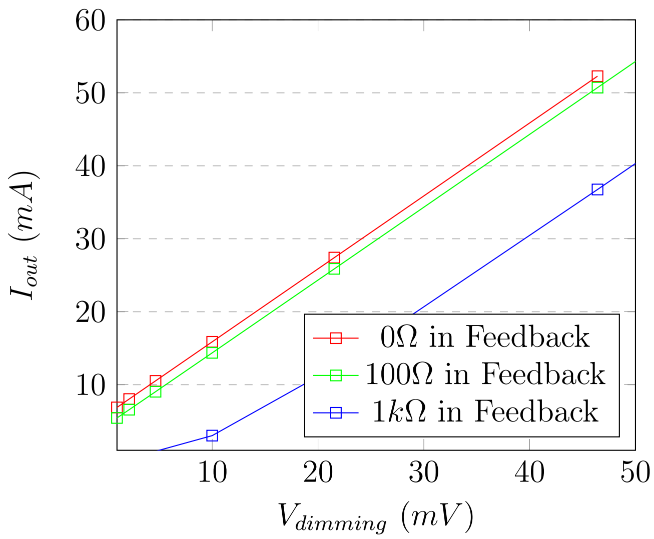


Figure 5.6: Dimming behaviour between 1% and 10% with OP284

current, leading to even higher losses.

$$V_{Dropout} = V_{1,BE,sat} + V_{2,CE,active} \tag{5.2}$$

The minimum dropout voltage can be determined by varying the dropout voltage in a stability analysis. Such an analysis was conducted according to the simplified schematic in figure 5.9. To conduct the stability analysis, a disturbance signal is injected into the feedback loop. This signal is typically a sine wave that is ramping up in frequency, generated by a signal generator. The signal is measured by an oscilloscope and the magnitude and phase of the signal is calculated for each frequency. The phase margin is the remaining phase angle at the frequency where the gain becomes $\beta = 1 = 0dB$. This phase margin is calculated by $180° - \phi$. As the signal is already inverted by the nature of the negative feedback loop of the operational amplifier.

A stability analysis in figure 5.10 shows the relation of dropout voltage and stability. The compensation network was tweaked to provide a high bandwidth. This also means that the gain over the lower frequency range

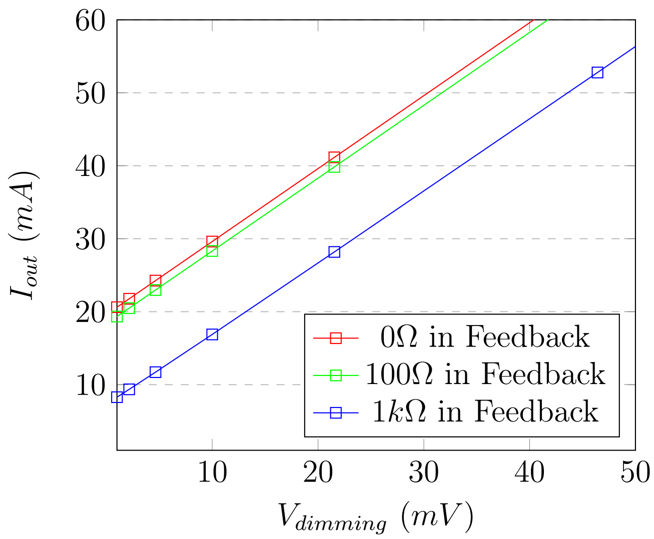


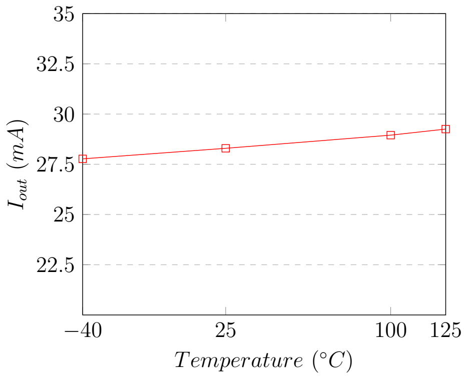
Figure 5.7: Dimming behaviour between 1% and 10% with LM358

is reduced. The lowest curve below $-10dB$ is simulated with a dropout of $0.9V$, the next with $0.95V$ and so on. Above $1.15V$, the transistor is fully in the active regime. With more than $1.15V$, the darlington stage provides a good stability, shown by the large phase reserve together with an acceptable bandwidth. This result also shows, that the bandwidth is too low to actively filter the current spikes that can be introduced by a switch-mode regulator which usually have a switching frequency of $150kHz$ to $2MHz$.

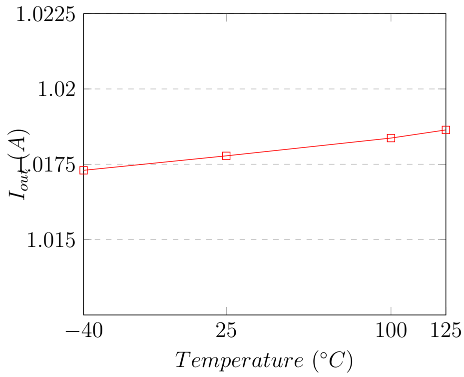
To improve the circuit, the type of amplifier and the type of transistor can be changed. While changing the parameters of the darlington transistor itself only gave minor improvements, changing the amplifier to a different type with higher bandwidth didn't change the bandwidth of the system due to the compensation circuit. In figure 5.10b, the OP284 was used, which has an open loop bandwidth of $4MHz$.

The result was later verified and measured. Figure 5.11 shows the bode plot running on a dropout voltage of $1.3V$ and a current of $0.4A$. The compensation reduces the gain to $< 10dB$ to reach a bandwidth of $39kHz$ with a phase margin of $60°$.

To get an impression of the influence of the dropout voltage on the ex-



(a) current set to $10mA$



(b) Current set to $1A$

Figure 5.8: Current vs. Temperature for different current settings

pected current ripple, the PSRR (Power Supply Rejection Ratio) was simulated for different dropout voltages beginning with $0.9V$. It has to be noted that the PSRR is usually defined as $20log\frac{\Delta V_{in}}{\Delta V_{out}}$. But here, we refer to the current ripple $20log\frac{\Delta I_{out}}{\Delta V_{in}}$ where the Rejection ratio is a transconductive function.

The line-regulation of the current sink is very good when the dropout voltage is high enough. The load-regulation was already included in the previous measurements, since for a current sink, it's the variation of the load voltage. This was simulated by varying the dropout voltage as seen in the previous measurements. For a dropout voltage above $1.15V$, the load-regulation is minor.

In figure 5.14, the line regulation was measured on a linear regulator based upon a TP1542 operational amplifier by 3peak and a MMBT5551 transistor. The minimum dropout voltage is clearly visible and rises with increasing load current because of the increasing saturation voltage. Once the $V_{dropout,min}$ is reached, the line regulation is below 0.5%. In theory, the line regulation depends on the early voltage, which is typically high for bipolar transistors.

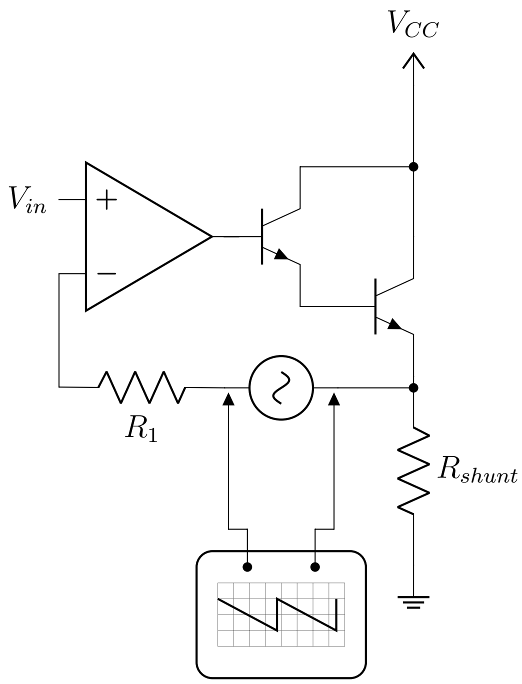
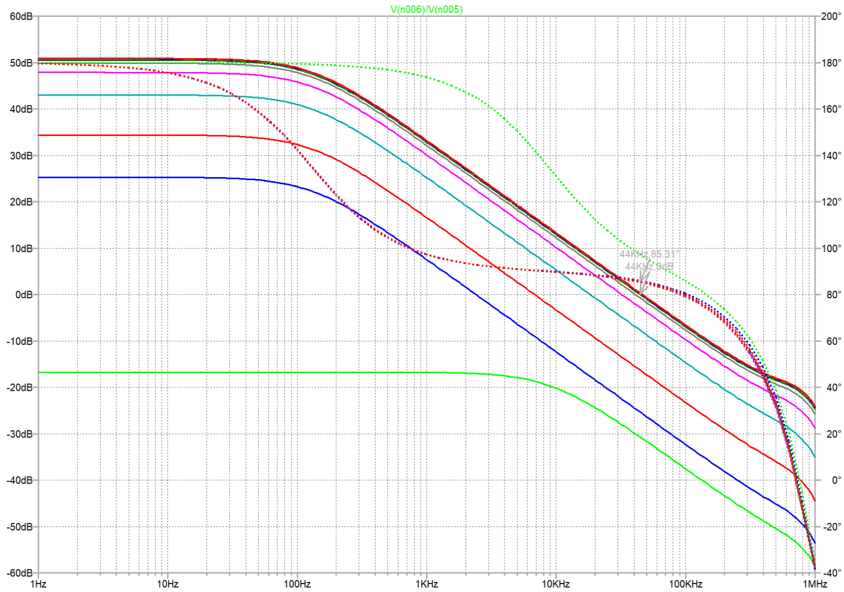
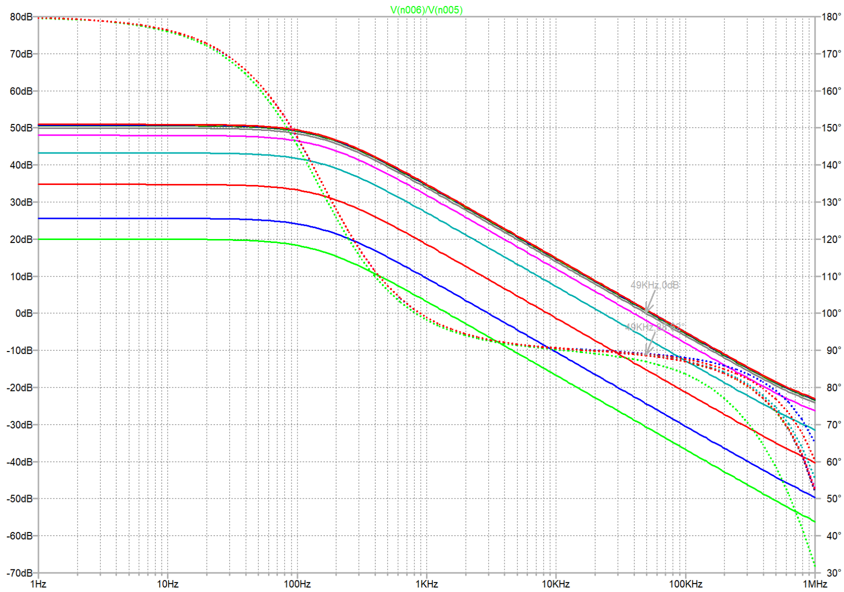


Figure 5.9: Measurement setup for the stability measurement of the linear regulator with darlington stage



(a) LM358 as operational amplifier



(b) OP284 as operational amplifier

Figure 5.10: Simulated stability analysis of a current sink using a darlington configuration

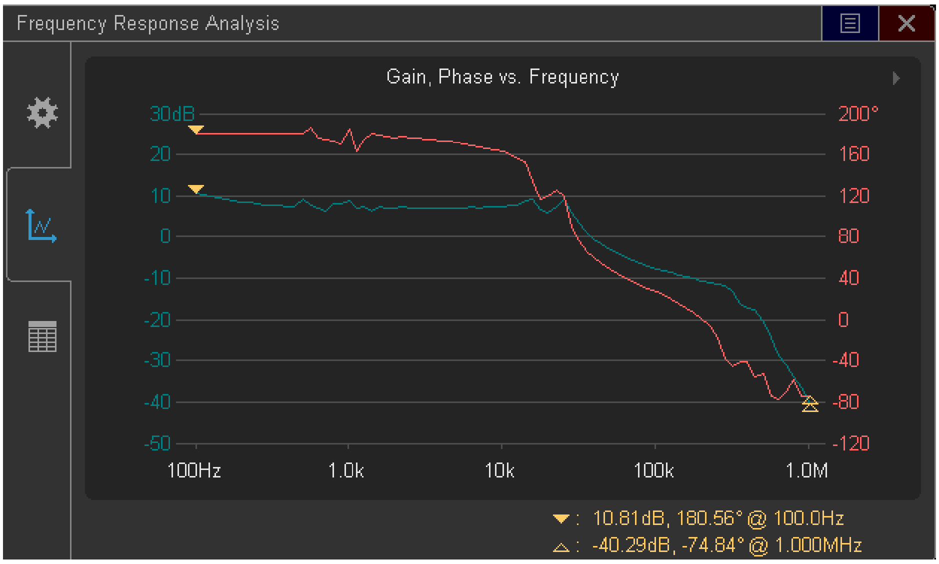
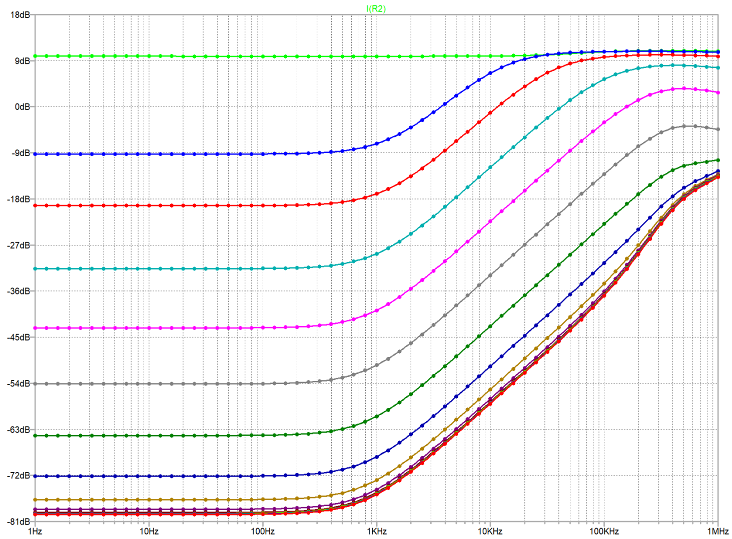
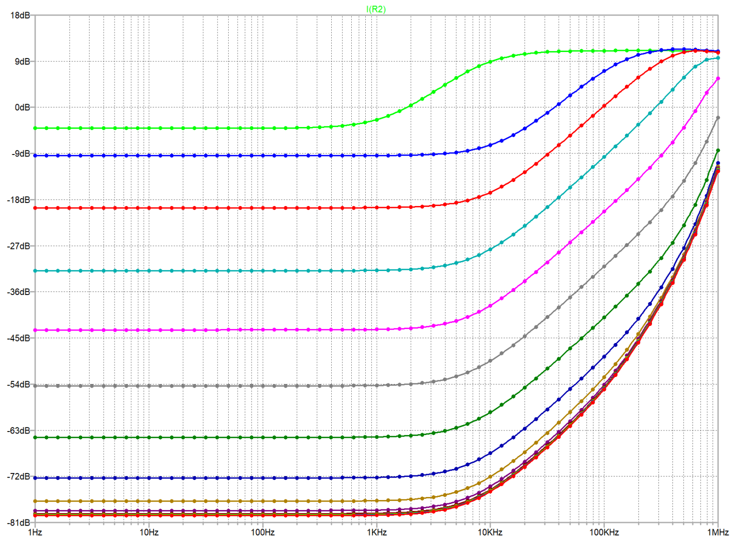


Figure 5.11: Measured stability analysis of a current sink using a darlington configuration

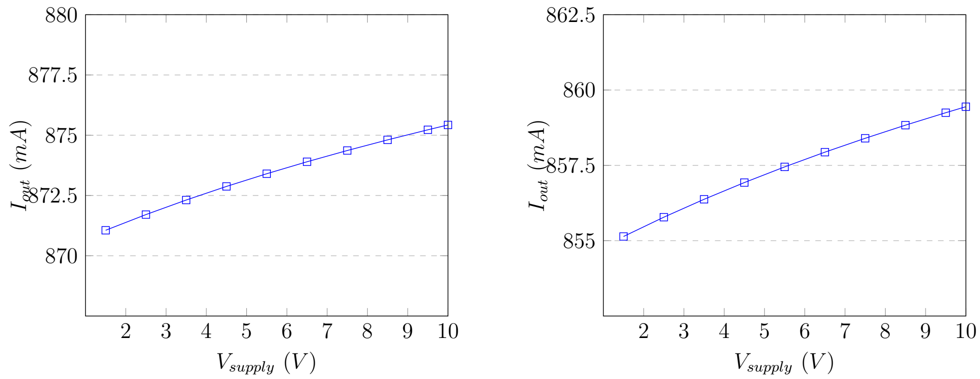


(a) LM358 as operational amplifier



(b) OP284 as operational amplifier

Figure 5.12: Rejection of a power supply ripple of a current sink using a darlington configuration



(a) LM358 as operational amplifier (b) OP284 as operational amplifier

Figure 5.13: Output current of a current sink using a Darlington for different supply voltages

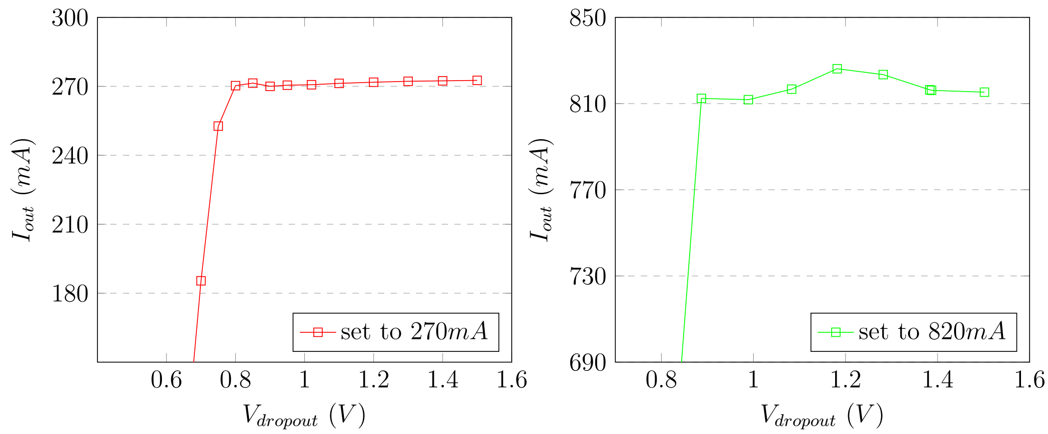


Figure 5.14: Line regulation across the output voltage range for different current settings

MOSFET Stages

It was already shown, that a low saturation voltage $V_{CE,sat}$ together with high gain or no current into the base is desired. This makes a MOSFET device almost perfect as an output stage. First, MOSFETs have no galvanic connection between Gate and Source, so that there won't be any DC current flowing into the Gate, besides leaking currents. Second, the saturation voltage of MOSFETs is much lower and can be approximated with equation.5.3, where the overdrive voltage $V_{GS} - V_{th}$ needs to be high enough that the desired current can flow. Since the voltage drop will cause losses, V_{DS} will be kept as low as possible.

$$V_{DS,sat} >= V_{GS} - V_{th} \tag{5.3}$$

The circuit in figure 5.15 was simulated. Since there is no significant DC Gate current and the gain is extremely high, a single transistor is capable of regulating accurate enough.

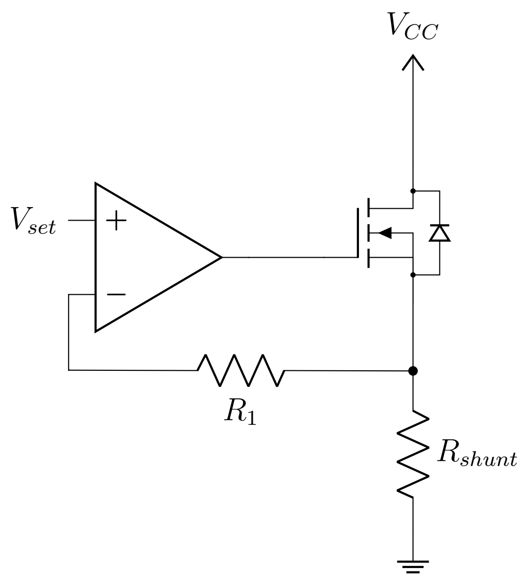


Figure 5.15: Schematic of a linear current regulator with N-channel MOSFET

The switch-off behavior is an important factor when it comes to LED lighting. LEDs can glow with currents below $1\mu A$, which is very undesirable because the light is visible especially during night. Figure 5.16 shows the dimming behavior without and with switch-off design. The red line in figure 5.16 shows still a current flowing because of offset voltage of the LM358 OPAMP and parasitic currents. The green and blue line in figure **??** shows the behaviour, when a resistor R_1 is introduced in the feedback path. The bias current into the OPAMP will lead to a voltage drop over the resistor, lowering the feedback voltage and introducing an offset in the linear dimming curve.

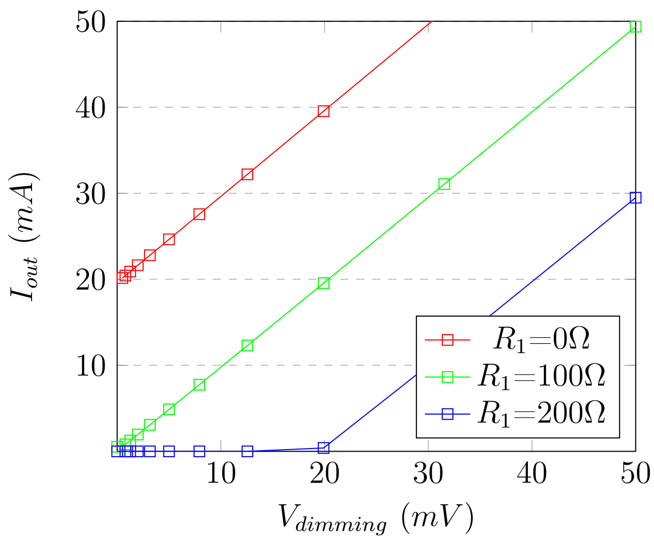
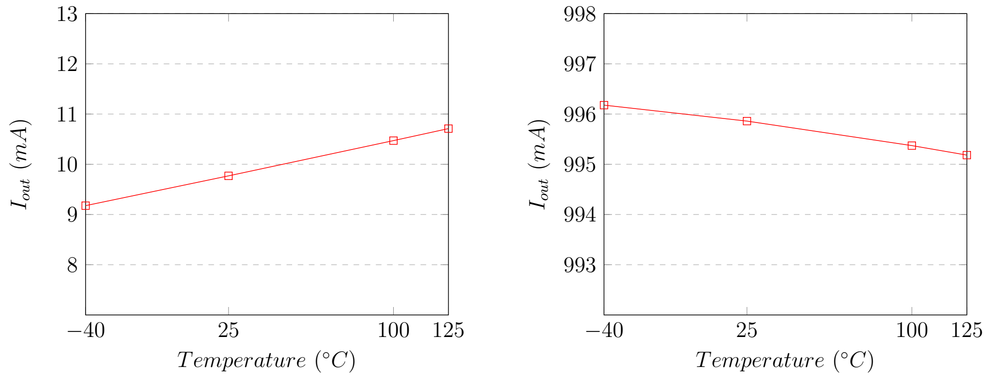


Figure 5.16: Dimming behaviour between 0% and 5% of a $1A$ current sink

The temperature stability of the MOSFET regulator shows a similar response as the darlington stage for $10mA$. But as it can be seen in figure 5.17b, it changes to a negative temperature coefficient.

The dropout for the previous simulations was set to $0.5V$, which is very low compared to the $1.3V$ for the darlington configuration and decreases losses by more than 60%. But MOSFET devices have a significant drawback in the sense of stability. The Gate capacitance, along with parasitic capacitances from Drain to Gate and Gate to Source make a regulation difficult.



(a) current set to $10mA$ (b) Current set to $1A$

Figure 5.17: Current vs. Temperature for different current settings

This means, the compensation circuit has to be much more sophisticated.

The stability analysis in figure 5.18 was done for different dropout voltages from $0.2V$ in green, up to $0.8V$. Here, it can also be seen that the MOSFET needs at least $0.4V$ (red curve) dropout. The simulation shows an unstable case with the OP284 OPAMP and a stable behavior with the LM358 which has a lower bandwidth.

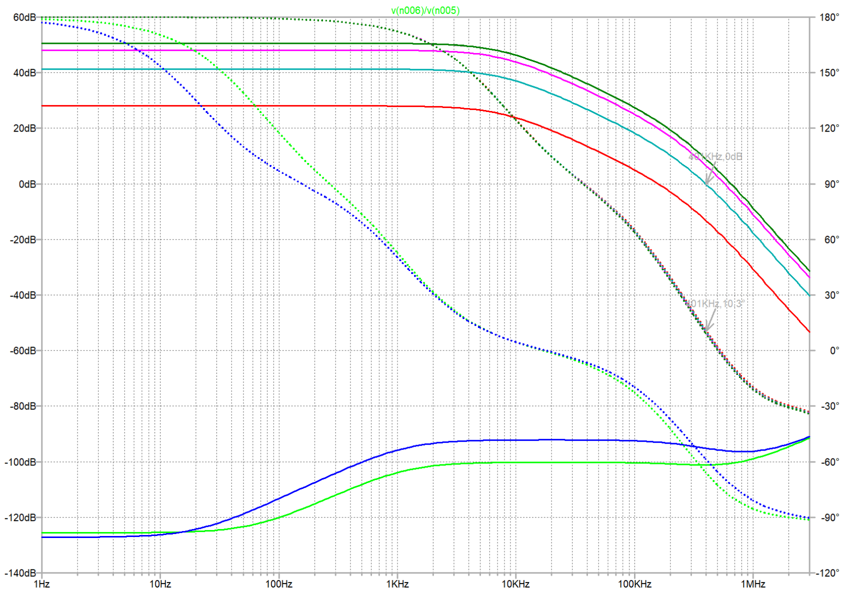
In figure 5.19, the frequency analysis was performed with an TP1542A and an ZXMN6A07ZTA N-channel MOSFET from Diodes. The dropout was set to $0.6V$ with a current of $0.2A$. The results show again, like with the darlington stage, that the gain was reduced to $< 10dB$ to reach the highest possible bandwidth. The bandwidth in this case reached $45kHz$ with a phase reserve of 96°.

The higher bandwidth of the OP284 gives a far better PSRR, as it can be seen in figure 5.20a and 5.20b.

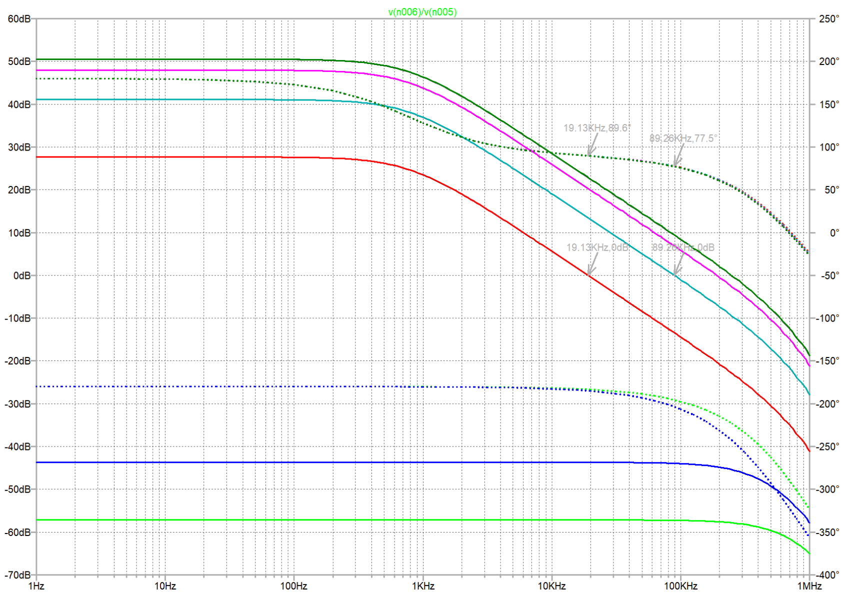
The line regulation of the MOSFET stage was simulated in figure 5.21 with a dropout of $0.5V$. Compared to the darlington stage, it is significantly worse, which means that the change of the output I_{out} has a stronger relation to the input voltage V_{in}. The reason is the V-I curve in the saturation region, which is described by the Early-Effect for the bipolar transistor and the Channel-Length-Modulation for the MOSFET.

Furthermore, the current spikes at the transition are as well much larger than with the darlington configuration. This is likely due to the worse PSRR of the MOSFET stage at higher frequencies.

In figure 5.22, the line regulation was measured for a linear regulator based on the TP1542A by 3peak and the ZXMN6A07ZTA n-channel MOSFET. The minimum required dropout voltage can be easily seen and is far lower than the dropout voltage of the darlington device. The line regulation of a MOSFET depends on the channel length modulation that leads to a result similar to the early effect. Usually the channel curve of a MOSFET in saturation is not as flat as a bipolar device, which is the reason, we see a slightly worse line regulation here.



(a) Unstable OP284, different compensation is needed



(b) Stable LM358

Figure 5.18: Stability analysis for different dropout voltages

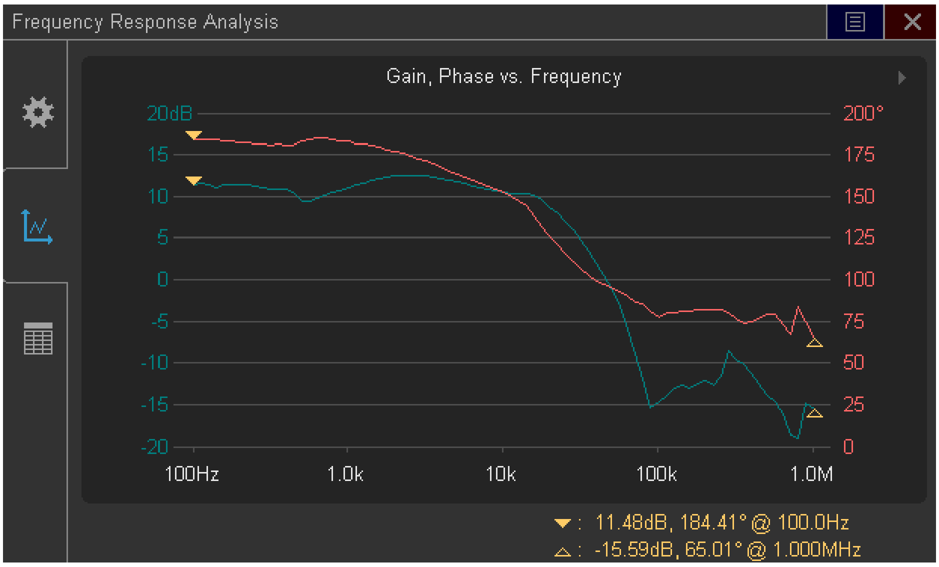
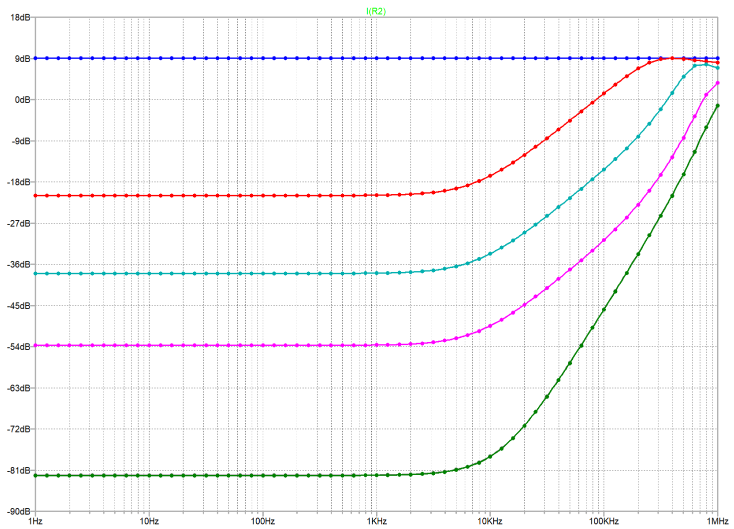
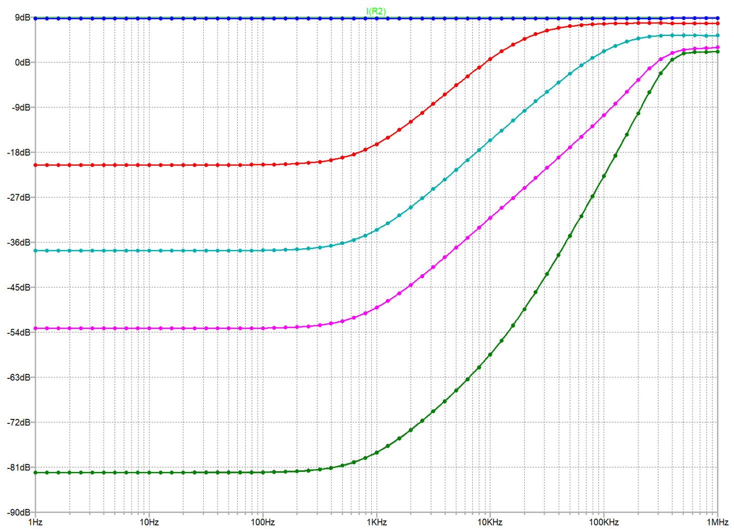


Figure 5.19: Measured stability analysis of a current sink using a MOSFET configuration

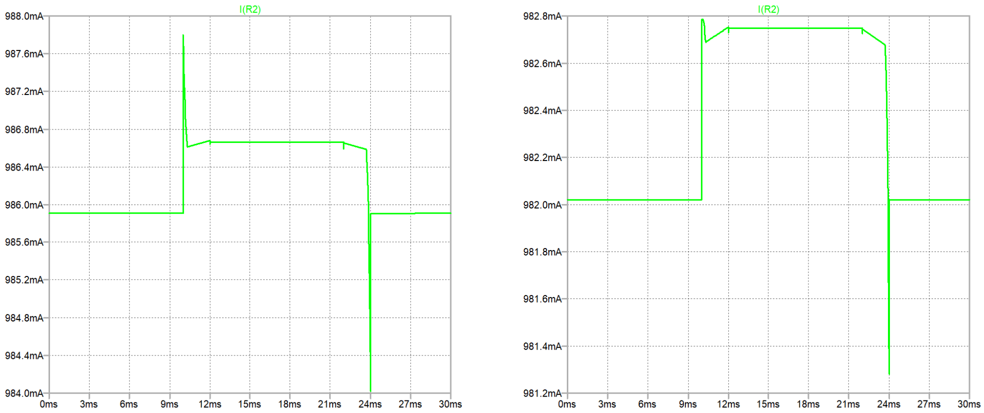


(a) OP284 as OPAMP



(b) LM358 as OPAMP

Figure 5.20: PSRR for different dropout voltages



(a) LM358 as OPAMP (b) OP284 as OPAMP

Figure 5.21: Line regulation for a $1V$ voltage step on the supply

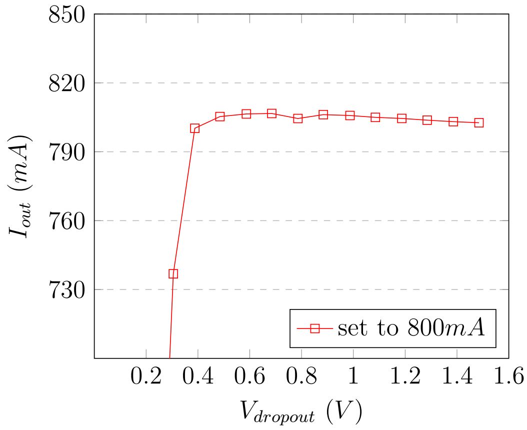


Figure 5.22: Measured output current across the output voltage range for 800mA

5.2.3 Improving Switch-Mode Stage

A hybrid system that consist of linear and switch-mode parts must be optimized individually first. This gives a good starting point because the individual parts are already stable and working. As mentioned in chapter **??**, the system will be a series topology featuring a switch-mode voltage regulator as a pre-regulator for the linear current regulator. The linear stage was already optimized in the previous chapter.

In this chapter, the focus lies on the switch-mode regulator. When choosing a switch-mode regulator, a number of parameters need to be evaluated such as switching frequency, voltage ranges, current capabilities as well as stability. Since the application is space and cost sensitive and only needs to lower the voltage from the $48V$ input, we will focus only on the single-inductor buck regulator.

There is a huge variety of switch-mode voltage regulator ICs to choose from. An important difference is the control principle. There are two popular control principles, namely voltage control mode (VCM) and current control mode (CCM). In VCM, a the output voltage is compared to a generated voltage ramp, controlling the switching transistor, while the CCM generates the ramp from the rising current in the inductor. In a technical note, Maniktala compares both modes with their control theory and deducts their individual strengths and weaknesses. While VCM suffers not only from a bad line regulation due to an independently generated voltage ramp, which can be accommodated for, it also has two poles in the transfer function $G(s)$ which requires a more sophisticated compensation network with two zeros that cancel out the poles of the transfer function. Since CCM only inherits one pole, it is much simpler to compensate which makes this control mode much more stable. Maniktala mentions subharmonic instability as a drawback of CCM in terms of stability, which usually occurs at half of the switching frequency $f_{switching}$ and is an irreversible state where the duty cycle of the switching alternates between two values (usually very low and very high) [27].

The specific application as a hybrid converter asks for a varying load current and thus, a good load regulation. Since the current varies so much (down to 1% of the maximum current), it is very challenging to ensure stability over this wide range. In the conventional design approach, all parts are specified around a sweet spot of the output conditions. Especially the inductor is very critical in this aspect as it must meet the demands of the maximum current, which usually give a too low inductance for an efficient operation at minimum current.

As part of a dynamic system, the output voltage well change as well from $3V$ to $45V$, depending on the number of connected LEDs. This means that a high efficiency must be granted, especially int the case of high voltage ratios. Stability has to be ensured even for changing target voltages. Furthermore, audible noise and EMI must be kept to a minimum.

These parameters, especially the stability issues, make CCM the choice to go for.

The MP4558 switching regulator was chosen for future optimization because it is based on CCM and programmable switching frequency besides having a high voltage ($55V$) input capability. The high switching frequency of up to $2MHz$ is necessary to keep the inductor small, especially for larger load currents. The light load capability is very important as well, since dimming down to 1% reduces the current considerably. This light load capability is achieved by lowering the switching frequency, so that the minimum ON-time can still me met. This is usually done by pulse skipping, which is a control strategy that does not switch on every cycle of the internally generated clock signal. Drawback of this behaviour is the lower frequency of the ripple current which can lead to audible noise in ceramic capacitors.

5.2.4 Adding Dynamic Feedback

Connecting both stages is simple for a static solution, where the load voltage and thus the forward voltage V_f and number of in series connected LED's does not change. However, since the forward voltage does change due to process variation, temperature and different types of LED's, it is necessary to introduce a dynamic feedback that regulates the voltage potential after the LED string close to the minimum voltage drop of the linear stage.

A first approach is connecting the node above the linear regulator directly to the feedback pin of the switch-mode regulator. Although this is a viable option, it has disadvantages since it will always regulate towards the same potential. The reference voltage for the MP4558, which is used in the chapter 5.2.3, is $0.8V$. This is too low for a bipolar or darlington regulation and too high for a MOSFET regulator.

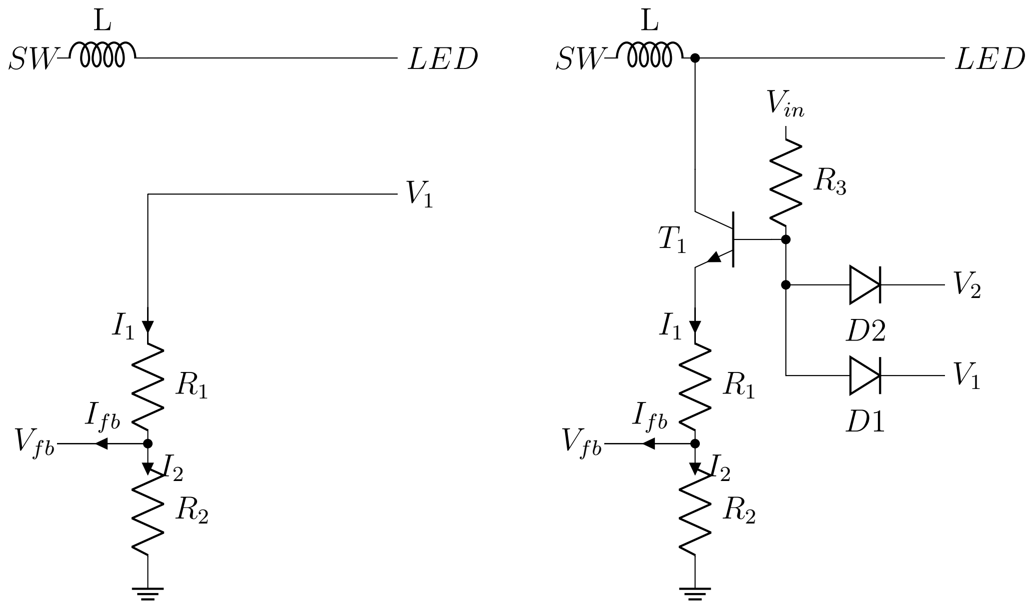
In order to control set the desired voltage, a voltage divider as well as an emitter follower as buffer stage is introduced to reduce the bias current through the LED's that is not regulated by the linear stage. The current through the voltage divider is set by the feedback voltage $V_{feedback} = 0.8V$ and R2. This current is set to be $0.8mA$. The voltage across the linear regulator can then be calculated with equation 5.4. For $R1 = 1k$ and $R2 = 10k$, we receive $1.48V$ which is enough drop for a darlington stage.

The equation 5.4 also shows that the lowest dropout voltage is the fixed feedback voltage $V_{feedback}$ of the switch-mode converter. Considering the MP4558, which was used in chapter 5.2.3 that has a reference voltage of $0.8V$, a linear MOSFET stage can not be optimized for efficiency.

$$V_{Drop,OP} = V_{feedback} + \frac{V_{feedback}}{R2} * R1 + V_{be} \tag{5.4}$$

The introduction of the diodes D1 and D2 in combination with the resistor to the supply voltage in figure 5.23 has the advantage of lowering the minimum dropout voltage by the forward voltage of the diodes as well as introducing a voltage selection in favor of the highest LED string voltage of a two channel system. In a two channel application where two separate LED strings are supplied, the LED string with the highest forward voltage is regulated, ensuring enough dropout voltage for the other channel. Because the channel with lower forward voltage is not regulated for optimal efficiency, its efficiency decreases. The current flowing through the diode into the linear regulator is $\frac{V_{supply} - V_{f,diode} - V_{dropout}}{R_{or}}$ which is designed to be $< 10\mu A$.

Unfortunately, the diode introduces an additional negative temperature coefficient. The temperature coefficient of the diode adds to the temperature



(a) Simple feedback path (b) Advanced feedback path

Figure 5.23: Schematic of circuit for a constant voltage over linear stage

coefficient of the bipolar transistor. Because the total coefficient is still negative, the feedback voltage will drop with increasing temperature, resulting in an increase of the voltage drop across the linear stage. Because of the negative temperature coefficient of the transistor inside the linear regulator, the efficiency drifts even further from the theoretical optimum.

Common diodes, such as the 1N4148 has a temperature coefficient of $-2mV/K$ while $\theta_{V_{be}}$ of a BC846 as typical NPN transistor also is around $-2mV/K$. Only considering these elements, and considering the same temperature on all parts, $\theta_{V_{drop}}$ will be $0mV/K$ according to equation 5.5 but the threshold voltage and thus the saturation voltage of the linear regulator depends on the device. For the N-Channel MOSFET, it drops with $\theta_{V_{th}} = -3mV/K$, which causes an efficiency loss. The imbalance becomes worse, when the diode is replaced by two diodes to reach efficient dropout voltages for MOSFET devices [28] [29].

By introducing an element with a negative temperature coefficient in R_1, or a positive temperature coefficient in R_2, the efficiency loss due to temperature drift can be reduced.

$$\theta_{V_{drop}} = -\theta_{V_{fb}} = -\theta_{V_{f,D1}} + \theta_{V_{be}} + \theta_{V_{R1}} - \theta_{V_{R2}} \tag{5.5}$$

$$\theta_{V_{drop}} = \theta_{V_{th}} \tag{5.6}$$

5.2.5 Characterization

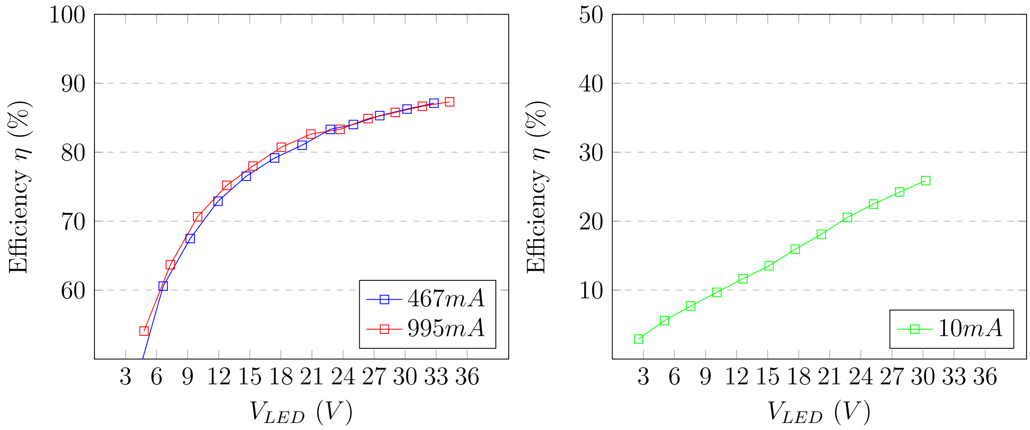


Figure 5.24: Efficiency across the output voltage range for different currents

In figure 5.25, the current of the driver was measured across the output voltage range. The current was set to $960mA$ and to $9.6mA$, which is 1% of the maximum current. It can be seen, the the current stays very stable. Especially for the low current, no major current difference can be seen. The load regulation is $< 0.1mA$ which is less than 1%.

5.2.6 Fault Analysis

Besides the theoretical design of a device, an appropriate analysis and risk management is essential to bring a product to market. It is necessary to figure out what can go wrong and how the device is affected.

The first fault condition is the open failure of an LED. This is the most common failure of an LED. Derived from the schematic in figure 6.5, the potential above the linear stage becomes ground. Therefore, the voltage at the feedback input of the switch-mode stage will also drop, resulting in a

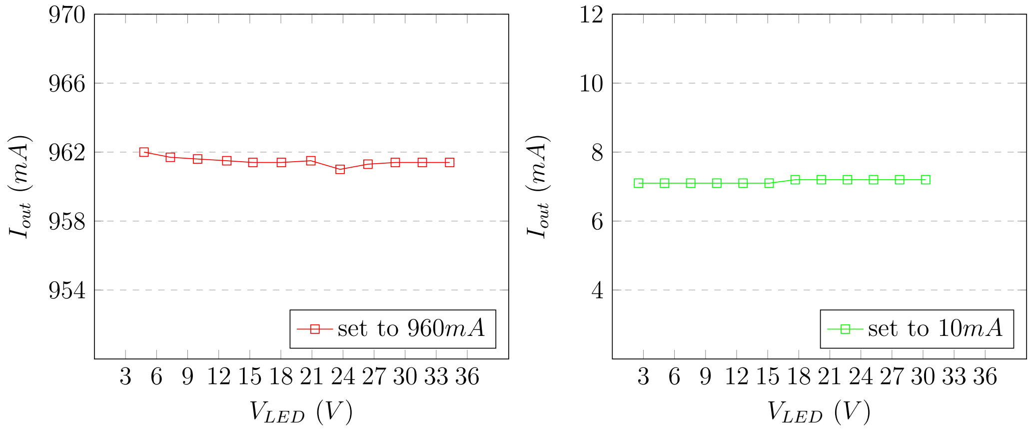


Figure 5.25: Load regulation across the output voltage range for different current settings

rising output voltage. The output voltage rises until it reaches its maximum. Since no current can flow through the LED anymore

Chapter 6

Conclusion

Throughout this thesis, two different hybrid topologies have been considered. In general, a parallel and a serial topology of switching converter and linear regulator are possible. The parallel topology has advantages in a very fast response and startup time. But the simulation already showed many obstacles like high voltage stress for all elements and difficult referencing of a dimming signal. Since the application of lighting can't the series approach was pursued.

The series approach, comprising a switch-mode voltage converter and discrete linear current regulator was showing less required complexity and reduce voltage stress due to the use of a low side linear stage. In order to optimize the converter, the discrete linear current regulator was analyzed and optimized in detail and losses have been reduced considerably. After optimizing the stages separately, a circuit has been established that reduces losses of the complete circuit.

As expected, the hybrid solution is able to maintain some important advantages of both technologies. Disadvantages are obviously the higher complexity compared to a non-hybrid solution.

6.1 Discovered Parameters

As expected, measurements of the final LED driver are showing a slightly decreased efficiency compared to the reference solution using a dedicated switch-mode LED driver. A comparison of the efficiency across the range of load voltages can be seen in figure 6.1.

While being vastly more efficient than a linear regulator, the load regulation as a week point of the switch-mode converter was improved as shown in figure 6.2. Especially in the low current operation, such a variation of the output current must be prevented as they can easily be observed with by the human eye due to its logarithmic sensitivity.

The line regulation is barely detectable, which is a typical feature of switch-mode converters.

A short circuit protection is inherently integrated in the design as the switch-mode stage reduces the output voltage to maintain a low voltage over the linear stage in order to minimize losses.

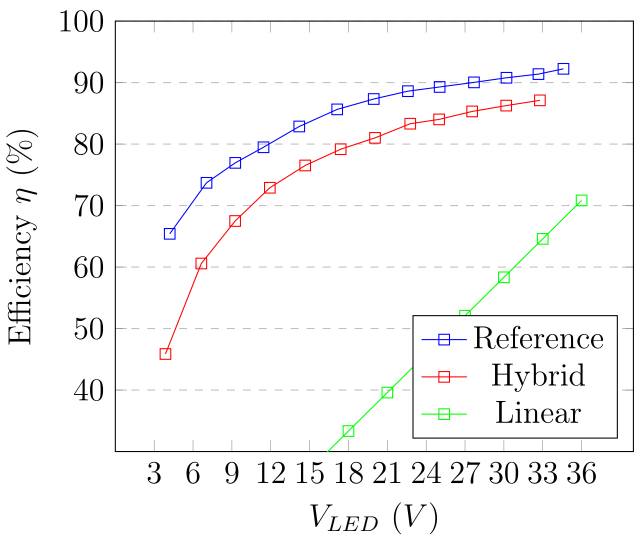


Figure 6.1: Efficiency comparison (measured) for $467mA$ LED current

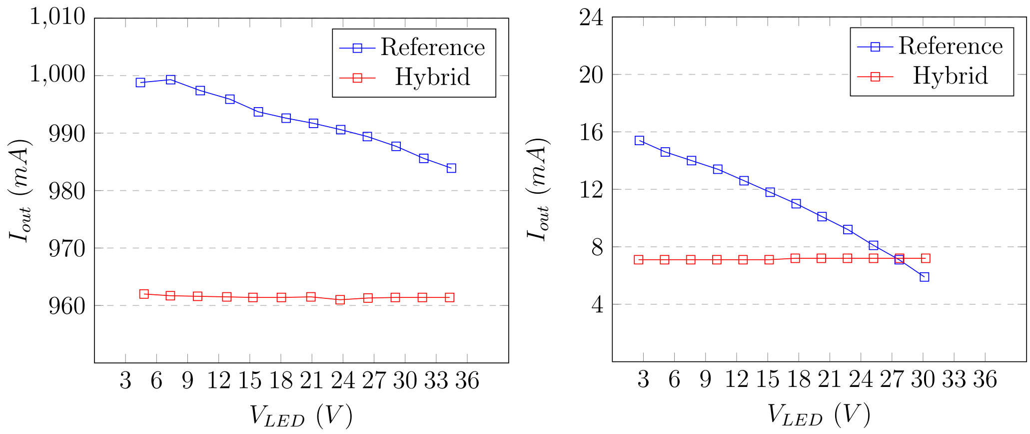


Figure 6.2: Measured load regulation across the output voltage range for different current settings

6.2 Future of DC LED Drivers

The development of LED drivers, especially DC to DC LED drivers, is targeting higher power levels as the losses decrease. The efficiency of switch-mode LED drivers is being pushed by the recent developments in GaN technology. Means to improve the dimming behaviour of switch-mode LED drivers are focusing on hybrid dimming, which combines analog current regulation up to a certain load level and using PWM dimming to achieve acceptable linearity below this current level.

List of Figures

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