Constant current paralleling controller for mid-power LEDs

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Abstract

A simple, cost-effective current sharing controller integrated in a LED module is presented. By this technology it is possible to drive high luminous flux, SELV-compliant LED arrays, using a large number of mid-power LEDs arranged in a series-parallel configuration. Still a conventional constant current power supply can be used.

This novel current sharing concept uses one dissipative, cost-effective, balancing current sink per string, implemented by using a single operational amplifier, and one global averaging circuit. The document describes the circuit and corresponding design equations. Operation is demonstrated by measurements.

1. Introduction

The nominal output power of LED floodlights is rising with recent development. For optimal performance, LED strings connected in parallel should be driven with matched current [3]. Currently paralleling is avoided by series-wiring all LEDs [4]. But that approach has safety limits: SELV limits the total DC voltage to 120 V [5]. To increase safety and lower touch protection requirements, some LED lamp manufacturers apply the lower SELV limit of 60 V. Assuming 3 V forward voltage per LED, the maximum number of standard series LEDs is hence limited to 20. Another industry trend in LED flood lighting is to use mid-power LEDs [6], instead of high power LEDs, which tend to glare and are less cost-effective.

As the luminance per area of mid-power LEDs is lower, they promise to meet anti-glare requirements without using lenses. Omitting the lenses reduces cost, increases efficacy and avoids yet another failure mechanism. To achieve high luminous flux with hundreds of mid-power LEDs, current single channel LED drivers require high output voltages and are therefore not SELV compliant.

In this paper a simple, safe SELV-compliant and cost-effective solution for paralleling LEDs is proposed, which allows to use a standard, single channel, constant-current LED driver to parallel connect LED strings.

2. State of the art

The state-of-the-art approach is to separate the LED driver, into input and output stage, shown in Fig. 1. The input stage converts mains AC to a fixed DC output. The output stage is responsible for providing the appropriate currents to each LED string.

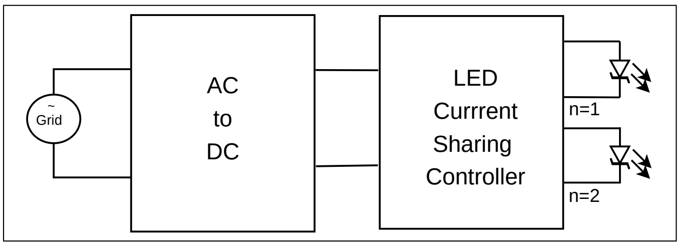

Fig. 1: Typical generic LED driver topology for parallel driving LEDs.

2.1. Resistors

The resistor method paralleling approach in Fig. 2 is widely used, e.g. in LED stripes. This paralleling approach can compensate the LED string currents, however LED string currents can only be compensated, but not matched. To achieve a decent and reliable equalization high resistances must be used. However using a high resistance value introduces losses, lowering the modules efficacy.

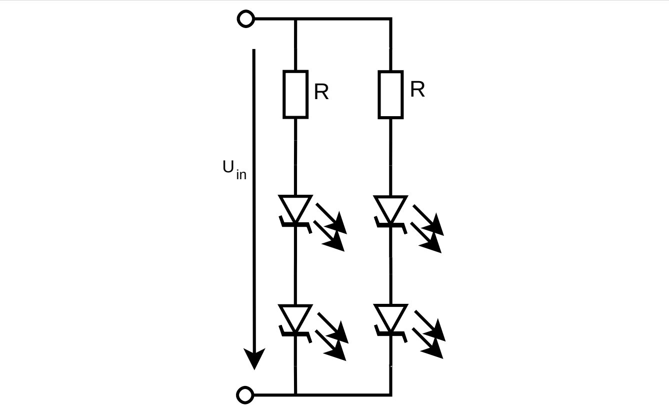

Fig. 2: Paralleling LEDs can be achieved in the simplest manner by using resistors. While being simple and straight forward, that solution has high electric losses.

2.2. Buck Converter

Another option for paralleling is by using a regulated buck converter for each string. In case one or several LED fail to short circuit, the buck converter can compensate for the lower string voltage. Using this approach the LEDs do not need to be matched by forward voltage – thus the driver is more versatile. This flexibility comes at increased cost – especially when tens or hundreds of parallel channels are required.

To replace a 2.8 A high power LED by 120 mA mid-power LED, 24 parallel strings are required. That's why the state-of-the-art constant current approach is highly costly and impracticable for mid-power LEDs. Additionally this approach is only possible if a higher entity, like a microcontroller, defines the desired forward current.

3. Self averaging topology

3.1. Overview

A new averaging regulation topology targeted for low cost is proposed in Fig. 3. The schematic is shown for two LED strings, but may be extended to $n \in \mathbb{N}$. The paralleling controller aims to distribute a constant current source equally to the LED strings. Therefore it measures all string currents, averages them, and provides that value as a reference to the dissipative constant-current sink of each string.

If an LED fails to short circuit, the string current does not change, as the additional voltage is dissipated in the constant current sink. The average current will also not change.

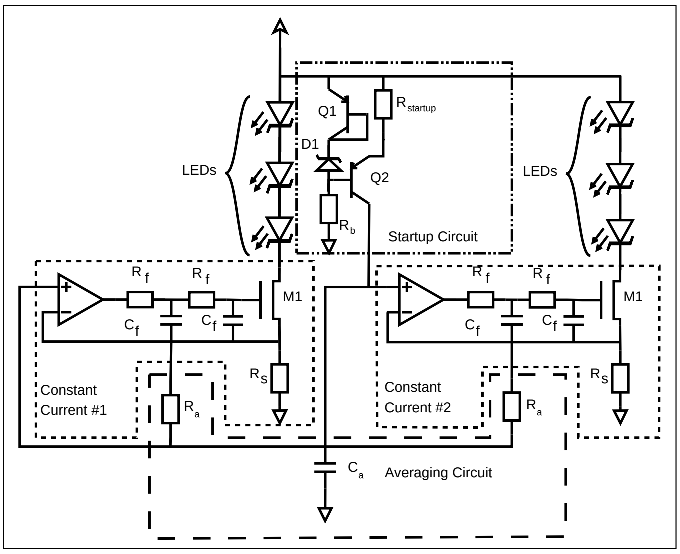

Fig. 3: Proposed novel current matching topology, based on a dissipative current limiter with self-determining set current. For startup an additional startup circuit is required.

This simple paralleling approach is easy to use and does not require a predefined reference value by the circuit or from a microcontroller. Hence it can follow the linear dimming of the primary constant current source while achieving good equalization of the led string currents at any operating point.

3.2. Constant current regulation

Formula derivation

The main constant current functionality is implemented using an operational amplifier combined with a mosfet, as shown in Fig. 4. The operational amplifier adjusts U_{RS} that is equal to U_{Iset}.

$$U_{\mathrm{Iset}} = R_{\mathrm{S}} I_{\mathrm{LED}} \tag{1}$$

The second order low pass filter reduces the gain at high frequencies. The additional gain G_{CS} of the mosfet and sense resistor can be calculated to:

$$G_{\mathrm{CS}} = g_{\mathrm{M1}} R_{\mathrm{S}} \tag{2}$$

For a simplified calculation it is assumed that the additional gain is frequency independent. For a stable operation the following equation must be fulfilled.

$$G_{\mathrm{CS}}\, G_{\mathrm{OPV}}(f)\, G_{\mathrm{filter}}(f) < 1 \tag{3}$$

The additional required damping of G_{filter} at high frequencies is done by using a second order low pass filter, however also lower and higher orders could be used. The second order low pass filter G_{filter} has 40dB attenuation per decade. The compensated constant current source is shown in Fig. 4.

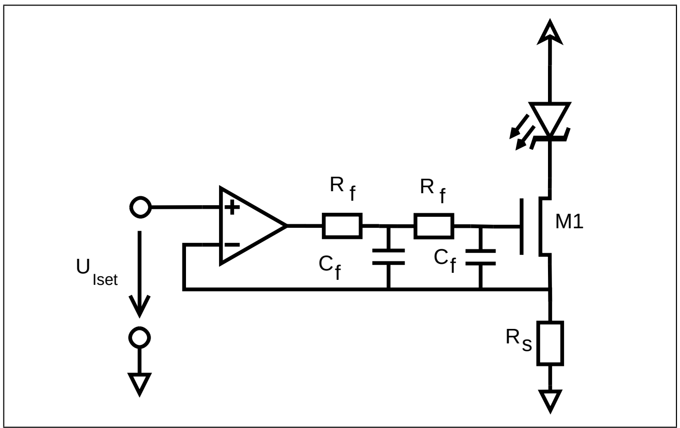

Fig. 4: Compensated constant current source, in this example using a second order low pass filter.

The cutoff frequency f_{B} of the second order low pass is calculated, so that the systems gain at the operational amplifiers GBW frequency is less then 1.

$$f_{\mathrm{B}} = GBW \left(10^{-\frac{G_{\mathrm{CS}}[dB]}{G_{\mathrm{Filter}}[dB]}}\right) \tag{4}$$

The cutoff frequency f_{B} of a coupled second order low pass filter can be calculated using the following formula. By this, the values of R_{f} and C_{f} are chosen.

$$f_{\mathrm{B}} = \frac{1}{4\pi\sqrt{2}R_{\mathrm{f}}C_{\mathrm{f}}} \tag{5}$$

The mosfet M1 switches only in dissipative mode if the individual string current is above the average current, to maintain a safe current. However, when all LED strings have the same forward voltage, negligible conduction loss by the current measurement and by the mosfet can be observed.

Constant current filter design example

For M1 the NTTFS4930 is chosen. From the datasheet a forward transconductance of $S = 19\frac{\mathrm{A}}{\mathrm{V}}$ is denoted. We determine the sense resistor $R_S = 1\Omega$ and calculate the gain.

$$G = R_{\mathrm{S}}S = 19 \tag{6}$$

The amplification is converted into log-scale:

$$20\log(G) = 26\,dB \tag{7}$$

To determine the number of decades of damping the system gain is set in relation to the used second order low pass filter, with an attenuation of 40dB per decade

$$\frac{26\,\mathrm{dB}}{40\,\mathrm{dB}} = 0.65\,dec \tag{8}$$

As an operational amplifier we choose LM324. From the datasheet of the LM324 we know, that the Gain-Bandwidth-Product (GBP) is 1MHz. So we can calculate the boarder frequency for our damping network.

$$f_{\mathrm{B}} = 1\mathrm{MHz}\,10^{-0.65} = 223\,\mathrm{kHz} \tag{9}$$

The frequency for boarder frequency for the converter can be calculated using (5). Choosing $C_{\mathrm{f}} = 1\,\mathrm{nF}$, $R_{\mathrm{f}} = 281\,\Omega$. For a practical value $R_{\mathrm{f}} = 330\,\Omega$ is chosen.

3.3. Averaging Circuit

Conventionally a set current, represented by U_{Iset}, is determined by a higher entity. This approach has two drawbacks: First, a higher instance is required to determine the current. Second, a connection is needed for transporting this information.

Therefore a new method is proposed to determine the set current. Each LED-String current is measured at $R_{\mathrm{S_x}}$ and produces the voltage U_{Rx}. These currents are averaged over the samples and filtered to determine the U_{Iset} voltage. The filter cutoff frequency f_{f} of the averaging circuit should be chosen to be lower than the cutoff frequency $\mathrm{f_B}$ to prevent oscillation, recommended is a factor of 5 to 10.

The filter cutoff frequency can be tuned with the capacitor C_{a}. It can be calculated with the following formula, where n is the number of parallel strings.

$$C_{\mathrm{a}} = \frac{n}{2\pi f_{\mathrm{f}} R_{\mathrm{a}}} \tag{10}$$

Fig. 5 shows the averaging circuit. The circuit shown is implemented in this setup as first order averaging circuit, but a higher order is also possible. The calculated average current is then used as set value U_{Iset}.

Note that the AC/DC power supply from Fig. 1 must operate in constant current mode, to determine the to led module current.

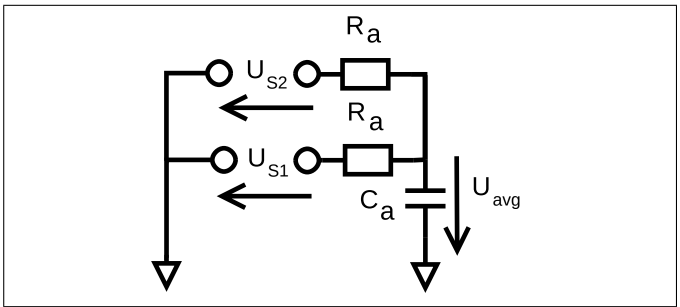

Fig. 5: Averaging circuit determining the set current U_{Iset} using the individual sense voltages.

3.4. Startup Circuit

As U_{avg} is fed back to U_{Iset}, the circuit has two stable operating points. The initial operating point is at 0 A and the second at equally shared current. The startup circuits task is to force a slightly higher current than that the averaging circuit determines. This is done by adding a voltage offset, which is produced by a current source. The required current can be calculated to:

$$I_{\mathrm{startup}} = \frac{U_{\mathrm{Offset}} n}{R_{\mathrm{A}}} \tag{11}$$

In case an increased mismatch current should be accepted by the circuit the following formula can be used.

$$I_{\mathrm{startup}} = \frac{R_{\mathrm{S}} I_{\mathrm{mismatch}} n}{R_{\mathrm{A}}} \tag{12}$$

To generate the startup current, two possible solutions are analyzed, a dual transistor approach and a zener diode approach.

Transistor approach

The transistor approach is shown in Fig. 6. This approach uses two PNP transistors. The following formula can be used to calculate the startup resistor:

$$R_{\mathrm{startup}} = \frac{U_{\mathrm{BE}}}{I_{\mathrm{startup}}} \tag{13}$$

The bias resistor R_{b} is chosen to provide always a minimum operating current for the startup circuit.

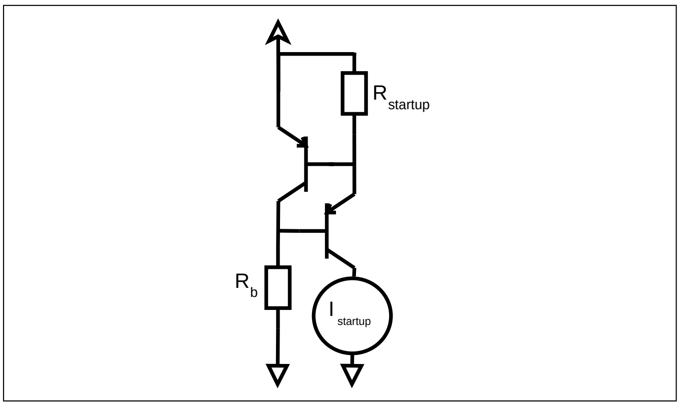

Fig. 6: Transistor constant current source, using two transistors. This constant current approach is known to be temperature sensitive.

This approach is highly temperature sensitive, as the bipolar transistors base emitter voltage U_{BE} changes with temperature.

Zener / Transistor approach

As LED lamps generate heat, it's important to have an improved temperature stability of the startup circuit. The zener constant current approach, shown in Fig. 7, is designed for advanced temperature stability.

Q1 and D1 produce a voltage offset, where R_b is used to bias Q1 and D1. Q2 mirrors the voltage of only D1, which translates the voltage to a current. As the transistor Q2 has a fairly high amplification, the following equation holds true:

$$I_\mathrm{SRC} \approx I_\mathrm{Rstartup} \tag{14}$$

A bipolar transistor has a temperature coefficient of $-2\ \frac{mV}{K}$ [1]. By subtracting this temperature coefficient again, the transistors temperature drift is compensated. As the dual transistors are fabricated on the same die, both temperature coefficients and temperature are likely to be matched. The additional required zener diode should ideally have no temperature variation, thus a 5.6V zener diode [2] is ideal, as temperature drift is centered around $0\ \frac{mV}{K}$.

For the design the following equations hold true:

$$R_\mathrm{src} = \frac{U_\mathrm{D1}}{I_\mathrm{src}} \tag{15}$$

$$R_\mathrm{b} = \frac{U_\mathrm{supply} - U_\mathrm{Q1} - U_\mathrm{D1}}{I_\mathrm{bias}} \tag{16}$$

I_bias is the minimum cathode current of the zener diode specified by the datasheet.

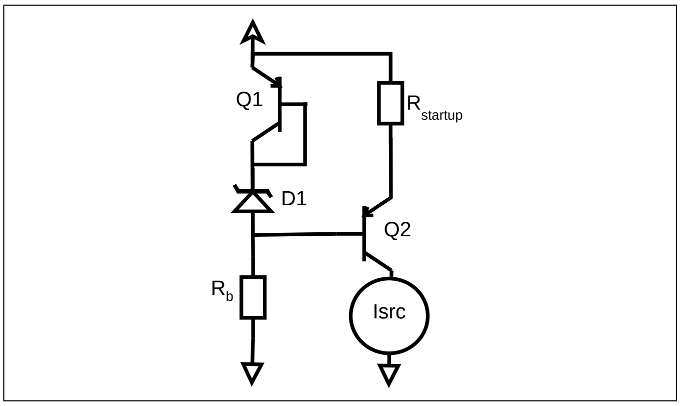

Fig. 7: Zener constant current source, which is temperature compensated using matched Q1 and Q2 same die transistors.

Comparison

The following simulation results of Fig. 8 compare the two solutions. It can be observed that the dual transistor solution has a significant temperature drift, while the zener transistor solution offers decent temperature stability. Therefore the zener transistor solution is favored in the complete design and verified in the section 4.6.

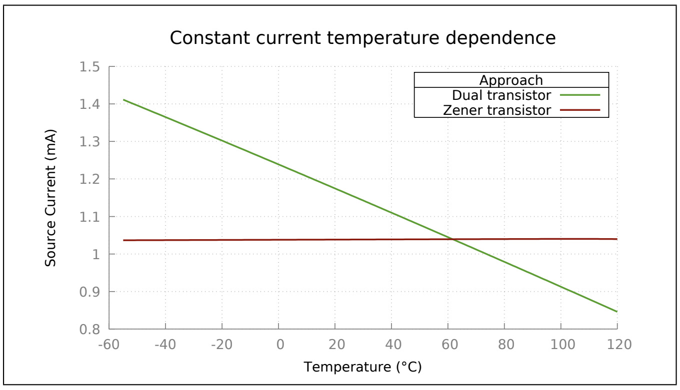

Fig. 8: Temperature dependence of the two constant current sources (spice simulation). It can be observed that the zener transistor solution offers improved temperature stability.

4. Measurements & Validation

The design and its equations are validated by a prototype to prove the proper operational behavior of the circuit.

4.1. Measurement conditions

The measurements where done under the conditions stated in tab. 1, unless not otherwise noted.

The used measurement equipment is stated in table 2.

4.2. Current Distribution

The current distribution over one module is depicted in Fig. 9. It can be observed that the LED currents are matched within a distribution of 7mA.

As a fairly low current sensing resistor is used to save power $(1\,\Omega)$, the current distribution is within allowable limits. The typical operational amplifiers

Parameter	Value
Number of parallel LED strings	24
Number of series LEDs	6
LED forward current	0.12A
LED type used	Lumiled 2835C
Op-amp	LM324
mosfet	NTTFS4930
R_{sense}	1Ω
R_{f}	330Ω
C_{f}	$1\mathrm{nF}$
R_a	4700Ω
C_a	$1\mu\mathrm{F}$
Dual transistor (Q1,Q2)	BC857S
Zener diode	MM3Z5V6
R_{b}	150k
R_{startup}	220k
T_{amb}	20 °C

Tab. 1: Measurement conditions

Type	Model number
Multimeter	Pico Test M3500A
Power Supply	Korad KA3005D

Tab. 2: Used measurement equipment for experiments.

offset (5mV) can be observed. Further a voltage difference between different ground points can be observed: Though a solid ground plane is used, potential differences of up to 3mV at the ground plane could be measured. Therefore strong attention must be given to obtain a solid strong ground plane.

Also the startup circuit allows an additional mismatch current for the LEDs.

For a better matching of LED currents the sense resistors could be increased, compromising efficiency. Using lower offset operational amplifiers can reduce mismatch.

4.3. Voltage Drop over mosfet

The dropout voltage over the mosfet without the sense resistor voltage is depicted in Fig. 10. It can be observed that approx 1/3 of the mosfet operate nearly free of losses. The better the LED strings are matched, the less power loss can be observed. The average voltage drop in this sample was 105mV.

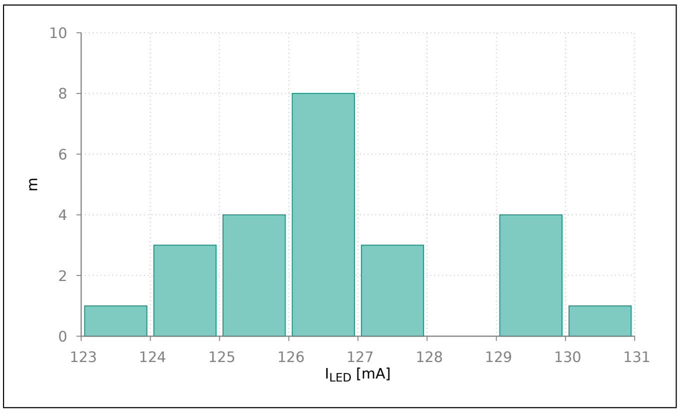

Fig. 9: LED current distribution in the range of 123 mA to 131 mA.

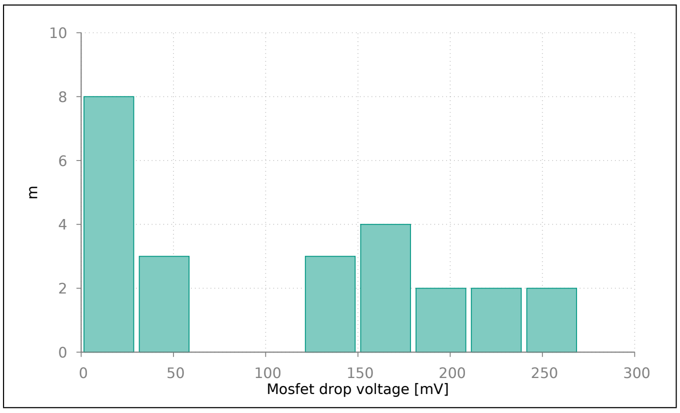

Fig. 10: The voltage drop measured over the mosfet without the sense resistor voltage is depicted. It can be observed that even though one batch is used, a significant difference in voltage drop can be observed.

4.4. Quiescent Current

As additional circuits are introduced, current for the regulation is required that intentionally should be used for lighting. The currents are listed in Tab. 3. However the quiescent currents are fairly low (<4.4 mA) compared to 2.88 A total LED drive current. Therefore the efficiency loss by quiescent currents can be neglected.

4.5. Efficiency

The table 4 depicts the losses. The sensing resistors are the most significant loss source followed by the equalization losses.

Type	Quiescent current	Total
1/4 LM324	175 µA	4.2 mA
Startup circuit	140 µA	0.14 mA
Total		4.4 mA

Tab. 3: Quiescent current of each individual sections and the sum of all loss sources.

Loss source	Value
Quiescent losses	90 mW
Equalizing losses	302 mW
Sensing losses	345 mW
Total	750 mW

Tab. 4: Losses calculation of each functional circuits section. The sensing resistors are the most significant loss source followed by the equalization losses.

4.6. Constant Current Source temperature drift

The constant current source for the startup circuit drifts by heat generated by the LEDs. For a stable performance, the constant current source should have a minimal sensitivity to the LEDs temperature. Fig. 11 demonstrates the temperature dependency of the constant current source I_{SRC}. A typical drift is 2.5% over 60 K, which equals to a temperature drift of 417 ppm/K.

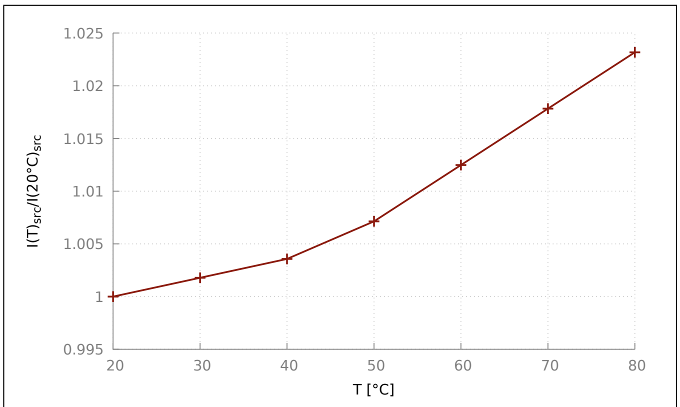

Fig. 11: I_{SRC} temperature drift of the dual transistor temperature compensated constant current startup circuit. (Section 4.6).

4.7. Reliability

So called catastrophic failures of LEDs are rare [7], and if so, typically a short circuit can be observed. When a short circuit failure occurs the additional voltage drop is taken by the mosfet, demonstrated in Fig. 12.

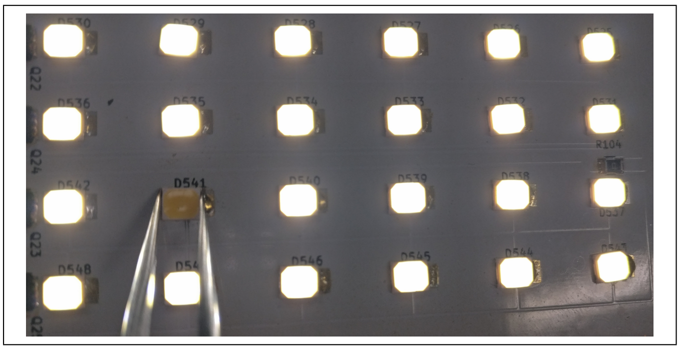

Fig. 12: A short circuit of a LED is compensated without affecting the LED drive current of the other LEDs.

Open circuit failure is not covered by the presented circuit diagram in Fig. 3, as it's not likely. However, if open circuit failure should be covered additional zener diodes anti-parallel to the LEDs may be used.

5. Design considerations

5.1. PCB Layout

LED PCBs are typically fabricated using single sided aluminum PCBs, therefore a layout can be quite complex. A solid ground plane is strongly advised to avoid different ground potentials. Even 3 mV difference can create a current mismatch in the mA range.

Also it's strongly recommended to add overvoltage protection to the system and the mosfet gate. This can be done using TVS or zener diodes.

5.2. Glare requirement

To reduce glare the mid-power LEDs should be positioned close to another, a distance of less then 5mm is recommended.

5.3. Legal

The legal patent situation should be checked in the individual countries for applied patents, especially for Germany.

6. Results

The LED driver can be integrated on a single sided aluminum PCB. It uses minimal board space, as shown in Fig. 13. A minimal PCB space requirement of only $A_{\mathrm{EQ}} = 110\,\mathrm{mm}^2$ per series LED channel has been demonstrated.

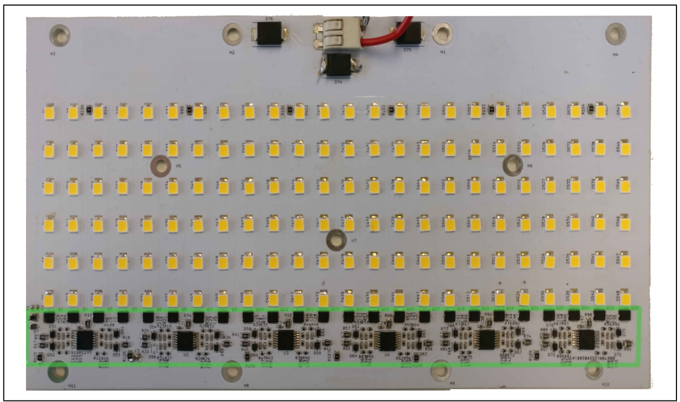

Fig. 13: Aluminum LED pcb board with integrated current sharing controller. The control section is highlighted in green.

A cost of approx. 20ct per series LED channel could be demonstrated, compared to 60ct of a buck converter solution. If the LED strings are constructed equally, minor losses can be observed.

Therefore the dissipative current sharing solution is superior in terms of simplicity, size, cost effectiveness, efficacy and efficiency compared to existing solutions.

7. Summery

In this document an easy paralleling approach for LEDs is demonstrated. In normal operation, the circuit consumes minimal power and balances differences between the individual LED strings accurately within $\pm 2.6\,\%$. In case of an LED short circuit failure the remaining LEDs are protected, as the mosfet of the constant current sink will take the additional voltage drop. The losses of the current sharing circuit are measured to be minimal and narrow current distribution among the parallel LED strings are observed.

As LED produce heat, a temperature stable startup circuit was proposed, allowing tight tolerances.

By this technology high luminance LED arrays with hundreds or thousands of mid-power LEDs can be constructed cost effectively and conventional safety extra low voltage (SELV) LED drivers may still be used.

As mid-power LEDs are typically far more efficient, a significant amount of energy can be saved. By using mid-power LEDs, thermal and glare requirements are easily fulfilled.

8. References

[1] *BC857 Datasheet NXP*.

[2] *Zener Diode BZX84 datasheet*.

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