Chipsmall Limited consists of a professional team with an average of over 10 year of expertise in the distribution of electronic components. Based in Hongkong, we have already established firm and mutual-benefit business relationships with customers from,Europe,America and south Asia,supplying obsolete and hard-to-find components to meet their specific needs.

With the principle of “Quality Parts,Customers Priority,Honest Operation,and Considerate Service”,our business mainly focus on the distribution of electronic components. Line cards we deal with include Microchip,ALPS,ROHM,Xilinx,Pulse,ON,Everlight and Freescale. Main products comprise IC,Modules,Potentiometer,IC Socket,Relay,Connector.Our parts cover such applications as commercial,industrial, and automotives areas.

We are looking forward to setting up business relationship with you and hope to provide you with the best service and solution. Let us make a better world for our industry!

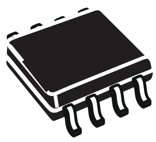
Contact us

Tel: +86-755-8981 8866 Fax: +86-755-8427 6832
Email & Skype: info@chipsmall.com Web: www.chipsmall.com
Address: A1208, Overseas Decoration Building, #122 Zhenhua RD., Futian, Shenzhen, China

LED5000

3 A monolithic step-down current source with dimming capability

Datasheet - production data



HSOP8

Features

- 5.5 V to 48 V operating input voltage range
- 850 kHz fixed switching frequency
- 200 mV typ. current sense voltage drop
- Buck / buck-boost / floating boost topologies
- PWM dimming
- ± 3% output current accuracy overtemperature
- 200 mΩ typical R_{DSON}
- Peak current mode architecture
- Short-circuit protection
- Compliant with ceramic output capacitors
- Inhibit for zero current consumption
- Thermal shutdown

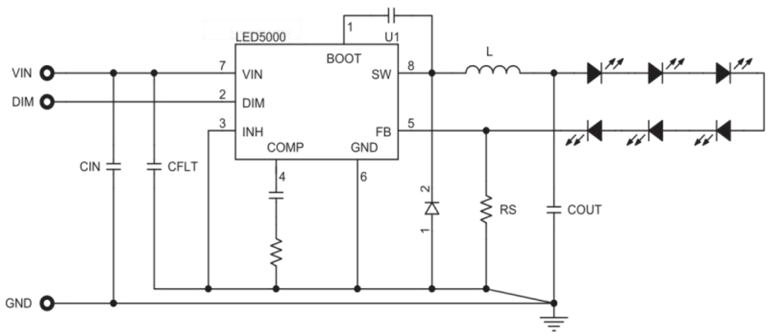
Applications

- High brightness LED driving
- Street lighting
- Signage
- Halogen bulb replacement
- General lighting

Description

The LED5000 device is an 850 kHz fixed switching frequency monolithic step-down DC-DC converter designed to operate as a precise constant current source with an adjustable current capability up to 3 A DC. The embedded PWM dimming circuitry features LED brightness control. The regulated output current level is set by connecting a sensing resistor to the feedback pin. The 200 mV typical R_{SENSE} voltage drop enhances performance in terms of efficiency. The size of the overall application is minimized thanks to the high switching frequency and its compatibility with ceramic output capacitors. The device is fully protected against overheating, overcurrent and output short-circuit. The LED5000 is available in an HSOP8 package.

Figure 1: Typical application circuit



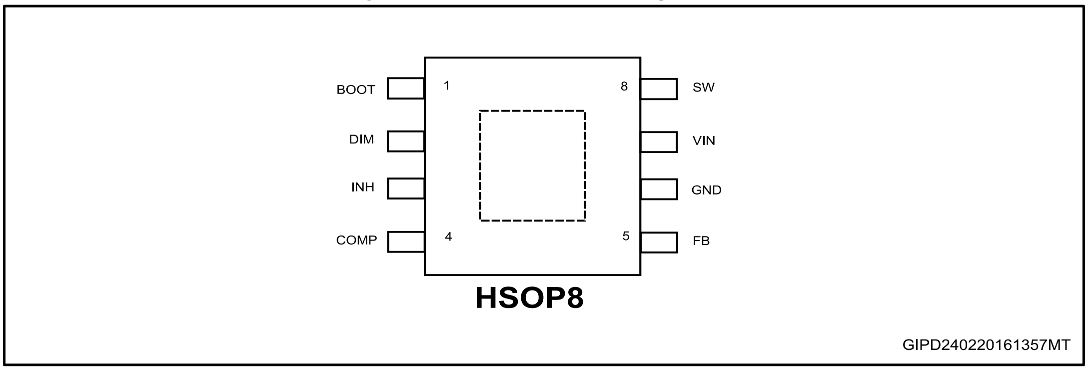
This is information on a product in full production. www.st.com

Contents

1 Pin settings

1.1 Pin connection

Figure 2: Pin connection (top view)



1.2 Pin description

Table 1: Pin description

	Type	Description
1	BOOT	Analog circuitry power supply connection
2	DIM	Dimming control input. Logic low prevents the switching activity, logic high enables it. A square wave on this pin implements LEDs current PWM dimming. Connect to VIN if not used (see *Section 5.8: "Dimming operation"*)
3	INH	Inhibit pin. Connect to GND if not used
4	COMP	Analog circuitry
5	FB	Feedback input. Connect a proper sensing resistor to set the LED current
6	GND	Ground connection
7	VIN	Power input voltage
8	SW	Switching node
-	e.p.	Exposed pad to be connected to GND to increase the package thermal performance and the device noise immunity

2 Maximum ratings

2.1 Maximum ratings

Table 2: Absolute maximum ratings

Symbol	Parameter	Value	Unit
V_{IN}	Power supply input voltage	-0.3 to 52	V
V_{INH}	Inhibit input	-0.3 to 7	V
V_{DIM}	Dimming input	-0.3 to (V_{IN} + 0.3)	V
V_{COMP}	Comp output	-0.3 to 3	V
BOOT	Bootstrap pin	-0.3 to 55	V
SW	Switching node	-1 to (V_{IN} + 0.3)	V
V_{FB}	Feedback voltage	-0.3 to 3	V
T_J	Operating junction temperature range	-40 to 150	°C
T_{STG}	Storage temperature range	-65 to 150	°C
T_{LEAD}	Lead temperature (soldering 10 sec.)	260	°C

2.2 Thermal data

Table 3: Thermal data

Symbol	Parameter	Value	Unit
$R_{th\ JA}$[(1)]	Thermal resistance junction ambient	40	°C/W

Notes:

[(1)]Device soldered to the STEVAL-ILL056V1 demonstration board.

2.3 ESD protection

Table 4: ESD protection

Symbol	Test condition	Value	Unit
ESD	HBM	4	KV
	MM	500	V

3 Electrical characteristics

All tests performed at T_J = 25 °C, V_{CC} = 12 V, V_{INH} = 0 V unless otherwise specified. The specification is guaranteed from -40 to +125 °C - T_J temperature range by design, characterization and statistical correlation.

Table 5: Electrical characteristics

Symbol	Parameter	Test condition		Min.	Typ.	Max.	Unit
V_{IN}	Operating input voltage range			5.5		48	V
$R_{DS(on)}$	MOSFET on resistance	I_{SW} = 1 A			0.2	0.4	Ω
I_{SW}	Maximum limiting current			3.7	4.5	5.2	A
t_{HICCUP}	Hiccup time				16		ms
f_{SW}	Switching frequency			600	850	1000	kHz
	Duty cycle		(1)		90		%
$T_{ON\ MIN}$	Minimum conduction time of the power element		(1)		90		ns
$T_{OFF\ MIN}$	Minimum conduction time of the external diode		(1)	75	90	120	ns
DC characteristics							
V_{FB}	Voltage feedback			194	200	206	mV
I_{FB}	FB biasing current				50		nA
I_q	Quiescent current	V_{DIM} > 1.5 V			1.3	2	mA
		V_{DIM} > 1.5 V, V_{IN} = 48 V			1.7	2.4	mA
I_{qst-by}	Standby quiescent current	V_{INH} > 1.5 V		12	16	34	mA
Inhibit							
V_{INH}	Inhibit levels	Device ON V_{IN} = 5.5 V to 48 V				0.5	V
		Device OFF V_{IN} = 5.5 V to 48 V		1.5			V
I_{INH}	Inhibit biasing current	V_{INH} = 5 V		0.7	1.6	2.5	mA
Dimming							
V_{DIM}	Dimming levels	Switching activity V_{IN} = 5.5 V to 48 V		2.2			V
		Switching activity prevented V_{IN} = 5.5 V to 48 V				0.5	V

Symbol	Parameter	Test condition		Min.	Typ.	Max.	Unit
Error amplifier							
V_{OH}	High level output voltage	V_{FB} = 0 V		3			V
V_{OL}	Low level output voltage	V_{FB} = 400 mV				150	mV
$I_{o\ source}$	Source output current	V_{COMP} = 1.5 V; V_{FB} = 0 V		16	23	30	µA
$I_{o\ sink}$	Sink output current	V_{COMP} = 1.5 V; V_{FB} = 0.4 V		16	23	30	µA
I_b	Source bias current	V_{FB} = 250 mV			50		nA
	DC open loop gain	RL = ∞	(1)		90		dB
gm	Transconductance	I_{COMP} = TBD; V_{COMP} = TBD			220		µS
Thermal shutdown							
T_{SHDWN}	Thermal shutdown temperature		(1)	140	150	160	°C
T_{HYS}	Thermal shutdown hysteresis		(1)		15		°C

Notes:

(1)Parameter guaranteed by design.

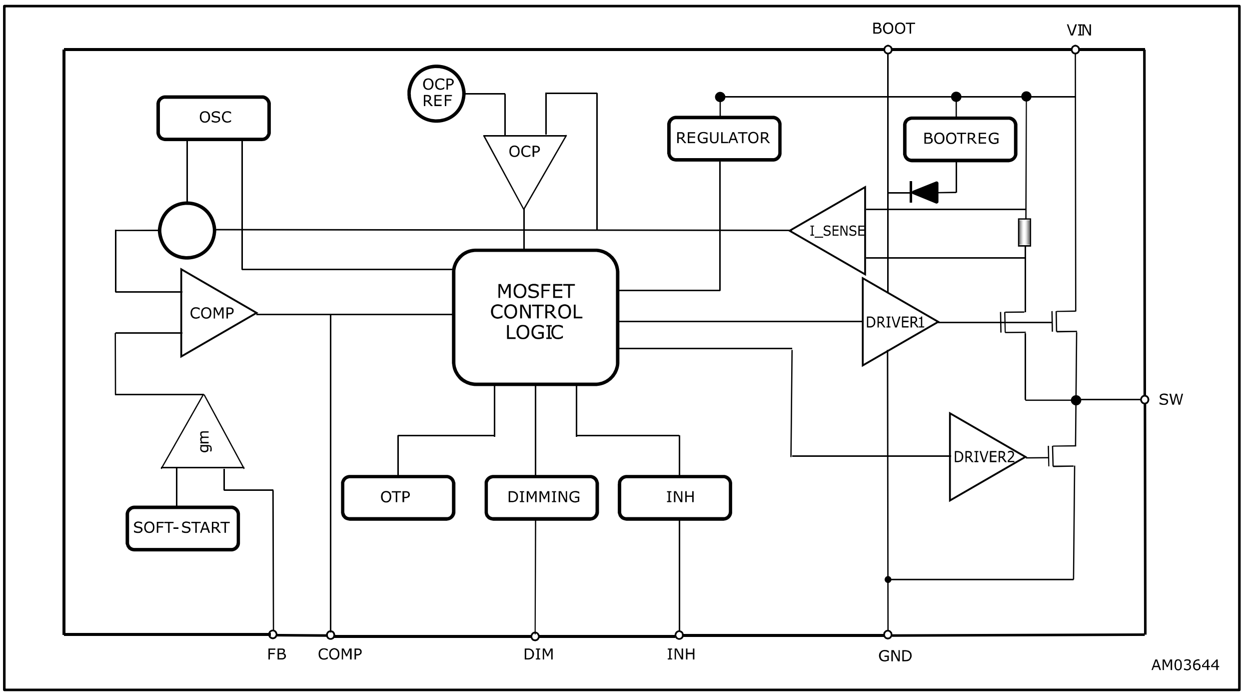
4 Functional description

The LED5000 is based on a “peak current mode” architecture with fixed frequency control. As a consequence the intersection between the error amplifier output and the sensed inductor current generates the control signal to drive the power switch.

The main internal blocks shown in the block diagram in *Figure 3: "block diagram"* are:

- A fully integrated sawtooth oscillator with a typical frequency of 850 kHz
- A transconductance error amplifier
- A high side current sense amplifier to track the inductor current
- A pulse width modulator (PWM) comparator and the circuitry necessary to drive the internal power element
- The soft-start circuitry to decrease the inrush current at power-up
- The dimming block to implement PWM dimming
- The inhibit block for standby operation
- The current limitation circuit based on the pulse-by-pulse current and the HICCUP protection
- The bootstrap circuitry to drive the embedded NMOS switch
- A circuit to implement the thermal protection function

Figure 3: LED5000 block diagram



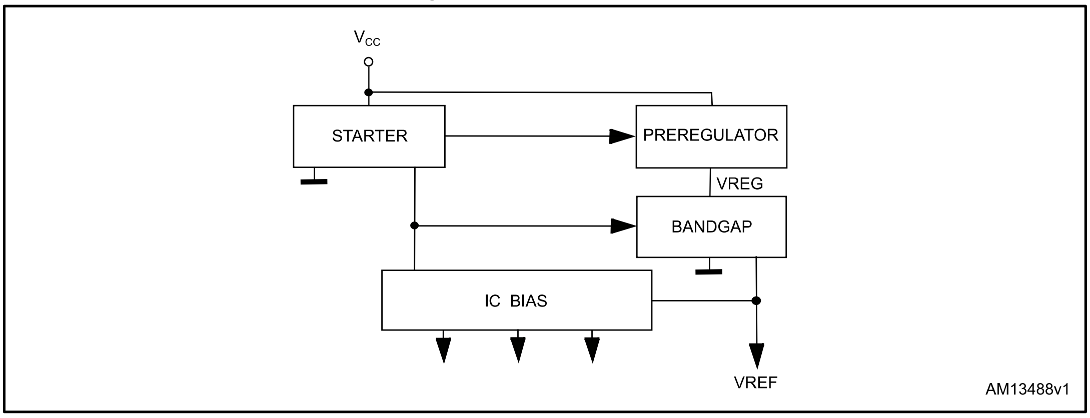
4.1 Power supply and voltage reference

The internal regulator circuit consists of a startup circuit, an internal voltage pre-regulator, the bandgap voltage reference and the bias block that provides current to all the blocks. The starter supplies the startup current to the entire device when the input voltage goes high and the device is enabled (inhibit pin connected to ground). The pre-regulator block supplies the bandgap cell with a pre-regulated voltage that has a very low supply voltage noise sensitivity.

4.2 Voltage monitor

An internal block continuously senses the V_{cc}, V_{ref} and V_{bg}. If the monitored voltages are good, the regulator begins operating. There is also a hysteresis on the V_{cc} (UVLO).

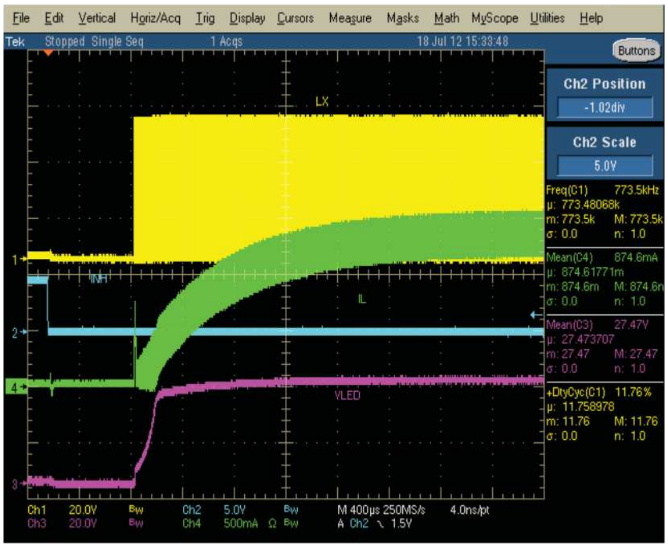
Figure 4: Internal circuit



4.3 Soft-start

The startup phase is implemented ramping the reference of the embedded error amplifier in 1 msec typ. time. It minimizes the inrush current and decreases the stress of the power components at power up.

Figure 5: Soft-start open



During normal operation a new soft-start cycle takes place in case of:

- thermal shutdown event
- UVLO event

The soft-start is disabled during the dimming operation to maximize the dimming performance.

4.4 Dimming block

The DIM input features the LED brightness control with the PWM dimming operation (see *Section 5.8: "Dimming operation"*).

4.5 Inhibit block

The inhibit block features the standby mode accordingly with *Table 5: "Electrical characteristics "*. The INH pin high level disables the device so the power consumption is reduced to less than 40 µA. The INH pin is 5 V tolerant.

4.6 Error amplifier

The voltage error amplifier is the core of the loop regulation. It is a transconductance operational amplifier whose non inverting input is connected to the internal voltage reference (200 mV), while the inverting input (FB) is connected to the output current sensing resistor.

Table 6: Uncompensated error amplifier characteristics

Description	Values
Transconductance	220 µS
Low frequency gain	90 dB

The error amplifier output is compared with the inductor current sense information to perform PWM control.

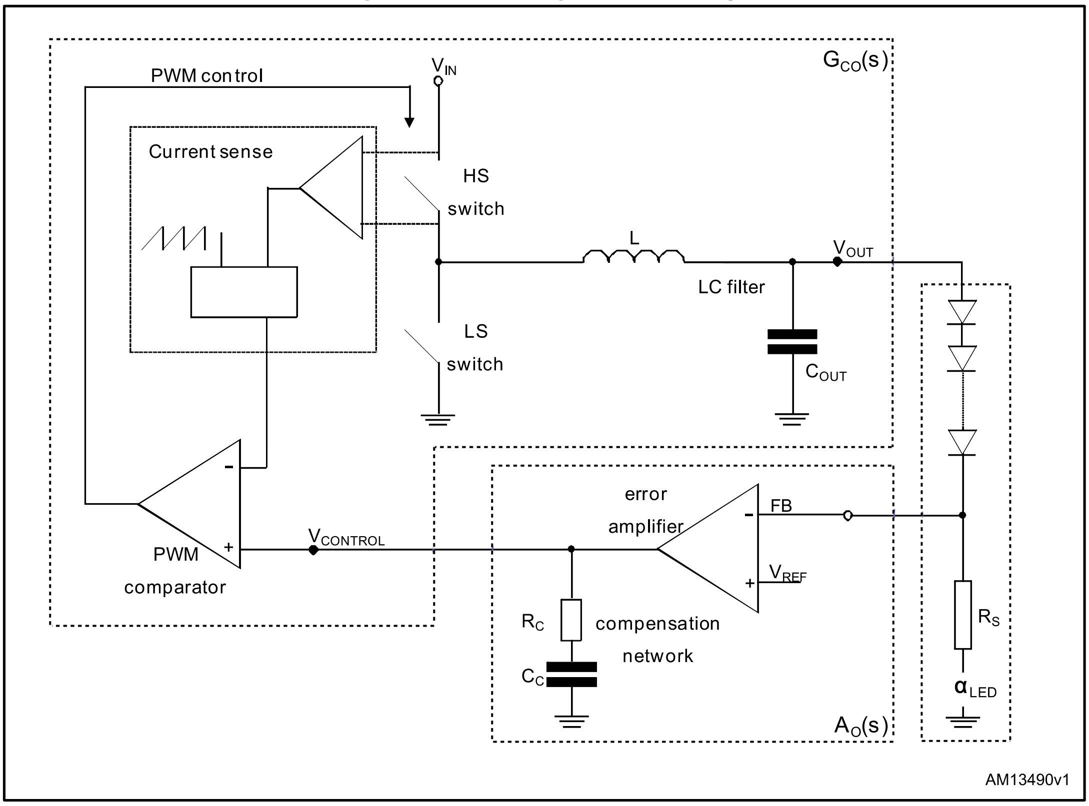
4.7 Thermal shutdown

The shutdown block generates a signal that disables the power stage if the temperature of the chip goes higher than a fixed internal threshold (150 ± 10 °C typical). The sensing element of the chip is close to the PDMOS area, ensuring fast and accurate temperature detection. A 15 °C typical hysteresis prevents the device from turning ON and OFF continuously during the protection operation.

5 Application notes - buck conversion

5.1 Closing the loop

Figure 6: Block diagram of the loop



5.2 $G_{CO}(s)$ control to output transfer function

The accurate control to output transfer function for a buck peak current mode converter can be written as:

Equation 1

$$G_{CO}(s) = \frac{R_{LOAD}}{R_{CS}} \cdot \frac{1}{1 + \frac{R_0 \cdot T_{SW}}{L} \cdot [m_C \cdot (1 - D) - 0.5]} \cdot \frac{\left(1 + \frac{s}{\omega_z}\right)}{\left(1 + \frac{s}{\omega_p}\right)} \cdot F_H(s)$$

where R_{LOAD} represents the load resistance (see *Section 5.4: "LED small signal model"*), R_{CS} the equivalent sensing resistor of the current sense circuitry equal to 0.38, ω_p the single pole introduced by the LC filter and ω_z the zero given by the ESR of the output capacitor.

$F_H(s)$ accounts the sampling effect performed by the PWM comparator on the output of the error amplifier that introduces a double pole at one half of the switching frequency.

Equation 2

$$\omega_z = \frac{1}{ESR \cdot C_{OUT}}$$

where ESR is the equivalent series resistor to the output capacitor.

Equation 3

$$\omega_P = \frac{1}{R_{LOAD} \cdot C_{OUT}} + \frac{m_C \cdot (1-D) - 0.5}{L \cdot C_{OUT} \cdot f_{SW}}$$

where:

Equation 4

$$\begin{cases} m_C = 1 + \dfrac{S_e}{S_n} \\ S_e = V_{pp} \cdot f_{SW} \\ S_n = \dfrac{V_{IN} - V_{OUT}}{L} \cdot R_{CS} \end{cases}$$

S_n represents the slope of the sensed inductor current, S_e the slope of the external ramp (V_{PP} peak to peak amplitude equal to 1.2 V) that implements the slope compensation to avoid sub-harmonic oscillations at duty cycle over 50%.

The sampling effect contribution $F_H(s)$ is:

Equation 5

$$F_H(s) = \frac{1}{1 + \dfrac{s}{\omega_n \cdot Q_P} + \dfrac{s^2}{\omega_n^2}}$$

where:

Equation 6

$$\omega_n = \pi \cdot f_{SW}$$

and

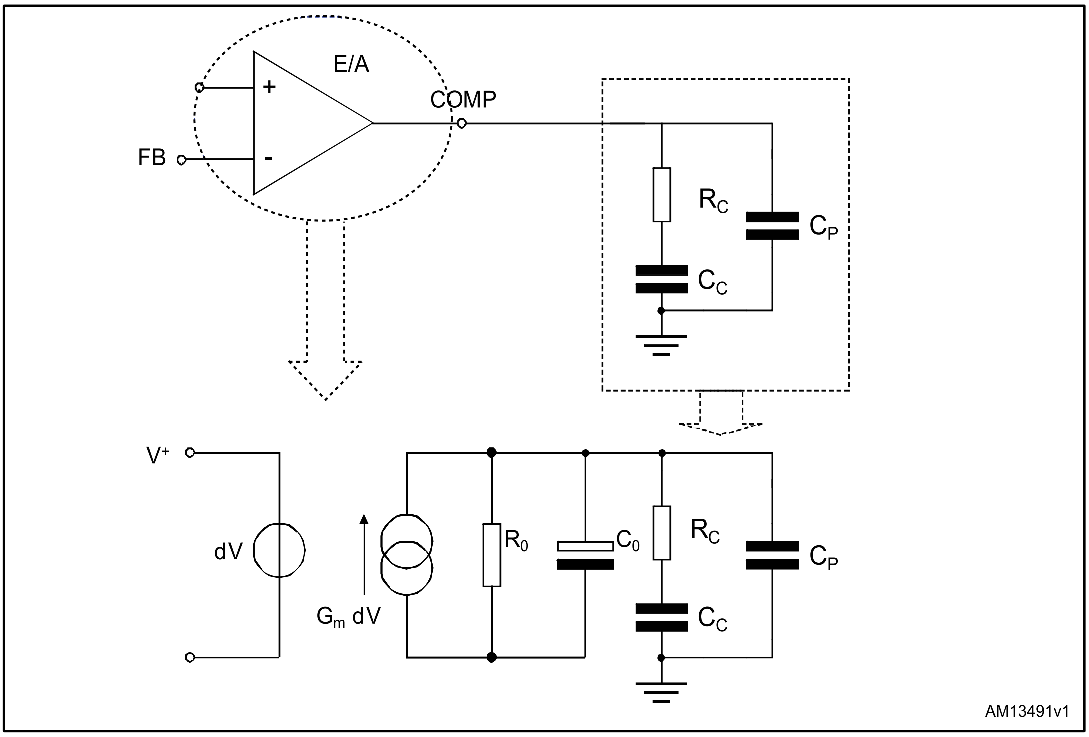
Equation 7

$$Q_P = \frac{1}{\pi \cdot [m_C \cdot (1-D) - 0.5]}$$

5.3 Error amplifier compensation network

The external compensation network connected at the output of the error amplifier is dimensioned to stabilize the system depending on the application conditions.

Figure 7: Transconductance embedded error amplifier



R_C and C_C introduce a pole and a zero in the open loop gain. C_P does not significantly affect system stability but it can be useful to reduce the noise at the output of the error amplifier.

The transfer function of the error amplifier and its compensation network is:

Equation 8

$$A_0(s) = \frac{A_{V0} \cdot (1 + s \cdot R_c \cdot C_c)}{s^2 \cdot R_0 \cdot (C_0 + C_p) \cdot R_c \cdot C_c + s \cdot (R_0 \cdot C_c + R_0 \cdot (C_0 + C_p) + R_c \cdot C_c) + 1}$$

Where $A_{V0} = G_m \cdot R_0$ (R_O = output resistor of OTA = 200 * 10 ^ 6 Ω).

The poles of this transfer function are (if $C_c >> C_0 + C_P$):

Equation 9

$$f_{P\,LF} = \frac{1}{2 \cdot \pi \cdot R_0 \cdot C_c}$$

Equation 10

$$f_{P\,HF} = \frac{1}{2 \cdot \pi \cdot R_c \cdot (C_0 + C_p)}$$

whereas the zero is defined as:

Equation 11

$$F_Z = \frac{1}{2 \cdot \pi \cdot R_c \cdot C_c}$$

5.4 LED small signal model

Once the system reaches the working condition the LEDs composing the row are biased and their equivalent circuit can be considered as a resistor for frequencies << 1 MHz.

The LED manufacturer typically provides the equivalent dynamic resistance of the LED biased at a different DC current. This parameter is required to study the behavior of the system in the small signal analysis.

For instance, the equivalent dynamic resistance of Luxeon III Star from Lumiled measured with different biasing current levels is reported below:

$$r_{LED} \begin{cases} 1.3\Omega & I_{LED} = 350mA \\ 0.9\Omega & I_{LED} = 700mA \end{cases}$$

In case the LED datasheet does not provide the equivalent resistor value, it can be easily derived as the tangent to the diode I - V characteristic in the present working point (see *Figure 8: "Equivalent series resistor"*).

Figure 8: Equivalent series resistor

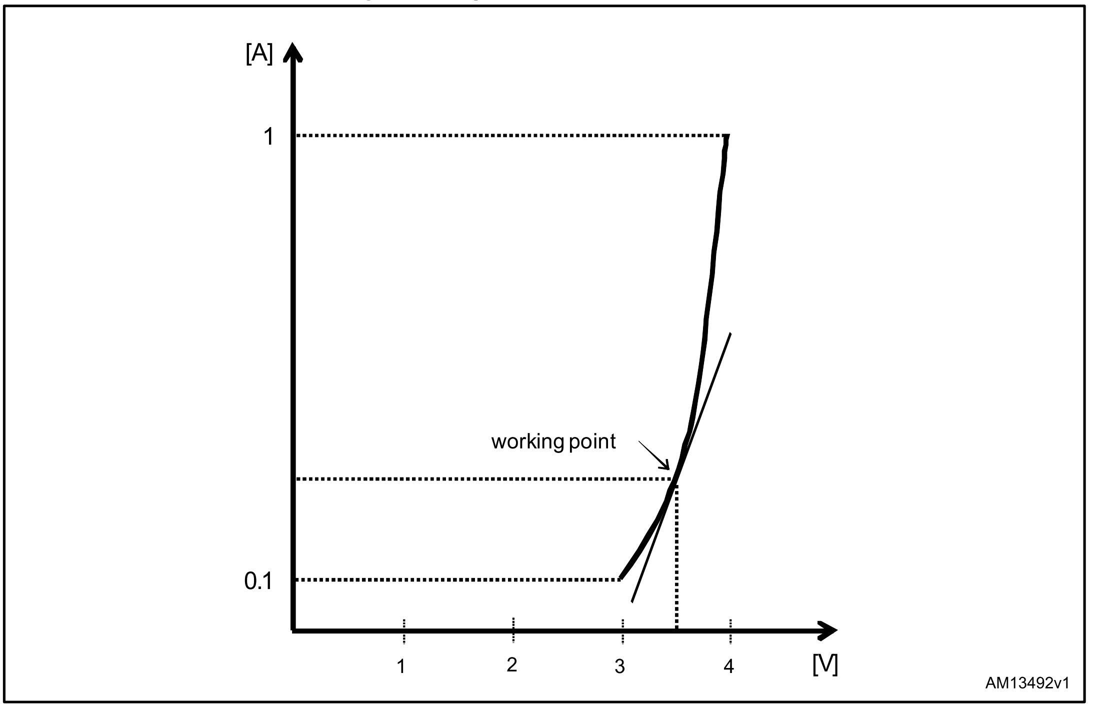


Figure 9: "Load equivalent circuit" shows the equivalent circuit of the LED constant current generator.

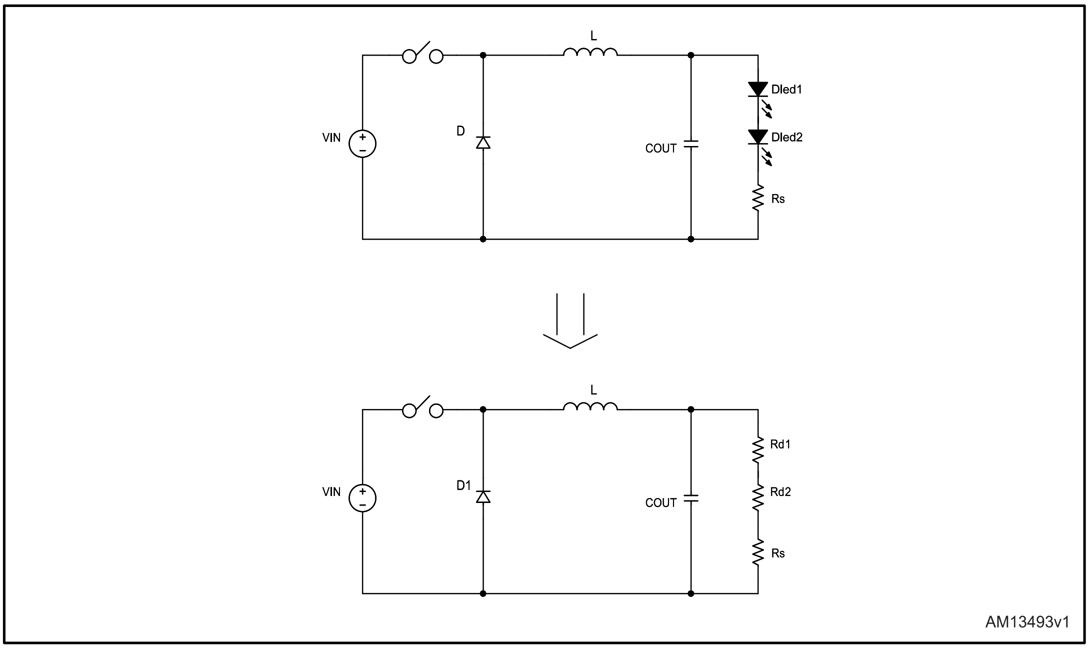
The equivalent loading resistor in the LEDs working point is:

Equation 12

$$R_{LOAD} = n_{LED} \cdot r_{LED} + R_S$$

where R_S is the resistor put in series to the LED string.

Figure 9: Load equivalent circuit



As a consequence the LED equivalent circuit gives the $\alpha_{LED}(s)$ term correlating the output voltage with the high impedance FB input:

Equation 13

$$\alpha_{LED}(n_{LED}) = \frac{R_{SENSE}}{n_{LED} \cdot r_{LED} + R_{SENSE}}$$

5.5 Total loop gain

In summary, the open loop gain can be expressed as:

Equation 14

$$G(s) = G_{CO}(s) \cdot A_0(s) \cdot \alpha_{LED}(n_{LED})$$

5.6 Compensation network design

The maximum bandwidth of the system can be designed up to $f_{SW}/6$ to guarantee a valid small signal model.

Equation 15

$$\frac{\omega_P}{2 \cdot \pi} < BW \le BW_{MAX} = \frac{f_{SW}}{6}$$

where ω_P (*Equation 3*) is the pole introduced by the power components. The following calculations are valid in the hypothesis that $BW > \omega_P$ which is the typical condition.

With the power components selected in accordance with *Section 5.9: "Component selection"* and given the BW specification, the components composing the compensation network can be calculated as:

Equation 16

$$R_C = \frac{1 + \frac{R_{LOAD} \cdot T_{SW}}{L} \cdot [m_C \cdot (1 - D) - 0.5]}{f_P} \cdot \frac{BW \cdot R_{CS}}{G_m \cdot R_S}$$

where the term m_C is represented in *Equation 4*, R_{LOAD} the equivalent loading resistor (*Equation 12*), R_S the resistor put in series to the LED string, G_m the error amplifier transconductance and R_{CS} the equivalent sensing resistor of the current sense circuitry equal to 0.38 (*Table 5: "Electrical characteristics "*).

Equation 17

$$C_C = \frac{K}{R_C \cdot BW}$$

where K represents the leading position of the F_Z (*Equation 11*) with respect to the system bandwidth. In general, a value of 2 gives enough phase margin to the overall small loop transfer function.

5.7 Example of system design

Design specification:

V_{IN} = 48 V, V_{FW_LED} = 3.7 V, n_{LED} = 10, r_{LED} = 1.1 Ω, I_{LED} = 1 A, $I_{LED\ RIPPLE}$ = 2%.

The inductor and capacitor value are dimensioned to meet the $I_{LED\ RIPPLE}$ specification (see *Section 5.9.2: "Inductor and output capacitor selection"* for output capacitor and inductor selection guidelines):

L = 22 µH, C_{OUT} = 1.0 µF mlcc (negligible ESR).

In accordance with *Section 5.9.1: "Sensing resistor"* the sensing resistor value is:

Equation 18

$$R_S = \frac{200\ mV}{1\ A} = 200\ m\Omega$$

Assuming a system bandwidth of:

Equation 19

$$BW = 70\ kHz < BW_{MAX}$$

The ideal values of the components making up the compensation network are:

Equation 20

$$R_C = 43\ k\Omega \qquad C_C = 650\ pF$$

Final component selection is based on commercial values and a small capacitor C_P is added to reduce noise at the error amplifier output. C_P slightly decreases the BW and phase margin.

Equation 21

$$R_C = 47\ k\Omega \qquad C_C = 680\ pF \qquad C_C = 12\ pF$$

The gain and phase margin bode diagrams are plotted, respectively, in *Figure 10: "Module plot"* and *Figure 11: "Phase plot"*.

Figure 10: Module plot

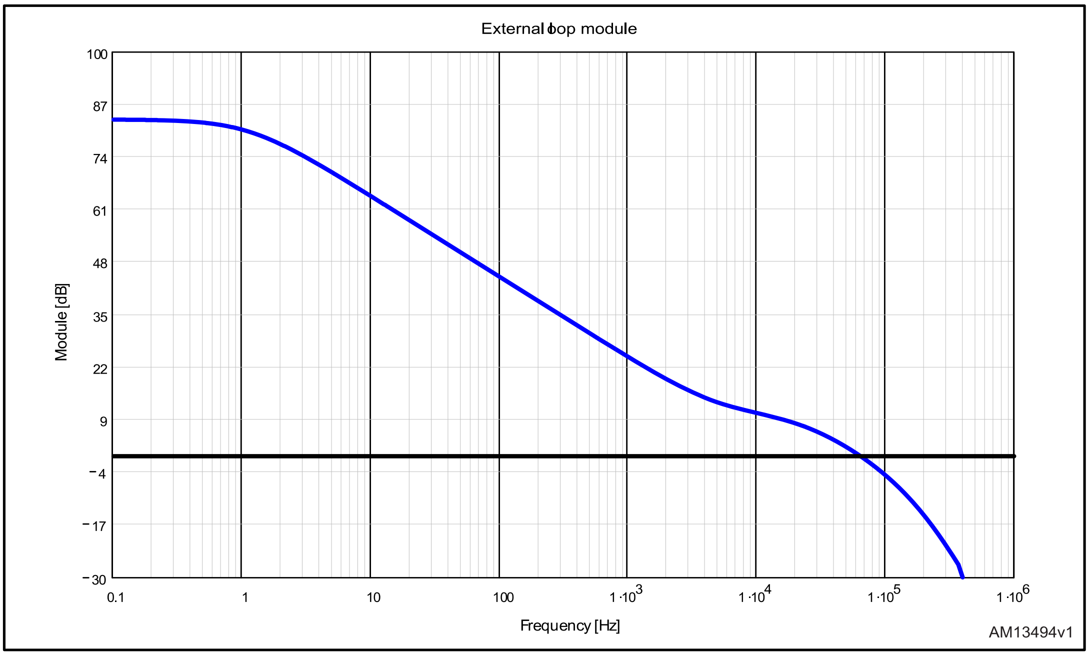
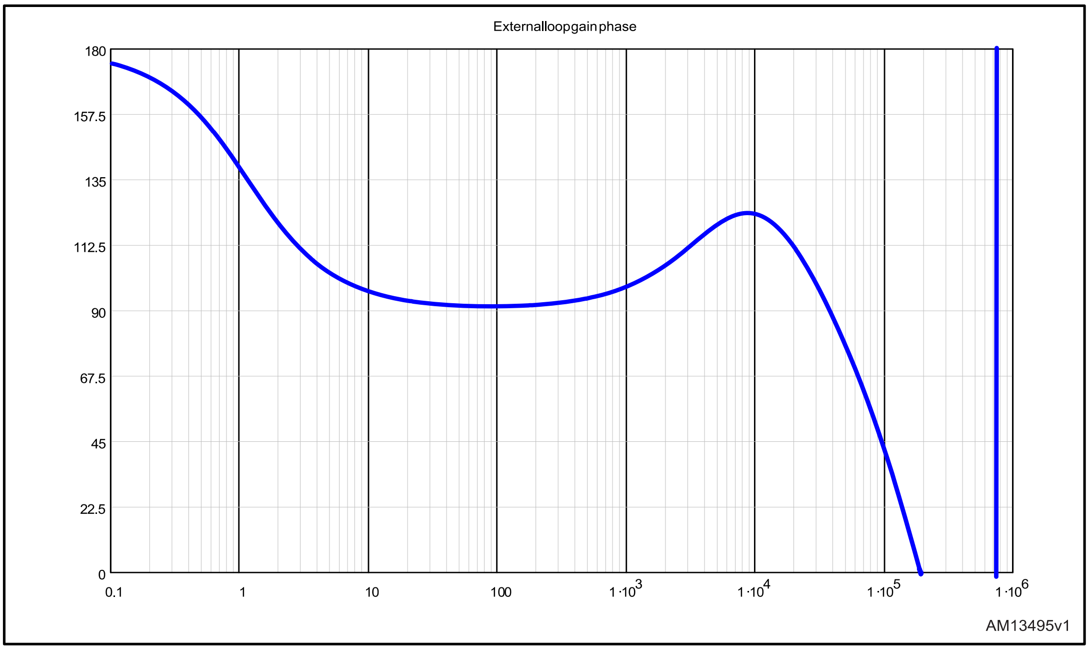


Figure 11: Phase plot



The cut-off frequency and the phase margin are:

Equation 22

$$f_C = 65\ \text{kHz} \qquad pm = 66°C$$

5.8 Dimming operation

The dimming input disables the switching activity, masking the PWM comparator output.

The inductor current dynamic performance when dimming input goes high depends on the designed system response. The best dimming performance is obtained by maximizing the bandwidth and phase margin, when possible.

As a general rule, the output capacitor minimization improves dimming performance.

Figure 12: dimming operation example

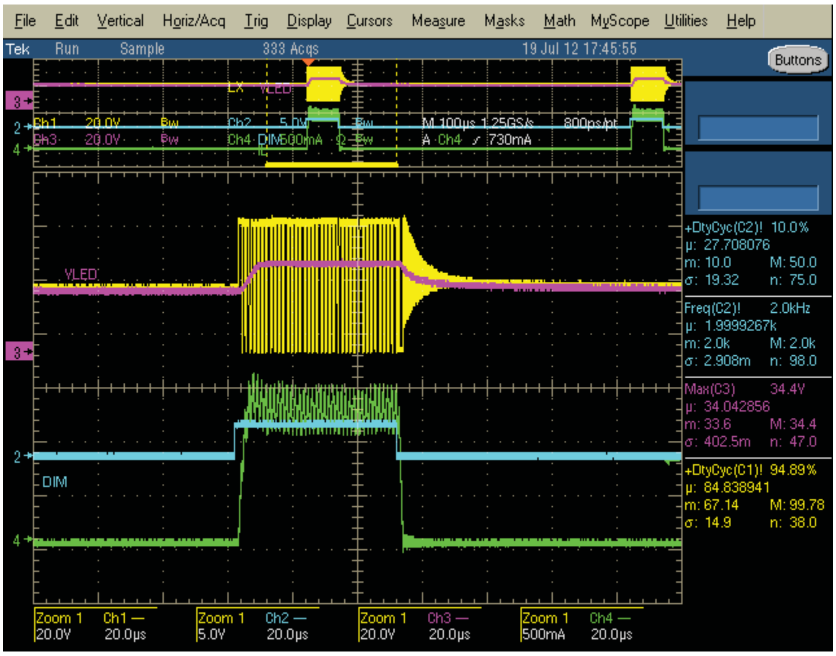
AM13496v1

In fact, when dimming enables the switching activity, a small capacitor value is fast charged with low inductor value. As a consequence, the LEDs current rising edge time is improved and the inductor current oscillation reduced. An oversized output capacitor value requires extra current for fast charge so generating an inductor current overshoot and oscillations.

The switching activity is prevented as soon as the dimming signal goes low. Nevertheless, the LED current drops to zero only when the voltage stored in the output capacitor goes below a minimum voltage determined by the selected LEDs. As a consequence, a big capacitor value makes the LED current falling time worse than a smaller one.

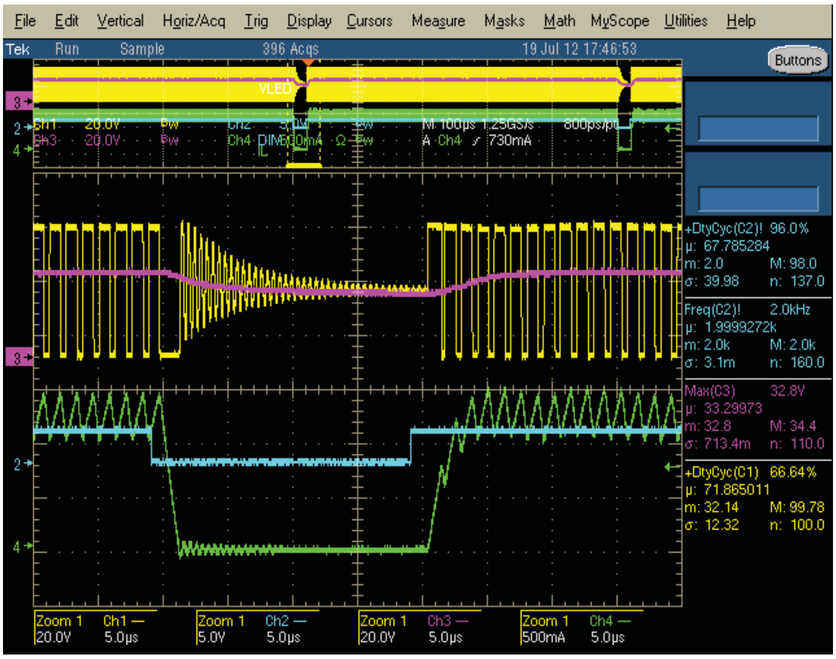
The LED5000 device embeds dedicated circuitry to improve LED current rising edge time.

Figure 13: LED rising edge operation



AM13497v1

Figure 14: LED rising edge operation (zoom)



AM13498v1

5.8.1 Dimming frequency vs. dimming depth

As seen in *Section 5.8: "Dimming operation"* the LEDs current rising and falling edge time mainly depends on the system bandwidth (T_{RISE}) and the selected output capacitor value (T_{RISE} and T_{FALL}).

The dimming performance depends on the minimum current pulse shape specification of the final application. The ideal minimum current pulse has rectangular shape, in any case it degenerates into a trapezoid or, at worst, into a triangle, depending on the ratio (T_{RISE} + T_{FALL})/ T_{DIM}.

Equation 23

$$\text{rectangle} \quad \frac{T_{RISE}+T_{FALL}}{T_{DIM}} \ll 1 \qquad \text{trapezoid} \quad \frac{T_{RISE}+T_{FALL}}{T_{DIM}} < 1 \qquad \text{triangle} \quad \frac{T_{RISE}+T_{FALL}}{T_{DIM}} = 1$$

The small signal response in *Figure 14: "LED rising edge operation (zoom)"* is considered as an example.

Equation 24

$$\begin{cases} T_{RISE} \approx 5\mu s \\ T_{FALL} \approx 2\mu s \end{cases}$$

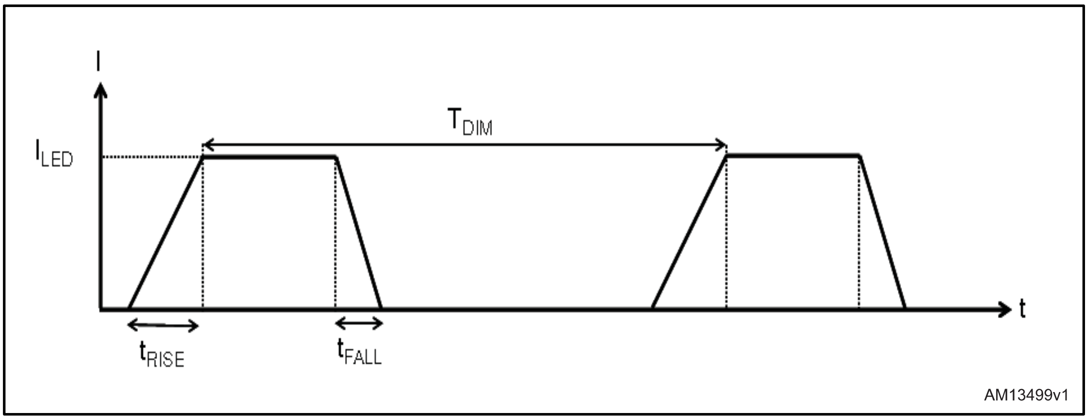
Assuming the minimum current pulse (T_{MIN_PULSE}) shape specification as:

Equation 25

$$T_{RISE} + T_{FALL} = 0.75 \cdot T_{MIN\_PULSE} = 0.75 \cdot D_{MIN} \cdot T_{DIMMING}$$

where $T_{DIMMING}$ represents the dimming period and D_{MIN} the minimum duty cycle which gives the T_{MIN_PULSE} charge. In the given example T_{MIN_PULSE} = 9 µs.

Figure 15: Dimming signal



Given T_{MIN_PULSE} it is possible to calculate the maximum dimming depth given the dimming frequency or vice versa.

For example, assuming a 10 KHz dimming frequency the maximum dimming depth is 9% or given a 5% dimming depth it follows a 5.5 KHz maximum f_{DIM}.

The LED5000 dimming performance is strictly dependent on the system small signal response. As a consequence, an optimized compensation network (good phase margin and bandwidth maximized) and minimized C_{OUT} value are crucial for best performance. Once the external power components and the compensation network are selected, a direct measurement to determine T_{RISE}, T_{FALL} (see *Equation 24*) is necessary to certify the achieved dimming performance.

5.9 Component selection

5.9.1 Sensing resistor

In closed loop operation the LED5000 feedback pin voltage is 200 mV, so the sensing resistor calculation is expressed as:

Equation 26

$$R_S = \frac{200\ mV}{I_{LED}}$$

Since the main loop (see *Section 5.1: "Closing the loop"*) regulates the sensing resistor voltage drop, the average current is regulated into the LEDs. The integration period is at minimum

5 * T_{SW} since the system bandwidth can be dimensioned up to $f_{SW}/5$ at maximum.

A system loop based on a peak current mode architecture features consistent advantages in comparison with simpler closed loop regulation schemes like the hysteretic or the constant ON/OFF control.

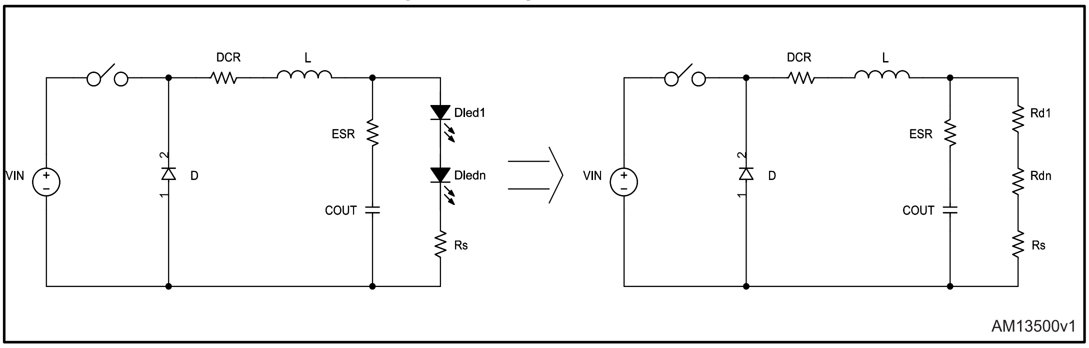
The system performs the output current regulation over a period which is at least five times longer than the switching frequency. The output current regulation neglects the ripple current contribution and its reliance on external parameters like input voltage and output voltage variations (line transient and LED forward voltage spread). This performance cannot be achieved with simpler regulation loops like hysteretic control.

For the same reason, the switching frequency is constant over the application conditions, that helps to tune the EMI filtering and to guarantee the maximum LED current ripple specification in the application range. This performance cannot be achieved using constant ON/OFF time architectures.

5.9.2 Inductor and output capacitor selection

The output capacitor filters the inductor current ripple that, given the application condition, depends on the inductor value. As a consequence the LED current ripple, that is the main specification for a switching current source, depends on the inductor and output capacitor selection.

Figure 16: Equivalent circuit



The LED ripple current can be calculated as the inductor ripple current ratio flowing into the output impedance using the Laplace transform (see *Figure 11: "Phase plot"*):

Equation 27

$$\Delta I_{RIPPLE}(s) = \frac{\frac{8}{\pi^2} \cdot \Delta I_L \cdot (1 + s \cdot ESR \cdot C_{OUT})}{1 + s \cdot (R_S + ESR + n_{LED} \cdot R_{LED}) \cdot C_{OUT}}$$

where the term $8/\pi^2$ represents the main harmonic of the inductor current ripple (which has a triangular shape) and ΔI_L is the inductor current ripple.

Equation 28

$$\Delta I_L = \frac{V_{OUT}}{L} \cdot T_{OFF} = \frac{n_{LED} \cdot V_{FW\_LED} + 200mV}{L} \cdot T_{OFF}$$

so L value can be calculated as:

Equation 29

$$L = \frac{n_{LED} \cdot V_{FW\_LED} + 200mV}{\Delta I_L} \cdot T_{OFF} = \frac{n_{LED} \cdot V_{FW\_LED} + 200mV}{\Delta I_L} \cdot \left(1 - \frac{n_{LED} \cdot V_{FW\_LED} + 200mV}{V_{IN}}\right)$$

where T_{OFF} is the OFF time of the embedded high switch, given by 1 - D.

As a consequence the lower is the inductor value (so higher the current ripple), the higher would be the C_{OUT} value to meet the specification.

A general rule to dimension L value is:

Equation 30

$$\frac{\Delta I_L}{I_{LED}} \leq 0.5$$

Finally the required output capacitor value can be calculated equalizing the LED current ripple specification with the module of the Fourier transformer (see *Equation 27*) calculated at f_{SW} frequency.

Equation 31

$$|\Delta I_{RIPPLE}(s=j\,.\,\omega)| = \Delta I_{RIPPLE\_SPEC}$$

(see *Section 5.6: "Compensation network design"*):

$V_{IN} = 48$ V, $I_{LED} = 700$ mA, $\Delta I_{LED}/I_{LED} = 2\%$, $V_{FW\_LED} = 3.7$ V, $n_{LED} = 10$.

A lower inductor value maximizes the inductor current slew rate for better dimming performance. *Equation 30* becomes:

Equation 32

$$\frac{\Delta I_L}{I_{LED}} = 0.5$$

which is satisfied selecting a 10 µH inductor value.

The output capacitor value has to be dimensioned according to *Equation 31*.

Finally, given the selected inductor value, a 1 µF ceramic capacitor value kesvg the LED current ripple ratio lower than the 2% of the nominal current. An output ceramic capacitor type (negligible ESR) is suggested to minimize the ripple contribution given a fixed capacitor value.

Table 7: Inductor selection

Manufacturer	Series	Inductor value (µH)	Saturation current (A)
Wurth Elektronik	WE-HCI 7040	1 to 4.7	20 to 7
	WE-HCI 7050	4.9 to 10	20 to 4.0
Coilcraft	XPL 7030	2.2 to 10	29 to 7.2