AN4501
Application note

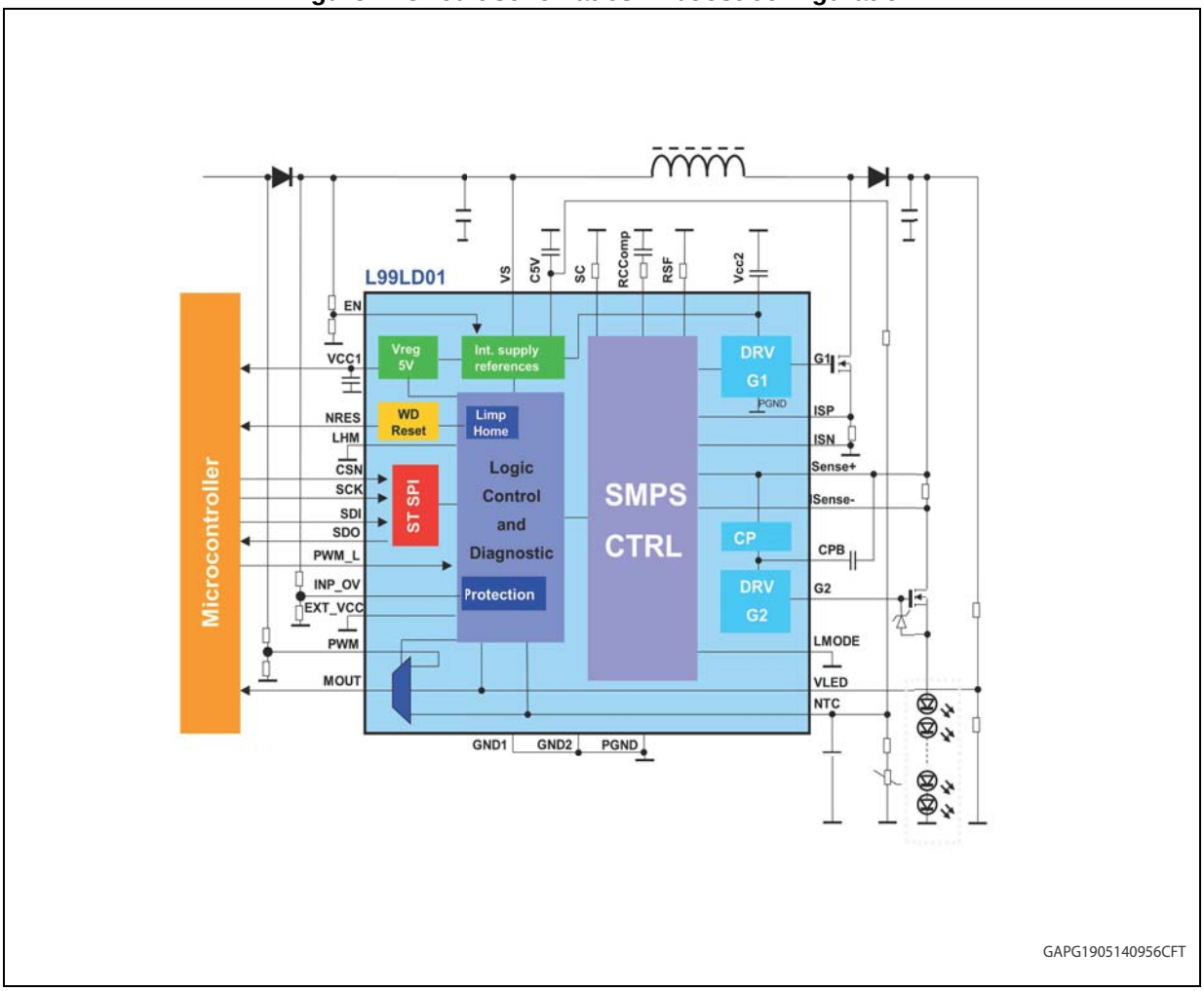
Design of a boost LED driver using L99LD01

Introduction

The L99LD01 is a boost controller dedicated to the control of high-brightness LEDs in automotive headlight applications.The device offers high software configurability thanks to its SPI interface. This scalable solution enables a cost-optimized selection of the power components and provides full diagnostics and protection for enhanced system reliability.

Moreover, the L99LD01 can supply a microcontroller and control its reset input, while a watchdog and a limp home input support safety relevant functions.

Figure 1. Circuit schematics in boost configuration



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1 Overview

This application note provides a guideline and an example of dimensioning of the power component around the boost controller using the L99LD01. This guideline takes into account the wide input voltage range of automotive applications and its implication on the device selection.

In a second part, additional information on the component dimensioning and the details of the calculations are provided.

2 Guideline for the selection of external components

This section provides a guideline for the dimensioning power components of the boost converter. The guideline takes into account the impact of the wide range of the input voltage on the component stress.

The reader can find the list of abbreviations in *Table 4*.

2.1 Guideline

Step 1: Calculate the extreme duty cycles and minimum converter on-time

- Calculate the minimum and maximum duty cycles D_{MIN} and D_{MAX}
- Verify that the operation is compatible with the converter minimum on-time ($t_{ON,MIN}$)
- Verify that D_{MAX} does not exceed the converter duty cycle limitation

Step 2: Selection of the inductor

- Calculate the maximum inductor DC current $I_{L,MAX}$
- Calculate the minimum inductance to comply with the requested ratio inductance current ripple to DC current at the minimum input voltage ($I_{L,PP@VINMIN}$ / $I_{L,MAX}$)
- Select the standardized inductor value
- Recalculate $I_{L,PP@VINMIN}$ with the selected standardized inductance value

Step 3: Selection of the freewheel diode

- Calculate the minimum breakdown voltage and the required current capability

Step 4: Selection of the output capacitor

- Choose a voltage capability, which is higher than the overvoltage protection
- Calculate the required output capacitance fulfilling the maximum output current ripple

Step 4: Selection of the input capacitor

- Calculate the minimum capacitance fulfilling the required input voltage ripple

Step 5: Selection of the switching MOSFET

- Choose a MOSFET and determine the corresponding switching losses (at $V_{IN,MIN}$)
- Calculate the MOSFET RMS current
- Make an iterative calculation of the total power dissipation (at $V_{IN,MIN}$)
- Verify that the maximum junction temperature does not exceed the MOSFET maximum rating.

2.2 Application example

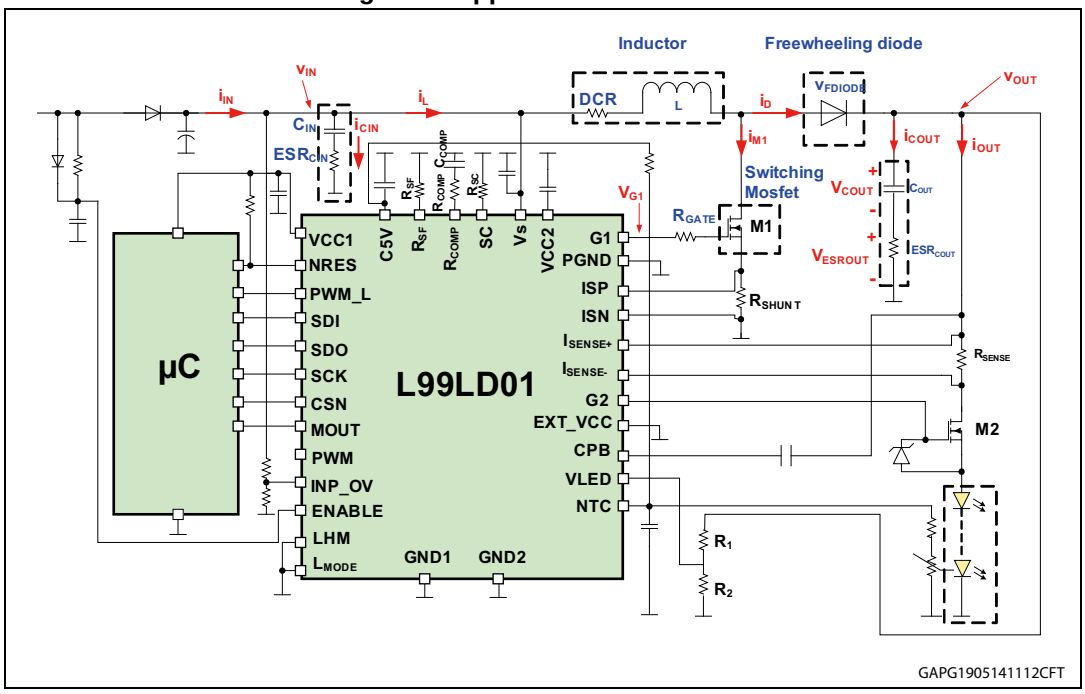
Table 1. Example of application conditions and requirements

Parameter	Value	Comments
V_{IN}	6 V to 18 V	Operating input voltage
I_{OUT}	1 A	Output current
V_{OUT}	2 6V	V_{OUT} ~ LED voltage

Table 1. Example of application conditions and requirements (continued)

Parameter	Value	Comments
$V_{OUT,OVTH}$	42 V	Output over-voltage threshold
r_{MAX}	0.5	Max. ratio at $V_{IN,MIN}$ between inductor current ripple ($I_{L,PP@VINMIN,TARGET}$) and inductor average current (I_L)
F_{SW}	450 kHz	Switching frequency
$I_{OUT,PP,MAX,TARGET}$	75 mA	Target maximum output current ripple
R_{DLED}	0.4 Ω	Dynamic resistance of one LEDs
N_{bLED}	8	Number of LEDs in series
$V_{IN,PP,MAX,TARGET}$	100 mV	Target maximum peak-to-peak input voltage ripple
$R_{THAMB,M1}$	25 K/W	Thermal resistance of the switching resistor M1
$T_{AMB,MAX}$	85 °C	Max. ambient temperature
$R_{DSON,25}$	16 mΩ	M1 RDSON at 25°C
h	90 %	Estimated converter's efficiency

Figure 2. Application schematics



Step 1: Calculation of the duty cycle

$$D_{MIN} = 1 - \frac{V_{INMAX} \times \eta}{V_{OUT}} = 1 - \frac{18 \times 0.9}{26} = 37.7\%$$

The minimum duty cycle is compatible with the device specification (datasheet parameter: T_{ON_MIN}, 14% maximum specification see *Appendix C: Document management*) of the converter is respected.

$$D_{MAX} = 1 - \frac{V_{INMIN} \times \eta}{V_{OUT}} = 1 - \frac{6 \times 0.9}{26} = 79.1\%$$

The maximal duty cycle of the L99LD01 is respected as well (datasheet parameter: Duty Cycle, 88% minimum specification).

Step 2: Selection of the inductor

Required inductance value:

The maximum inductor DC current is given by:

$$I_{L,MAX} = \frac{I_{OUT}}{1 - D_{MAX}} = \frac{1.0}{1 - 0.791} = 4.81A$$

According to the definition of the inductor current ripple r_{MAX}, we have:

$$I_{L,PP@VINMIN,TARGET} = r_{MAX} \times I_{L,MAX} = 0.5 \times 4.81 = 2.41A$$

The minimum inductance value is given by:

$$L_{MIN} = \frac{V_{OUT} \times D_{MAX} \times (1 - D_{MAX})}{F_{SW} \times I_{L,PP@VINMIN,TARGET} \times \eta} = \frac{26 \times 0.791 \times (1 - 0.791)}{450 \cdot 10^3 \times 2.41 \times 0.9} \sim 4.4\mu H$$

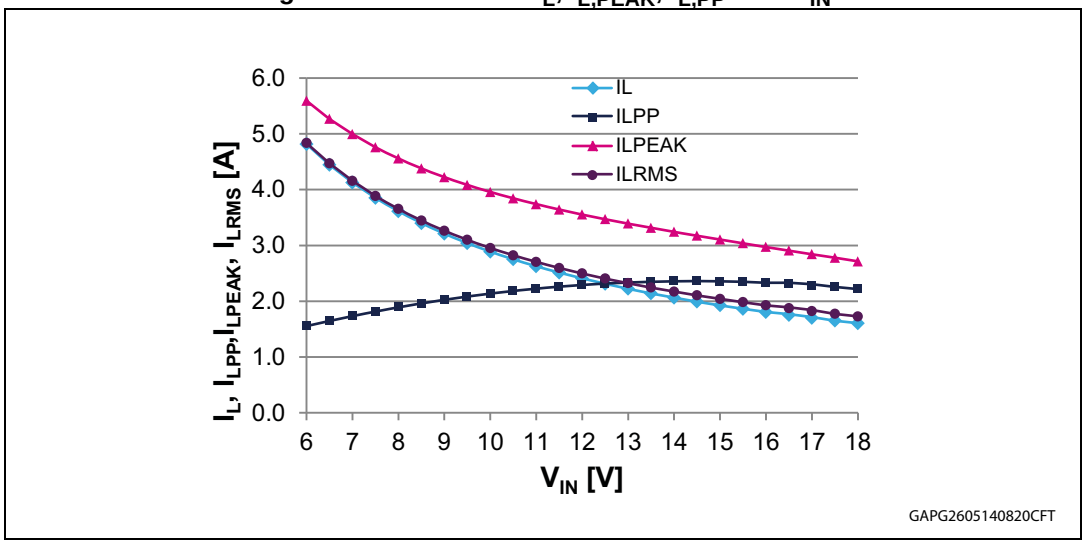
An inductance of 6.8µH can be selected, considering a tolerance of +/- 20%.

Inductor maximum peak and RMS current

$$I_{L,PEAK,MAX} = \frac{I_{OUT}}{1 - D_{MAX}} + \frac{V_{OUT} \times D_{MAX} \times (1 - D_{MAX})}{2 \times F_{SW} \times L \times \eta}$$

$$I_{L,PEAK,MAX} = \frac{1}{1 - 0.791} + \frac{26 \times 0.791 \times (1 - 0.791)}{2 \times 6.8 \cdot 10^{-6} \times 450 \cdot 10^3 \times 0.9} = 5.59A$$

Figure 3. Evolution of I_L, $I_{L,PEAK}$, $I_{L,PP}$ with V_{IN}



Considering the selected standardized inductance, the inductor maximum current ripple is:

$$I_{L,PP@VINMAX} = \frac{26 \times 0.791 \times (1 - 0.791)}{6.8 \cdot 10^{-6} \times 450 \cdot 10^{3} \times 0.9} = 1.55A$$

The inductor maximum RMS current is:

$$I_{L,RMS,MAX} = \sqrt{I_{L,MAX}^{2} + \frac{I_{L,PP@VIN,MIN}^{2}}{12}} = \sqrt{4.81^{2} + \frac{1.55^{2}}{12}} = 4.84A$$

The 6.8 µH inductor must have an RMS current and peak current of at least 4.84 A and 5.59 A.

Step 3: Selection of the freewheeling diode

The minimum diode breakdown voltage with 20% margin is:

$$V_{BRMIN} = 1.2 \times V_{OUT,OVTH} = 1.2 \times 42 = 50V$$

The next standardized breakdown voltage for a Schottky diode is 60 V.

It must be able to withstand the inductor peak current (5.59 A) and an average current which is equal to the LED current (1.0 A).

Step 4: Selection of the output capacitor

Calculation of the minimum required capacitance

In general multilayer ceramic capacitors (MLCC) allow neglecting the ESR contribution to the output voltage and current ripples. Assuming that the output current ripple is mainly due to the bulk capacitance, the minimum required output capacitance is:

$$C_{OUT,MIN} = \frac{1}{Nb_{LED} \times R_{DLED}} \times \frac{I_{OUT} \times D_{MAX}}{I_{OUT,PP,MAX,TARGET} \times F_{SW}}$$

$$C_{OUT,MIN} = \frac{1}{8 \times 0.4} \times \frac{1.0 \times 0.791}{0.075 \times 450 \cdot 10^{3}} = 7.34\mu F$$

Two 4.7 µF MLCC with a voltage capability of 50 V and with an ESR of 4 mΩ can be placed in parallel. The resulting ESR is 2 mΩ (ESR_{COUT}). The resulting maximum output current ripple due to the bulk capacitance is:

$$I_{OUT,PP,MAX} = \frac{1}{Nb_{LED} \times R_{DLED}} \times \frac{I_{OUT} \times D_{MAX}}{C_{OUT} \times F_{SW}} =$$

$$= \frac{1}{8 \times 0.4} \times \frac{1.0 \times 0.791}{9.4 \cdot 10^{-6} \times 450 \cdot 10^{3}} = 58.5mA$$

The additional current ripple caused by the ESR is:

$$\frac{ESR_{COUT} \times I_{L,PEAK,MAX}}{Nb_{LED} \times R_{DLED}} = \frac{2 \cdot 10^{-3} \times 5.59}{8 \times 0.4} = 4.7mA$$

which is indeed negligible compared to $I_{OUT,PP,MAX}$.

The total output current ripple is 62 mA, fulfilling the target.

Step 5: Selection of the input capacitor

We assume that MLCC with a very low ESR is used. The contribution of the ESR to the input voltage ripple can be neglected as well, compared to the contribution of the charge and discharge of the input capacitor itself.

The maximum inductor peak-to-peak current $I_{L,PP,MAX}$ is:

$$I_{L,PP,MAX} = \frac{V_{OUT} \times D_{50} \times (1 - D_{50})}{L \times F_{SW} \times \eta} = \frac{26 \times 0.5^{2}}{6.8 \cdot 10^{-6} \times 450 \cdot 10^{3} \times 0.9} = 2.36A$$

The minimum input capacitor is calculated by:

$$C_{IN,MIN} = \frac{I_{L,PP,MAX}}{8 \times F_{SW} \times V_{IN,PP,MAX,TARGET}} = \frac{2.36}{8 \times 450 \cdot 10^{3} \times 0.100} = 6.6\mu F$$

Using two 4.7 µF, 50 V input capacitor with an ESR of 4mΩ (ESR_{CIN}), the maximum input voltage ripple caused by the bulk capacitance is:

$$\frac{I_{L,PP,MAX}}{8 \times F_{SW} \times C_{IN}} = \frac{2.36}{8 \times 450 \cdot 10^{3} \times 9.4 \cdot 10^{-6}} = 70mV$$

The additional input voltage ripple caused by the ESR is:

$$ESR_{CIN} \times I_{L,PP,MAX} = 2 \cdot 10^{-3} \times 2.36 = 4.8mV$$

which is indeed negligible compared to 70 mV. The sum of both contributions is below the maximum target of 100 mV ($\Delta V_{IN,MAX}$).

Step 6: Selection of the switching MOSFET

The minimum required breakdown voltage is the same as the breakdown voltage of the diode.

A 60 V, 16 mΩ max @ 25°C is considered.

Estimation of the switching losses

The turn-on and turn-off times of the considered MOSFET with is estimated to 20 ns.

The maximum switching losses are given by:

$$P_{M1,SW} = \frac{V_{OUT} \times I_{L,MAX} \times F_{SW}}{2} \times (t_{M1,SWON} + t_{M1,SWOFF})$$

$$P_{M1,SW} = \frac{26 \times 4.81 \times 450 \cdot 10^{3}}{2} \times (20 \cdot 10^{-9} + 20 \cdot 10^{-9}) = 1.13W$$

Calculation of the max. RMS current

The MOSFET's maximum RMS current is estimated by:

$$I_{M1,RMS,MAX} = \frac{I_{OUT}}{1 - D_{MAX}} \sqrt{D_{MAX} \times \left(1 + \frac{1}{12}\left(\frac{I_{L,PP,VINMIN}}{I_{L,MAX}}\right)^2\right)}$$

$$I_{M1,RMS,MAX} = \frac{1.0}{1 - 0.791} \sqrt{0.791 \times \left(1 + \frac{1}{12}\left(\frac{1.55}{4.81}\right)^2\right)} = 4.30A$$

Iterative calculation of the total power losses and of the maximum junction temperature

Considering at maximum ambient temperature of 85°C, a thermal coefficient of the R_{DSON} of 0.006 K^{-1}, 20 ns of rise and fall times and a thermal resistance of 25 K/W, an iterative calculation leads to a junction temperature of 125°C and a R_{DSON} of 25.6 mΩ with a precision of 0.2 mΩ . Refer to *Section 8.3.2* for the calculation details.

This confirms that the MOSFET can operate in the worst case conditions, without exceeding the maximum rating of the junction temperature (175°C in general).

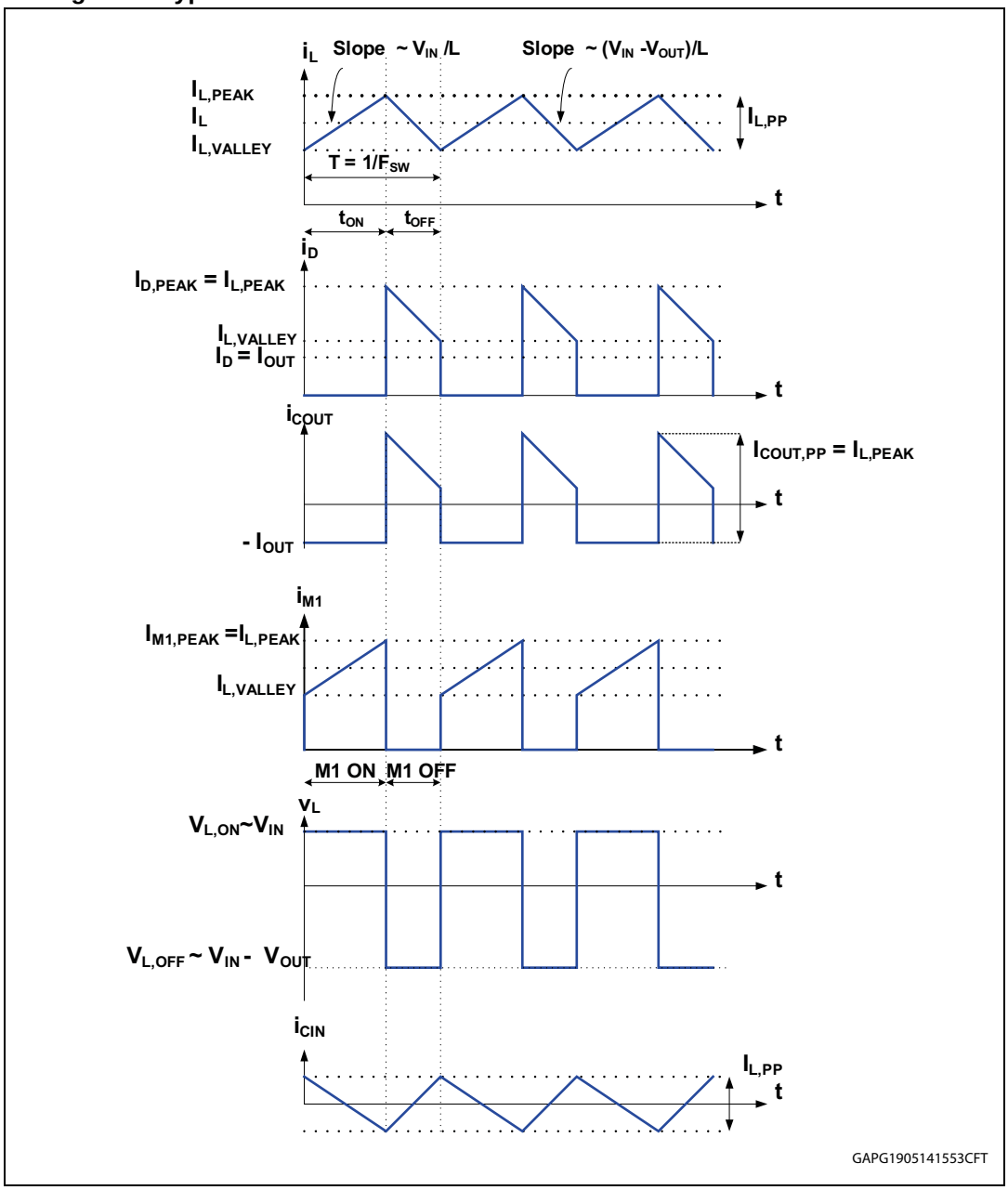
Table 2. Results of the iterative calculation of the worst case junction temperature

Iteration	T_{J1} [°C]	$R_{ON,M1@TJ1}$ [mΩ]	$P_{M1,COND@TJ1}$ [W]	$P_{M1@TJ1}$ [W]	T_{J2} [°C]	$R_{ON,M1@TJ1}$ [mΩ]	DiffRon [mΩ]
1	85	21.8	0.403	1.53	123	25.4	3.67
2	123	25.4	0.471	1.60	125	25.6	0.16

Table 3. Formula for currents and voltages of power components

Parameter	Value	Worst case V_{IN} (corresponding to the max parameter value)	Comment
D	$1-\frac{V_{IN}\times\eta}{V_{OUT}}$	$V_{IN,MIN}$	Duty cycle η represents the efficiency of the boost converter
I_L	$I_L=\frac{I_{OUT}}{1-D}$	$V_{IN,MIN}$	Inductor average current
$I_{L,PP}$	$\frac{V_{OUT}\times D\times(1-D)}{L\times F_{SW}\times\eta}$	$V_{IN,50}$	Inductor peak-to-peak current ripple
$I_{L,PEAK}$	$\frac{I_{OUT}}{1-D}+\frac{V_{OUT}\times D\times(1-D)}{2\times L\times F_{SW}\times\eta}$	$V_{IN,MIN}$ (in general)	Inductor peak current. The worst case V_{IN} must be verified case by case
$I_{L,RMS}$	$\sqrt{\frac{I_{OUT}^2}{(1-D)^2}+\frac{I_{L,PP}^2}{12}}$	$V_{IN,MIN}$ (in general)	The worst case V_{IN} must be verified case by case as the second term is not a monotonic increasing function
I_D	I_{OUT}	—	Diode average current
$I_{D,PEAK}$	$\frac{I_{OUT}}{1-D}+\frac{V_{OUT}\times D\times(1-D)}{2\times L\times F_{SW}\times\eta}$	$V_{IN,MIN}$ (in general)	Diode peak current. The worst case V_{IN} must be verified case by case
$V_{OUT,PP}$	$\frac{I_{OUT}\times D}{C_{OUT}\times F_{SW}}$	$V_{IN,MIN}$	Output peak-to-peak voltage ripple $ESRC_{OUT}$ is neglected
$I_{OUT,PP}$	$\frac{I_{OUT}\times D}{Nb_{LED}\times R_{DLED}\times C_{OUT}\times F_{SW}}$	$V_{IN,MIN}$	Boost output and LED peak-to-peak current ripple
$V_{CIN,PP}$	$\frac{I_{L,PP}}{8\times F_{SW}\times C_{IN}}$	$V_{IN,50}$	Input peak-to-peak current ripple $ESRC_{IN}$ is neglected
$I_{M1,RMS}$	$I_{M1,RMS}=\frac{I_{OUT}}{1-D}\sqrt{D\times\left(1+\frac{1}{12}\left(\frac{I_{L,PP}}{I_L}\right)^2\right)}$	$V_{IN,MIN}$ (in general)	MOSFET RMS current

Figure 4. Typical waveforms of a boost converter in continuous current mode



3 General considerations

3.1 Operating input voltage range

The input voltage range of automotive applications is usually wide, stretching from the cold cranking (below 5 V, depending on the car makers) or warm cranking (~7 V) to the jump start (~24 V). It is important to consider the component current, voltage and power dissipation over the whole operating range and not only for the minimum, typical or maximum input voltages. As we will see, some parameters reach their maximum value at a duty cycle of 50% and not at the minimum or maximum input voltages (V_{INMIN}, V_{INMAX}).

3.2 Output voltage and current

A slight variation of the voltage applied to a LED string results in a large variation of its forward current. As the light output and the color of the LEDs vary with the current, the best control strategy is a constant current generator to keep a constant brightness and the color.

The forward voltage of the LEDs (V_{FLED}) depends on the LED type, the process, the current, the temperature etc… The boost output voltage (V_{OUT}) is given by the formula:

$V_{OUT} = Nb_{LED} \times V_{FLED} + V_{RSENSE} + V_{RON,M2}$

Where Nb_{LED} is the number of LEDs in the string, V_{RSENSE} and $V_{RON,M2}$ are the voltage drop across R_{SENSE} and M2 (see *Figure 2*).

V_{SENSE} and $V_{RON,M2}$ can be neglected compared to $Nb_{LED} * V_{F,LED}$, therefore we will consider: $V_{OUT} - Nb_{LED} * V_{F,LED}$

The maximum V_{OUT} leads to the highest peak current in the inductor, in the switching transistor and in the diode. Therefore, a worst case calculation of those parameters must consider $V_{OUT,MAX}$.

The output current (I_{OUT}) is set by the choice of the sense resistor R_{SENSE} and a specific SPI control register of the L99LD01. To simplify, we consider only the case where this SPI register is set at its default value:

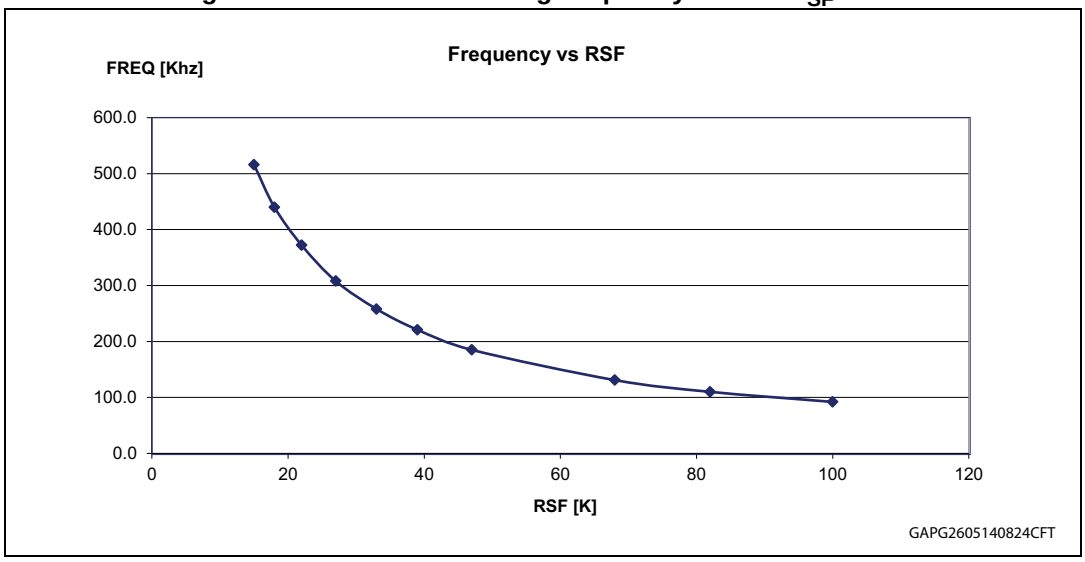
$I_{OUT} = 150\ mV\ (typ.) / R_{SENSE}$

3.3 Switching frequency

The switching frequency F_{SW} is a key parameter in the design of a DC-DC converter. Increasing the frequency allows in general the use of smaller capacitors and inductors, but as a drawback, it also leads to higher switching losses. Therefore the choice of the switching frequency is a tradeoff between costs, PCB area and efficiency.

The L99LD01 uses a constant frequency architecture, designed to operate from 100 kHz to 500 kHz. The switching frequency is set by the resistor R_{SF} as shown in *Figure 5*.

Figure 5. Converter switching frequency versus R_{SF} value



The L99LD01 works as a fixed frequency boost converter. It is possible to add a pseudo-random frequency modulation, which is controlled by a specific control register (see datasheet). The so-called spread spectrum technique distributes the electromagnetic disturbance over a wide frequency range, resulting in a reduction of the peak emission.

For clarity, in the rest of the document, we assume that this function is disabled.

4 Selection of the inductor

The selection of the inductor in a DC-DC converter has a direct influence on the performance and the selection of the other power devices. Therefore, care must be taken for its choice, as it dictates the cost and the overall performance of the system.

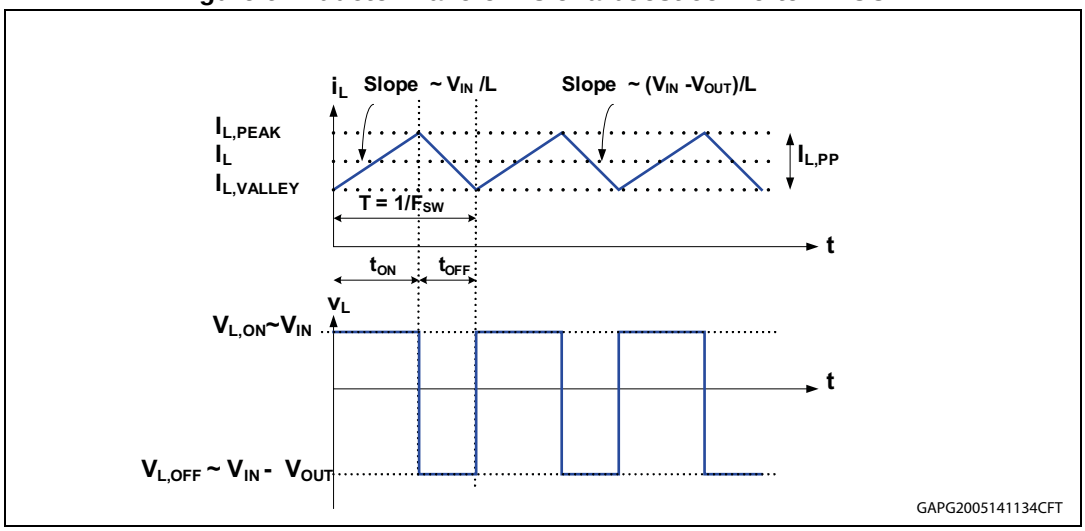
Like the switching frequency, the choice of the inductance value is a tradeoff between its size, its cost and the inductor current ripple.

A larger inductance results in:

- Smaller inductance current ripples
- Smaller input voltage and current ripples
- Smaller output voltage and current ripples
- Smaller current peaks in the converter switching MOSFET
- Smaller diode peak currents

However, a larger inductance value means higher cost, larger PCB surface and in general slower response time to transients.

Figure 6. Inductor waveforms of a boost converter in CCM



4.1 Converter Duty cycle

The expression of the ideal duty cycle of a boost converter, without power losses is:

Equation 1:

$$D_{ideal} = 1 - \frac{V_{IN}}{V_{OUT}}$$

The effective duty cycle taking into account the different sources of power losses is:

Equation 2:

$$D_{effective} = D = 1 - \frac{V_{IN} \times \eta}{V_{OUT}}$$

where η is the converter efficiency.

The duty cycle is a decreasing function of V_{IN}. In particular, the maximal duty cycle, noted D_{MAX}, is reached for the minimum input voltage, $V_{IN,MIN}$.

The duty cycle range must be compliant with two device parameters with the device minimum duty cycle (parameter T_{ON_MIN}, maximum specification: 14%) and maximum duty cycle (parameter Duty Max, minimum specification: 88%)

4.2 Inductor average current

In steady state, the average current of the output capacitor over one period must be equal to zero. Since the inductor delivers current to the load only during the converter's off-phase (see *Figure 7*), inductor current averaged during t_{OFF} is equal to the output current:

$$I_L \times \frac{t_{OFF}}{T} = I_L \times (1 - D) = I_{OUT}$$

Extracting I_L from this equation gives:

Equation 3:

$$I_L = \frac{I_{OUT}}{1 - D}$$

We can see that I_L is independent from the inductor value. Moreover, the worst case average inductor DC current is maximal for the maximal duty cycle. As a consequence, a worst case calculation of the inductor DC current must consider $V_{IN,MIN}$.

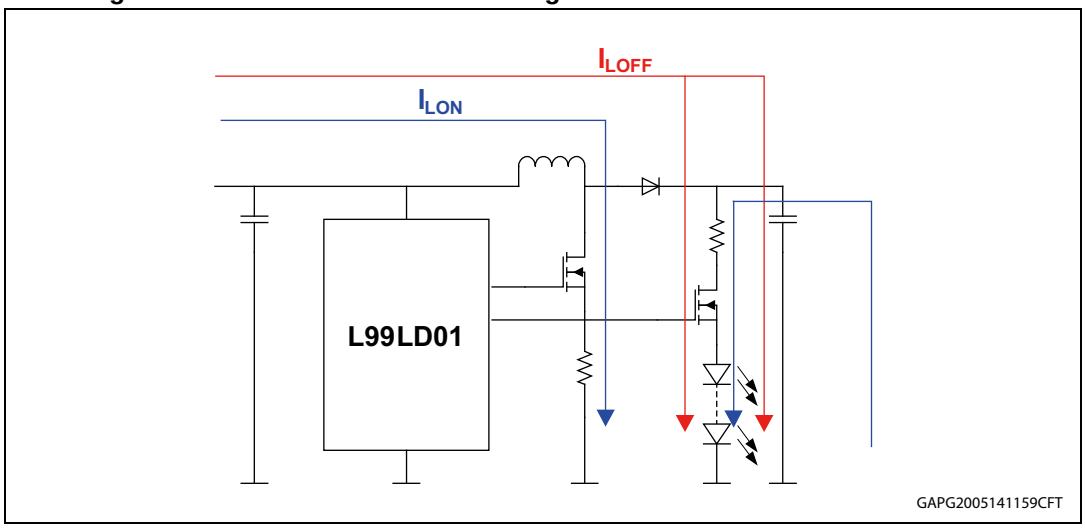
Equation 4:

$$I_{L,MAX} = \frac{I_{OUT}}{1 - D_{MAX}}$$

with

$$D_{MAX} = 1 - \frac{V_{IN,MIN} \times \eta}{V_{OUT,MAX}}$$

Figure 7. Inductor current flow during the converter on-time and off-time



4.3 Inductor peak-to-peak current ripple

$I_{L,PP}$ designates the inductor peak-to-peak current ripple:

$$I_{L,PP} = \frac{(V_{IN} - (R_{ONM1} + R_{SHUNT}) \times I_L)}{L \times F_{SW}} \times D$$

As a first approximation, we can neglect R_{ONM1} + R_{SHUNT} x I_L, compared to V_{IN}. The simplified expression of $I_{L,PP}$ is:

$$I_{L,PP} = \frac{V_{IN} \times D}{L \times F_{SW}}$$

Replacing V_{IN} by V_{OUT} x (1 - D)/η (see *Equation 2*) we obtain:

Equation 5:

$$I_{L,PP} = \frac{V_{OUT} \times D \times (1 - D)}{L \times F_{SW} \times \eta}$$

Extracting the inductance:

Equation 6:

$$L = \frac{V_{OUT} \times D \times (1 - D)}{I_{L,PP} \times F_{SW} \times \eta}$$

The derived function of $I_{L,PP}$ over D is:

Equation 7:

$$\frac{dI_{L,PP}}{dD} = \frac{V_{OUT} \times (1 - 2D)}{L \times F_{SW} \times \eta}$$

This shows that the maximum value of the inductor peak-to-peak current is reached for a duty cycle of D_{50} = 50%.

4.4 Inductor current ripple ratio

The inductor current ripple ratio r is defined as the ratio between the peak-to-peak current ripple and the average current.

Equation 8:

$$r = \frac{I_{L,PP}}{I_L}$$

Increasing the value of the inductance, we reduce the inductor current ripple and the output voltage ripple, as we will see in the section *Section 6.1.2*. In general, the max allowed inductor current ripple ratio is optimal for a value in the range of 0.3 to 0.5, from the standpoint of the cost / current ripple.

Indeed, reducing r to a value much lower than 0.3 leads a very large inductor size. Increasing r to a value which is much higher than 0.5, does not lead to a significant size reduction (see *Appendix C: Document management*). Therefore setting r to 0.4 or 0.5 is a good starting point.

$$I_{L,PP,MAX} = r \times I_{L,MAX} = r \times \frac{I_{OUT}}{1 - D_{MAX}}$$

Once the maximal inductor peak-to-peak current is fixed, we can estimate the minimum required inductance value, using *Equation 6* applied at $V_{IN,MIN}$ which corresponds to a duty cycle D_{MAX}.

Equation 9:

$$L_{MIN} = \frac{V_{OUT} \times D_{MAX} \times (1 - D_{MAX})}{I_{L,PP,MAX} \times F_{SW} \times \eta}$$

4.5 Inductor peak current

The inductor peak current must be calculated to make sure that in all cases, the inductor saturation current is not reached.

The inductor peak current, I_{LPEAK} is given by:

Equation 10:

$$I_{L,PEAK} = I_L + \frac{I_{L,PP}}{2}$$

Using the expression of $I_{L,PP}$ from *Equation 5* gives:

Equation 11:

$$I_{L,PEAK} = \frac{I_{OUT}}{1-D} + \frac{V_{OUT} \times D \times (1-D)}{2 \times L \times F_{SW} \times \eta}$$

As we will see, in most of the cases, the variation of $I_{L,PEAK}$ with the duty cycle is dictated by the term $I_{OUT}/(1 - D)$ if the $I_{L,PP}$ is lower than I_L. Therefore, in general $I_{L,PEAK\ MAX}$ is reached at the D_{MAX} (and V_{INMIN}). This property can be verified by calculating the derived function of $I_{L,PEAK}$:

Equation 12:

$$\frac{dI_{L,PEAK}}{dD} = \frac{I_{OUT}}{(1-D)^2} + \frac{V_{OUT} \times (1-2D)}{2L \times F_{SW} \times \eta}$$

In general the ratio between the inductor current peak-to-peak current and the inductor average current is kept below 1. Therefore:

Equation 13:

$$\frac{V_{OUT} \times D \times (1-D)}{L \times F_{SW} \times \eta} \leq \frac{I_{OUT}}{1-D}$$

Therefore:

$$\frac{V_{OUT} \times D}{L \times F_{SW} \times \eta} \leq \frac{I_{OUT}}{(1-D)^2}$$

$$\frac{dI_{L,PEAK}}{dD} = \frac{I_{OUT}}{(1-D)^2} - \frac{V_{OUT} \times D}{L \times F_{SW} \times \eta} + \frac{V_{OUT}}{2 \times L \times F_{SW} \times \eta} \geq \frac{V_{OUT}}{2 \times L \times F_{SW} \times \eta}$$

Under these conditions, $\frac{dI_{L,PEAK}}{dD} > 0$ therefore the peak current is a monotonically increasing function which reaches its maximum at D_{MAX} (and $V_{IN,MIN}$):

Equation 14:

$$I_{L,PEAK,MAX} = \frac{I_{OUT}}{1-D_{MAX}} + \frac{V_{OUT} \times D_{MAX} \times (1-D_{MAX})}{2 \times L \times F_{SW} \times \eta}$$

4.6 Input current limiter

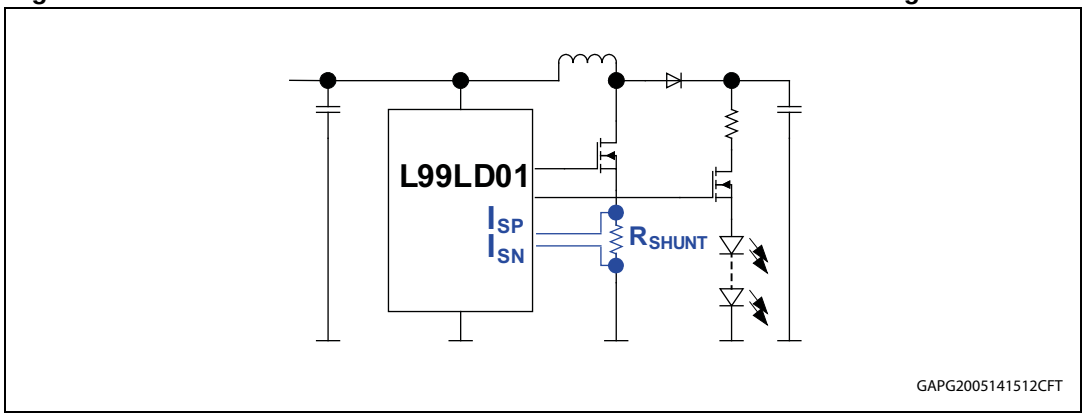
The L99LD01 offers a monitoring of the inductor current, which is sensed through the shunt resistor R_{SHUNT}. This feature ensures that the inductor current is always below the saturation current if R_{SHUNT} is correctly selected:

The threshold of the input current limitation is set by default to:

$$I_{INPUT,MAX} = \frac{2V}{G_{LAMP} \times R_{SHUNT}}$$

where G_{LAMP} (typ. value ~ 9.8, refer to datasheet of the L99LD01) is the gain of the linear amplifier.

Figure 8. A shunt resistor is used to monitor the inductor current during the on-state



The input current limiter can be configured by a specific SPI control register from

$$\frac{0.5V}{G_{LAMP} \times R_{SHUNT}}$$

to

$$\frac{3.5V}{G_{LAMP} \times R_{SHUNT}}$$

with a step of

$$\frac{3V}{31 \times G_{LAMP} \times R_{SHUNT}}$$

4.7 Limit of the continuous conduction mode

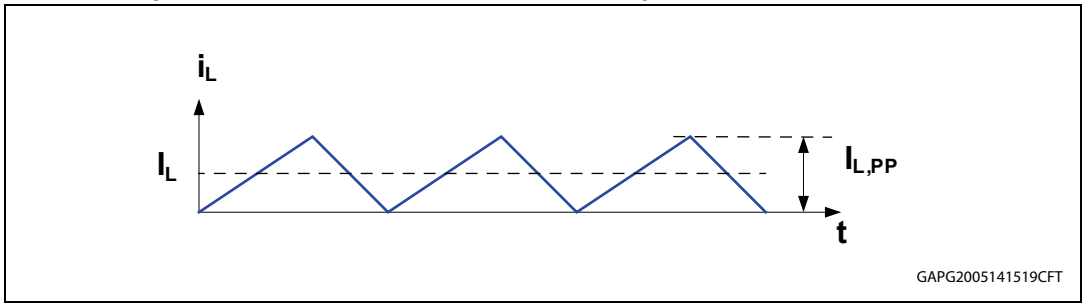
As stated in the introduction, all the considerations are valid only if the boost converter works in CCM, in other words, inductor current does not decay to zero.

An operation in discontinuous conduction mode (DCM) must be sometimes avoided, because of the increased electromagnetic emission. Indeed, when the output current or the duty cycle are low enough to allow the inductor current to decay to zero, we can observe a

ringing at the drain of the MOSFET with a typical frequency of some MHz. The effect is caused by the inductor in conjunction with the parasitic capacitances of the freewheeling diode and of the switching MOSFET, when the inductor current is close to zero.

We propose to calculate the minimum inductance value, which guarantees an operation in CCM over the whole input voltage range.

Figure 9. Inductor current at the boundary between CCM and DCM



The *Figure 9* shows the inductor waveform at the boundary between CCM and DCM. This condition can be expressed as: $I_{L,PP} = 2 \times I_L$.

Using *Equation 3* and *Equation 5*, we obtain:

$$\frac{V_{OUT} \times D \times (1 - D)}{L_{BOUNDARY} \times F_{SW} \times \eta} = \frac{2 \times I_{OUT}}{1 - D}$$

where $L_{BOUNDARY}$ is the inductance, for which the boost converter operates at the boundary between CCM and DCM

Extracting the inductance value yields:

Equation 15:

$$L_{BOUNDARY} = \frac{V_{OUT} \times D \times (1 - D)^2}{2 \times I_{OUT} \times F_{SW} \times \eta}$$

The expression of the derived function of $L_{BOUNDARY}$ over D is:

$$\frac{dL_{BOUNDARY}}{dD} = \frac{V_{OUT} \times (1 - D) \times (1 - 3D)}{2 \times I_{OUT} \times F_{SW} \times \eta}$$

This equation shows that, for a given output voltage and output current, L_{CRIT} reaches its maximum for a duty cycle of 33%:

- an operation in CCM at a duty cycle of 33% guarantees the CCM over the whole duty cycle range
- an inductance value higher than

$$L_{BOUNDARY} = \frac{2 \times V_{OUT}}{27 \times I_{OUT} \times F_{SW} \times \eta}$$

ensures the operation in CCM over the whole operating range

4.8 Inductor RMS current

The inductor RMS current ($I_{L,RMS}$) is needed to calculate the inductor copper loss (power dissipation caused by resistance of the inductor wires, noted DCR).

The waveform of the inductor current in CCM is a triangular signal with an average current of I_L and a peak-to-peak current $I_{L,PP}$ (see *Figure 4*).

$I_{L,RMS}$ is given by (see *Section 8.7* for the details of the calculations):

Equation 16:

$$I_{L,RMS} = \sqrt{I_L^2 + \frac{I_{L,PP}^2}{12}}$$

Similarly to $I_{L,PEAK}$, $I_{L,RMS}$ is also in general reached at D_{MAX} (and $V_{IN,MIN}$)

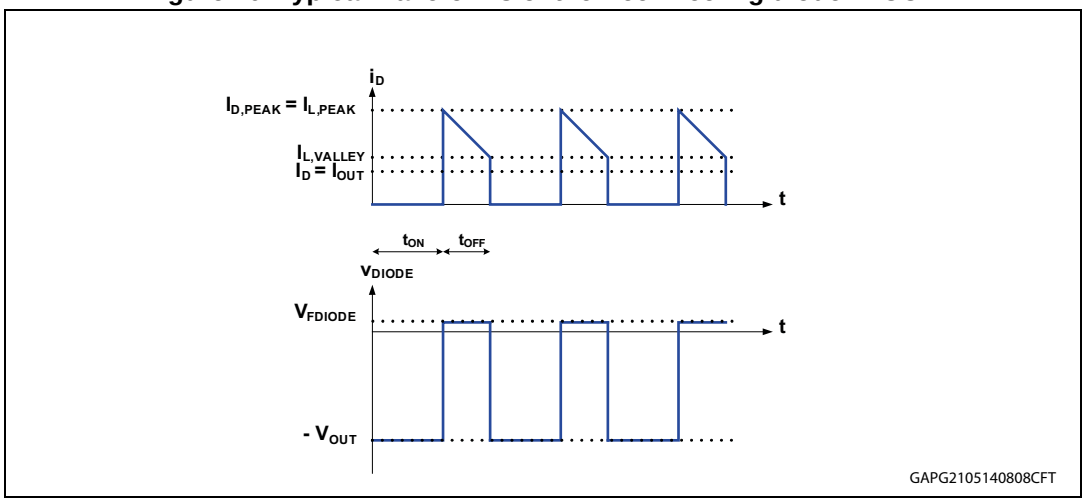
$$I_{L,RMS,MAX} = \sqrt{I_{L,MAX}^2 + \frac{I_{L,PP(V_{INMIN})}^2}{12}} = \sqrt{\frac{I_{OUT}^2}{(1-D_{MAX})^2} + \frac{I_{L,PP(V_{INMIN})}^2}{12}}$$

The maximum copper loss is $P_{COPPER,MAX} = DCR \times I^2_{L,RMS,MAX}$

5 Selection of the freewheeling diode

Schottky diodes are recommended to maximize the efficiency of the DC-DC converter thanks to their low forward voltage and their fast recovery time.

Figure 10. Typical waveforms of the freewheeling diode in CCM



Diode current capability

The freewheeling diode conducts only during the converter's off-phase (see i_D on *Figure 7* and *Figure 10*).Therefore, the average diode current over one switching period is equal to the output current. However, the diode peak current is equal to the inductor's peak current. The maximum rating of the Schottky diode must be chosen accordingly.

$I_{D,MAX} = I_{OUT}$

$I_{D,PEAK,MAX} = I_{L,PEAK,MAX}$ (see *Equation 14*)

Diode power dissipation and temperature

The power dissipation in the freewheeling diode is $P_D = V_{FDIODE} \times I_{OUT}$

The cooling of the diode must guarantee that its max. junction temperature is not exceeded even at the maximum ambient temperature:

$T_{J,DIODE} = T_{AMB} + R_{THJ\text{-}AMB,DIODE} \times P_D$

$T_{J,DIODE,MAX} = T_{AMB,MAX} + R_{THJ\text{-}AMB,DIODE} \times P_D$

Diode breakdown voltage

The diode is reverse biased during the on-phase of M_1. The maximum reverse voltage in normal operation must be at least higher than the max. LED voltage (neglecting the voltage drop across R_{SENSE} and M_2). However, the diode can see a reverse voltage, which is even higher in case of open load.

Indeed, as the output current goes to zero, the L99LD01 increases the duty cycle and the output voltage until an over-voltage condition on the output is detected.

The freewheeling diode must also withstand a reverse voltage up to the output over-voltage threshold ($V_{OUT,OVTH}$).

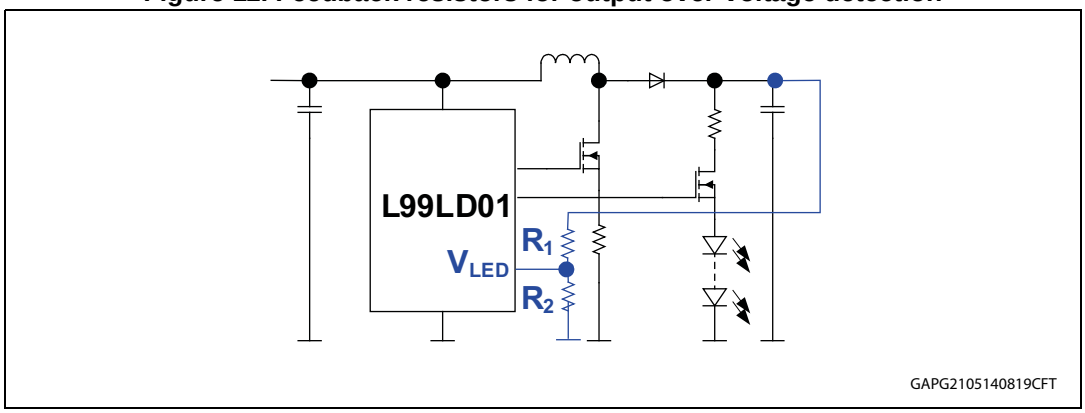
$V_{OUT,OVTH}$ must be set to a value which is higher than $V_{LED,MAX}$.

The overvoltage threshold is determined by the resistors R_1 and R_2 (see *Figure 11*):

$$V_{OUT,OVTH} = OV_TH1 \times \left(1 + \frac{R_1}{R_2}\right)$$

where OV_TH1 is typically 3.5 V (refer to datasheet of the L99LD01).

Figure 11. Feedback resistors for output over-voltage detection



The minimum breakdown, with a margin of 20% gives:

Equation 17:

$V_{BRMIN} = 1.2 \times V_{OUT,OVTH}$

6 Selection of the output capacitor

The output capacitor determines the output voltage and current ripples.

For a current source, the choice of the output capacitor begins with the specification of the maximum output current ripple $I_{OUT,PP,MAX}$.

The capacitor voltage capability must be higher than the maximum output voltage. Note that some margin must be taken, as this parameter has a non-negligible tolerance and varies with the temperature and the applied DC voltage (if a MLCC capacitor is used).

6.1 Output voltage ripple

The main causes of the output voltage ripples are:

- The charge, respectively the discharge, of the ideal capacitor without equivalent series resistor (ESR_{COUT}) during t_{ON} and t_{OFF}. The ideal capacitance is called the bulk capacitance in the rest of the document.
- The voltage drop caused by the output capacitor's current ripple across ESR_{COUT}.

6.1.1 Contribution of the charge and discharge of the bulk capacitance

During the on-phase, the output current is exclusively delivered by the output capacitor C_{OUT}.

During t_{ON}: $i_{COUT} = -i_{OUT} \sim -I_{OUT}$ (see*Figure 12*)

As a first approximation, we assume that the output current ripples are negligible compared to the average value. This assumption is justified by the fact that the selection of the output capacitor should limit the conducted emission at the output in order to fulfill stringent specifications on the electromagnetic emissions.

$$i_{COUT} = C_{OUT} \times \frac{dV_{COUT}}{dt} \sim -I_{COUT}$$

The integration of the output voltage over the on-phase gives:

$$\left|\Delta V_{COUT}\right| = \frac{I_{OUT}}{C_{OUT}} \times t_{ON} = \frac{I_{OUT} \times D}{C_{OUT} \times F_{SW}}$$

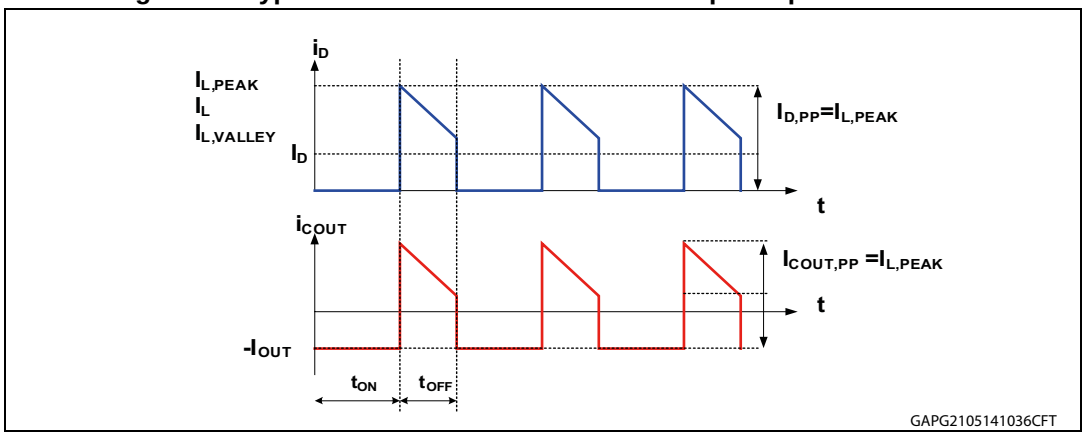
6.1.2 Contribution of the output capacitor ESR to the output voltage ripple

Considering that:

- During t_{ON}, $i_{COUT} = -i_{OUT} \sim -I_{OUT}$ and the diode is reverse biased: $i_D = 0$
- During t_{OFF}, the diode is conducting and charges C_{OUT} and delivers current to the LED strings: $i_{COUT} \sim i_D - I_{OUT}$

We can conclude that at anytime, $i_{COUT} \sim i_D - I_{OUT}$ and $I_{COUT,PP} = I_{D,PP} = I_{L,PEAK}$ (see *Figure 12*).

Figure 12. Typical current waveforms of the output capacitor current



Equation 18:

$$I_{COUT,PP} = I_{D,PP} = I_{L,PEAK} = \frac{I_{OUT}}{1-D} + \frac{V_{OUT} \times D \times (1-D)}{2 \times L \times F_{SW} \times \eta}$$

The contribution ESR_{COUT} to the output voltage ripple is:

Equation 19:

$$\Delta V_{COUT,ESR} = ESR_{COUT} \times I_{L,PEAK}$$

6.1.3 Total output voltage ripple

Summing both contributions, we obtain:

$$V_{OUT,PP} = \Delta V_{COUT} = \Delta V_{COUT,ESR} = \frac{I_{OUT} \times D}{C_{OUT} \times F_{SW}} + ESR_{COUT} \times I_{L,PEAK}$$

The substitution of $I_{L,PP}$ by its expression from *Equation 11* yields:

Equation 20:

$$V_{OUT,PP} = \frac{I_{OUT} \times D}{C_{OUT} \times F_{SW}} + ESR_{COUT} \times \left(\frac{I_{OUT}}{1-D} + \frac{V_{OUT} \times D \times (1-D)}{2 \times L \times F_{SW} \times \eta} \right)$$

6.2 Output current ripple

The output current ripple is the ratio between the total output voltage ripple and the dynamic resistance of the LED string (note that R_{SENSE} and R_{ONM2} can be neglected compared to the dynamic resistance of the LED string).

$R_{D,OUT} \sim Nb_{LED} \times R_{D,LED}$ where Nb_{LED} is the number of LEDs in the string and $R_{D,LED}$ is the dynamic resistance one LED at the considered output current.

The resulting output current ripple is:

Equation 21:

$$I_{OUT,PP} = \frac{\Delta V_{COUT} + \Delta V_{COUT,ESR}}{R_{DOUT}} =$$

$$= \frac{1}{Nb_{LED} \times R_{DLED}} \times \left(\frac{I_{OUT} \times D}{C_{OUT} \times F_{SW}} + ESR_{COUT} \times I_{L,PEAK} \right)$$

6.3 Calculation of the output capacitor

Ceramic capacitors are recommended. As they have a low ESR, we can first select the output capacitance, neglecting the contribution of ESR_{COUT}. The *Equation 21* becomes:

$$I_{OUT,PP,MAX} \sim \frac{1}{Nb_{LED} \times R_{DLED}} \times \frac{I_{OUT} \times D_{MAX}}{C_{OUT,MIN} \times F_{SW}}$$

Extracting C_{OUT} yields:

Equation 22:

$$C_{OUT,MIN} = \frac{1}{Nb_{LED} \times R_{DLED}} \times \frac{I_{OUT} \times D_{MAX}}{I_{OUT,PP,MAX} \times F_{SW}}$$

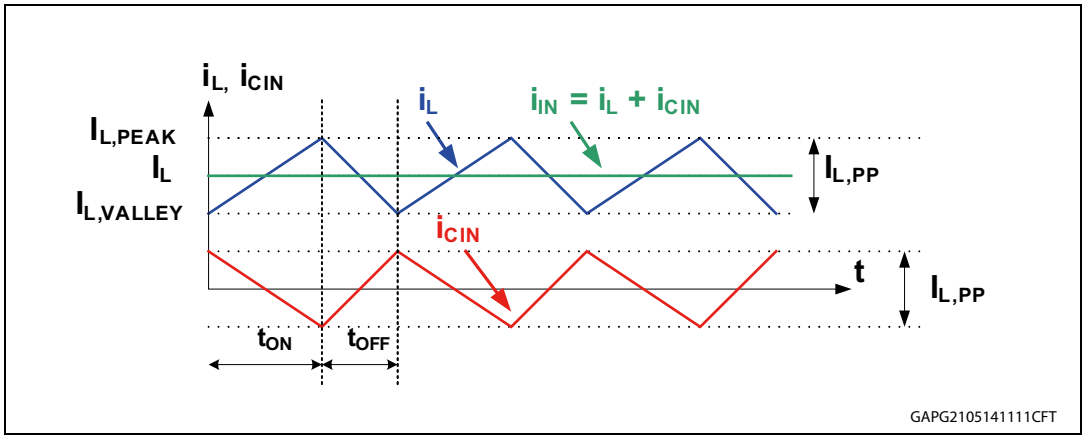
Once the output capacitor is chosen based on *Equation 22*, its ESR is known, and the contribution of the ESR to the output current ripple can be calculated, so that the assumption can be confirmed.

7 Selection of the input capacitor

An input capacitor is required to provide the AC current to the inductor and to reduce the input voltage ripple. Therefore, its choice has an impact on the electromagnetic emission at the input.

Note: this section considers only the input voltage ripples caused by the operation of the boost converter in steady state. An additional (and bigger) capacitor might be necessary to buffer the input voltage, for example in case of line transients.

Figure 13. Current waveforms of the inductor and of the input capacitor for a boost converter in CCM



7.1 Input voltage ripples

Like the output voltage ripples, the input voltage ripples are due to the charge and the discharge of the (ideal) input bulk capacitance and of the current ripple across the ESR.

7.1.1 Contribution bulk capacitance to the input voltage ripple

The peak-to-peak voltage ripple in the input capacitor (noted $V_{CIN,PP}$) corresponds to the voltage increase during the charge of the capacitor ($i_{CIN} > 0$).

We have:

$$\Delta q_{CIN} = \int_{\frac{t_{ON}}{2}}^{\frac{t_{ON}+t_{OFF}}{2}} i_{CIN}dt = C_{IN} \times V_{CIN,PP}$$

$\int_{\frac{t_{ON}}{2}}^{\frac{t_{ON}+t_{OFF}}{2}} i_{CIN}dt$ is represented by the blue area (see *Figure 14*).

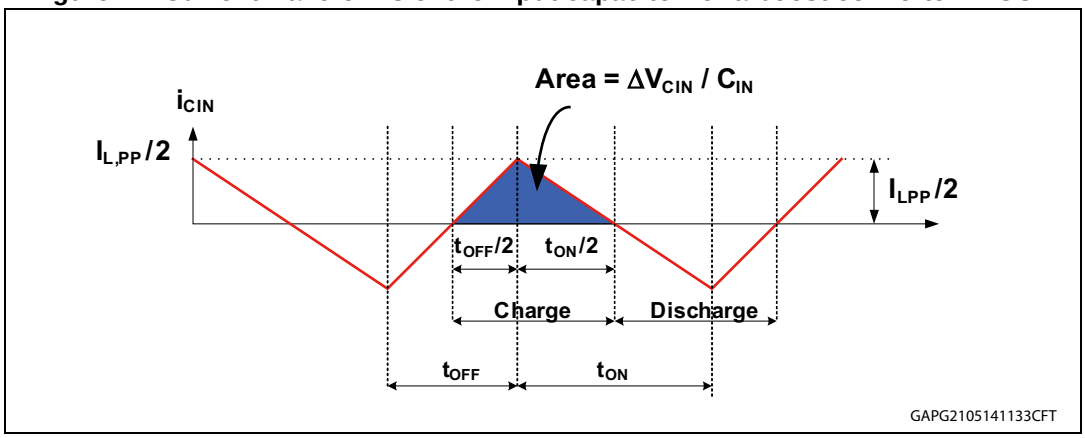
$$\int_{\frac{t_{ON}}{2}}^{\frac{t_{ON}+t_{OFF}}{2}} i_{CIN}dt = \frac{1}{2} \times \frac{I_{L,PP}}{2} \times \frac{t_{ON}+t_{OFF}}{2}$$

The substitution $t_{ON} + t_{OFF}$ by $1/F_{SW}$ gives:

Equation 23:

$$V_{CIN,PP} = \frac{I_{L,PP}}{8 \times F_{SW} \times C_{IN}}$$

Figure 14. Current waveforms of the input capacitor for a boost converter in CCM



Similarly to $I_{L,PP}$, $V_{CIN,PP}$ reaches its maximum when the duty cycle is 50%.

7.1.2 Contribution of the capacitor ESR to the input voltage ripple

As the peak-to-peak current ripple of the input capacitor is equal to $I_{L,PP}$, the input voltage ripple caused by the ESR_{CIN} is given by:

$$V_{ESR,CIN,PP} = ESR_{CIN} \times I_{L,PP}$$

Here again, the worst case corresponds to a duty cycle of 50%.

Therefore:

$$V_{ESR,CIN,PP,MAX} = ESR_{CIN} \times I_{L,PP,MAX}$$

7.2 Maximum input voltage ripple

The max input voltage ripple is the sum of $\Delta V_{CIN,MAX}$ and $\Delta V_{ESR,CIN,MAX}$:

Equation 24:

$$V_{IN,PP,MAX} = \frac{I_{L,PP,MAX}}{8 \times F_{SW} \times C_{IN}} + ESR_{CIN} \times I_{L,PP,MAX}$$

8 Selection of the switching MOSFET

The MOSFET M1 is the main switching element of the boost converter. The most important parameters for its selection are:

- the breakdown voltage
- the peak and the RMS currents
- the R_{DSON}
- the thermal resistance $R_{TH\text{-}J\text{-}AMB}$
- the turn-on and the turn-off time

Note: The gate driver of the L99LD01 controls M1 with a typ. voltage, V_{G1}, of ~ 10V, provided that the supply voltage is high enough (10 V + Dropout of the V_{CC2} internal regulator ~ 10.2 V). If V_{IN} is below ~10.2 V, V_{CC2} will be ~V_{IN} – 0.2 V and so does V_{G1}. Therefore, a logic level MOSFET is required to keep the converter's performance at V_{IN} below ~ 10.2 V.

8.1 Breakdown voltage

During t_{OFF} the drain-source voltage of the MOSFET M1 is equal to $V_{OUT} + V_{F,DIODE}$. Some margin must be added in case of ringing at the switching node.

8.2 MOSFET peak current

During t_{ON}, the MOSFET M1 is turned on and the inductor current flows into M1 (see *Figure 4* and *Figure 7*). Therefore, the MOSFET maximum peak current is equal to the inductor maximum peak current:

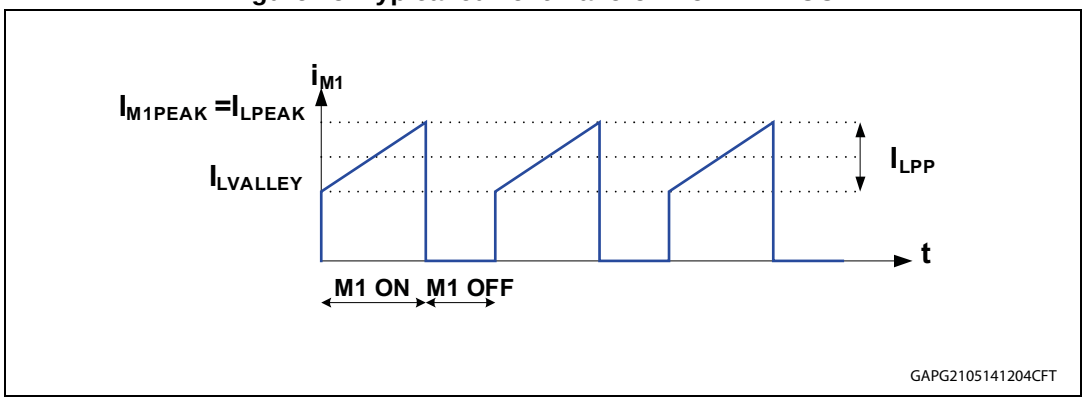
$$I_{M1,\,PEAK,\,MAX} = I_{L,\,PEAK,\,MAX} = \frac{I_{OUT}}{1 - D_{MAX}} + \frac{V_{OUT} \times D_{MAX} \times (1 - D_{MAX})}{2 \times L \times F_{SW} \times \eta}$$

8.3 MOSFET power dissipation

The MOSFET's power dissipation mainly comes from the the conduction losses and the switching losses.

8.3.1 Conduction losses

Figure 15. Typical current waveform of M1 in CCM



The conduction losses of M1 are given by $P_{M1,COND} = R_{ON,M1} \times I^2_{M1,RMS}$, where $R_{ON,M1}$ is the Rdson of the MOSFET M1 and $I_{M1,RMS}$ is the root means square current (see *Section 8.5* for the calculation details):

Equation 25:

$$I_{M1,RMS} = \frac{I_{OUT}}{1-D}\sqrt{D \times \left(1 + \frac{1}{12}\left(\frac{I_{L,PP}}{I_L}\right)^2\right)}$$

$\frac{1}{12}\left(\frac{I_{L,PP}}{I_L}\right)^2$ is in general negligible compared to 1 and the MOSFET RMS current is dominated by the term $\frac{\sqrt{D}}{1-D}$, which increases as D increases. The RMS current reaches its maximum value at $V_{IN,MIN}$:

$$I_{M1,RMS,MAX} = \frac{I_{OUT}}{1-D_{MAX}}\sqrt{D_{MAX} \times \left(1 + \frac{I_{L,PP(VINMIN)}^2}{12 \times \left(\frac{I_{OUT}}{1-D_{MAX}}\right)^2}\right)}$$

Finally, the conduction losses are given by:

$$P_{M1,COND} = R_{ON,M1} \times \left(\frac{I_{OUT}}{1-D_{MAX}}\right)^2 \times D_{MAX} \times \left(1 + \frac{I_{L,PP(VINMIN)}^2}{12 \times \left(\frac{I_{OUT}}{1-D_{MAX}}\right)^2}\right)$$

$$P_{M1,\,COND} \sim R_{ON,\,M1} \times \left(\frac{I_{OUT}}{1 - D_{MAX}} \right)^2 \times D_{MAX}$$

The Rdson is temperature dependant:

$R_{ON,M1} = R_{ON,M1@T25°C} \times (1 + \alpha\,(T_j - 25))$

Where:

- $R_{ON,M1@T25°C}$ is the R_{DSON} at 25°C
- T_j is the junction temperature of M1
- α is the temperature coefficient, which is in general in the range of $6 \cdot 10^{-3}\ K^{-1}$

In return the junction temperature depends on the power losses. Therefore, an iterative calculation is necessary for an accurate estimation of the conduction loss. *Figure 18* displays the flowchart for the estimation of the MOSFET power losses and R_{DSON}.

8.3.1 Switching losses of M1

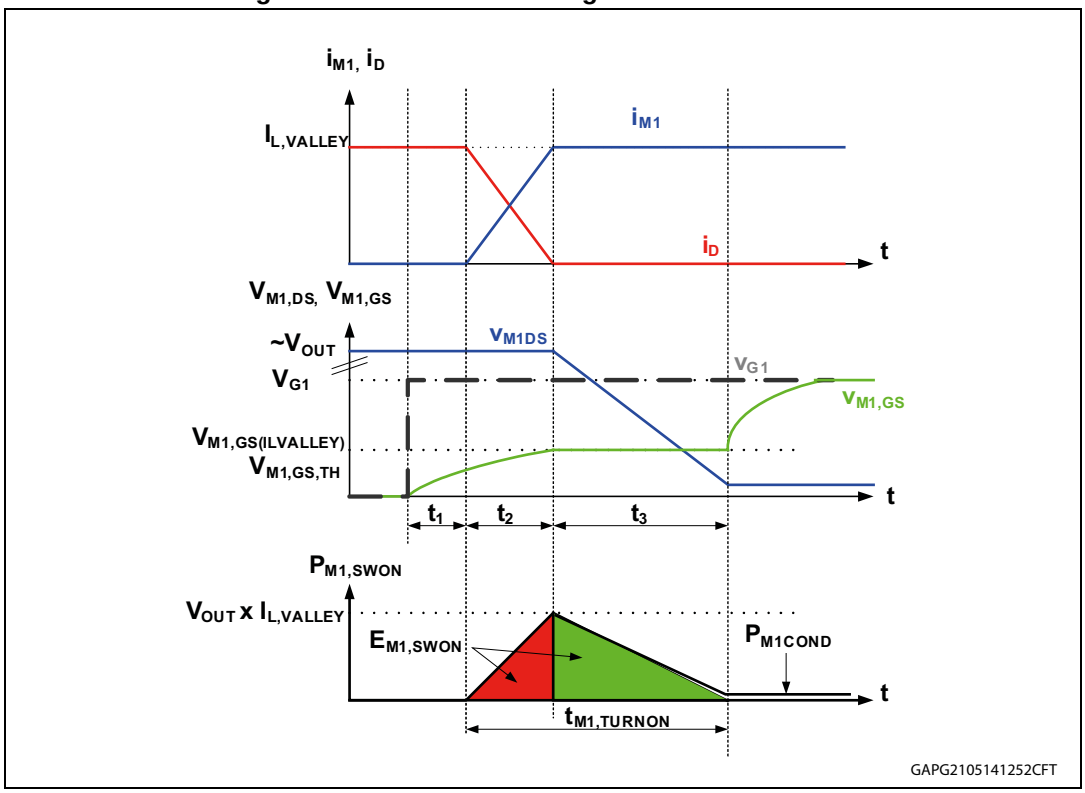
Some care must be taken for the calculation of the switching losses. Often, the turn-on and the turn-off times which are specified in datasheets of the MOSFET are applicable for a resistive load under very specific conditions (current and drain-source voltage).They are not valid for the switching of inductive loads.

Switching losses in the M1 MOSFET occur when the drain-source voltage is high, while the M1 current is not negligible. We can split the switching losses between the turn-on and the turn-off transitions: $P_{M1,SWON}$ and $P_{M1,SWOFF}$.

Switch-on phase

During t_1, the gate-source voltage of M1 ($v_{M1,GS}$) ramps up to the gate-source threshold voltage, $V_{M1,GS,TH}$. During this phase, there is no change in the drain-source voltage of M1 ($V_{M1,DS}$) and M1 is not yet conducting. The inductor's current ($I_{L,VALLEY}$) still flows through the diode and there is no switching loss in M1 (see *Figure 16*).

Figure 16. Waveforms during the switch-on of M1



During t_2, $v_{M1,GS}$ is now above $V_{M1,GS,TH}$,and M1 starts to conduct, and overtakes partially the diode current. As the diode still conducts, $V_{M1,DS}$ remains unchanged at $V_{OUT} + V_{FDIODE} \sim V_{OUT}$. This phase ends up, when the $v_{M1,GS}$ reaches the value which allows M1 to drive the whole inductor current (called $V_{M1,GS,ILVALLEY}$).

During t_2, the switching energy is estimated by:

$$\frac{V_{OUT} \times I_{L,VALLEY} \times t_2}{2}$$

which represents the area of the red triangle on *Figure 16*.

At the beginning of t_3, The diode stops conducting, allowing $V_{M1,DS}$ to drop. The gate-drain capacitor of M1 (C_{GD}) is being discharged. Apparently, the value of C_{GD} increases, which explains constant value of $V_{M1,GS}$ despite the current injected in the gate by the gate driver G1.

The switching energy during on transition is represented by the green triangle on *Figure 16* the:

$$\frac{V_{OUT} \times I_{L,VALLEY} \times t_3}{2}$$

After t_3, the M1 is fully turned on. The power dissipation during this phase is considered as conduction losses. C_{GS} is charged by G1 and $V_{M1,GS}$ further increases to reach the gate driver's voltage.

The switch-on losses are the total switching energy multiplied by the switching frequency:

Equation 26:

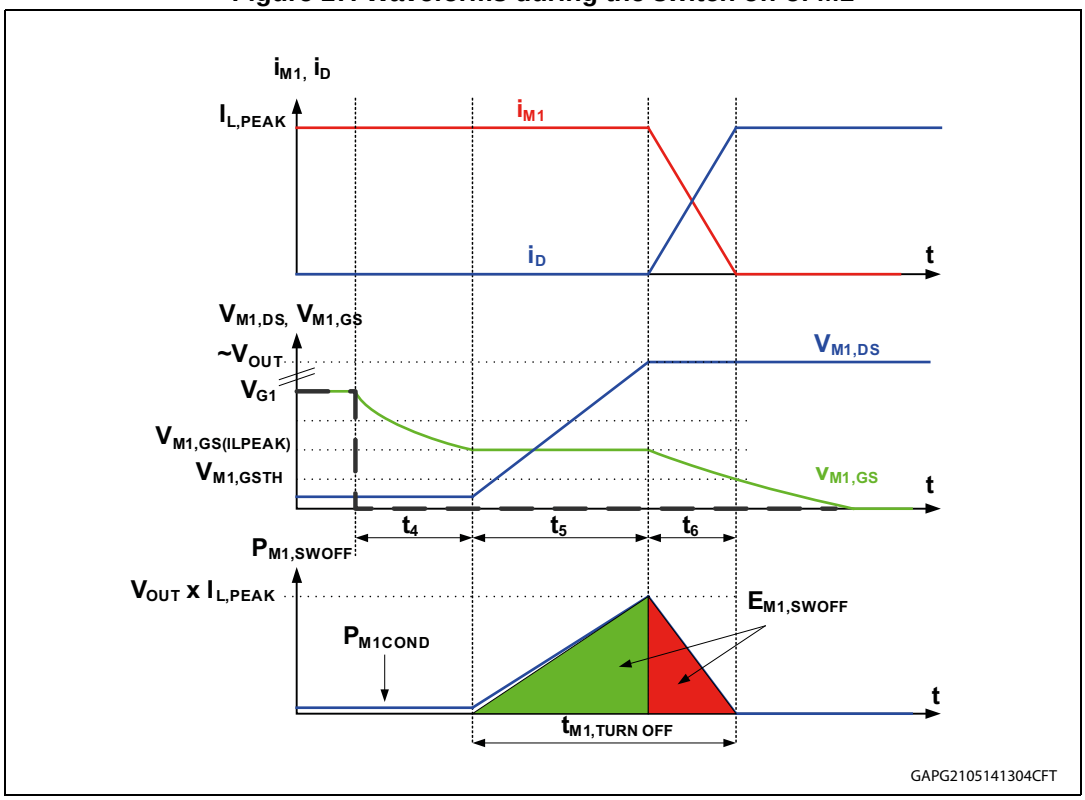
$$P_{M1,SWON} = \frac{V_{OUT} \times I_{L,\,VALLEY} \times t_{M1,SWON} \times F_{SW}}{2}$$

Where $t_{M1,SWON} = t_2 + t_3$.

Switch-off phase

Figure 17 shows the waveforms during the switch-off of M1, which are relevant for the switching losses. Prior to the phase t_4, M1 is fully on and conducts the current $I_{L,PEAK}$. The diode is reverse biased.

Figure 17. Waveforms during the switch off of M1



At the beginning of the t_4 phase, the voltage of the gate drive G1 drops and $v_{M1,GS}$ decreases down to the sustaining voltage $V_{M1,GS(ILPEAK)}$ (minimum voltage required to sustain the drain current $I_{L,PEAK}$). There is no switching loss during this phase.

During t_5, $v_{M1,DS}$ increases again up to ~ V_{OUT}. As a consequence C_{GD} is charged and $v_{M1,GS}$ reaches again the plateau. The switching energy represented by the green area on *Figure 17* is estimated by the formula:

$$\frac{V_{OUT} \times I_{L,\,PEAK} \times t_5}{2}$$

At the beginning of the t_6 phase, $v_{M1,DS}$ is high enough to allow the freewheeling diode to conduct. As the drain current of M1 decreases, so does $v_{M1,GS}$.

The related switching energy is estimated by:

$$\frac{V_{OUT} \times I_{L,PEAK} \times t_6}{2}$$

At the end of the t_6 phase, $v_{M1,GS}$ reaches $v_{M1,GS,TH}$. Therefore, the drain current of M1 is zero and the diode conducts the inductor current $I_{L,PEAK}$. The switching loss in this interval is zero.

To sum up, the switching losses during the turn-off of M1 are equal to the total switching energy during this phase, multiplied by the switching frequency.

Equation 27:

$$P_{M1,SWOFF} = \frac{V_{OUT} \times I_{L,PEAK} \times t_{M1,SWOFF} \times F_{SW}}{2}$$

Where $t_{M1,SWOFF} = t_5 + t_6$.

Total power switching losses of M1

From *Equation 26* and *Equation 27*, we can estimate the total switching losses of M1:

Equation 28:

$$P_{M1,SW} = P_{M1,SWON} + P_{M1,SWOFF} =$$
$$= \frac{V_{OUT} \times F_{SW}}{2}(I_{L,VALLEY} \times t_{M1,SWON} + I_{L,PEAK} \times t_{M1,SWOFF})$$

If the switch-on and switch-off times do not significantly differ, we can approximate the term $I_{L,VALLEY} \times t_{M1,SWON} + I_{L,PEAK} \times t_{M1,SWOFF}$ by $I_L (t_{M1,SWON} + t_{M1,SWOFF})$:

Equation 29:

$$P_{M1,SW} = \frac{V_{OUT} \times I_L \times F_{SW}}{2}(t_{M1,SWON} + t_{M1,SWOFF})$$

We see that the switching losses reach the maximum value at $V_{IN,MIN}$, which corresponds to $I_{L,MAX}$.

8.3.2 Iterative calculation of the junction temperature of M1

The temperature dependence of the Rdson requires an iterative calculation. We propose here a way how to proceed.

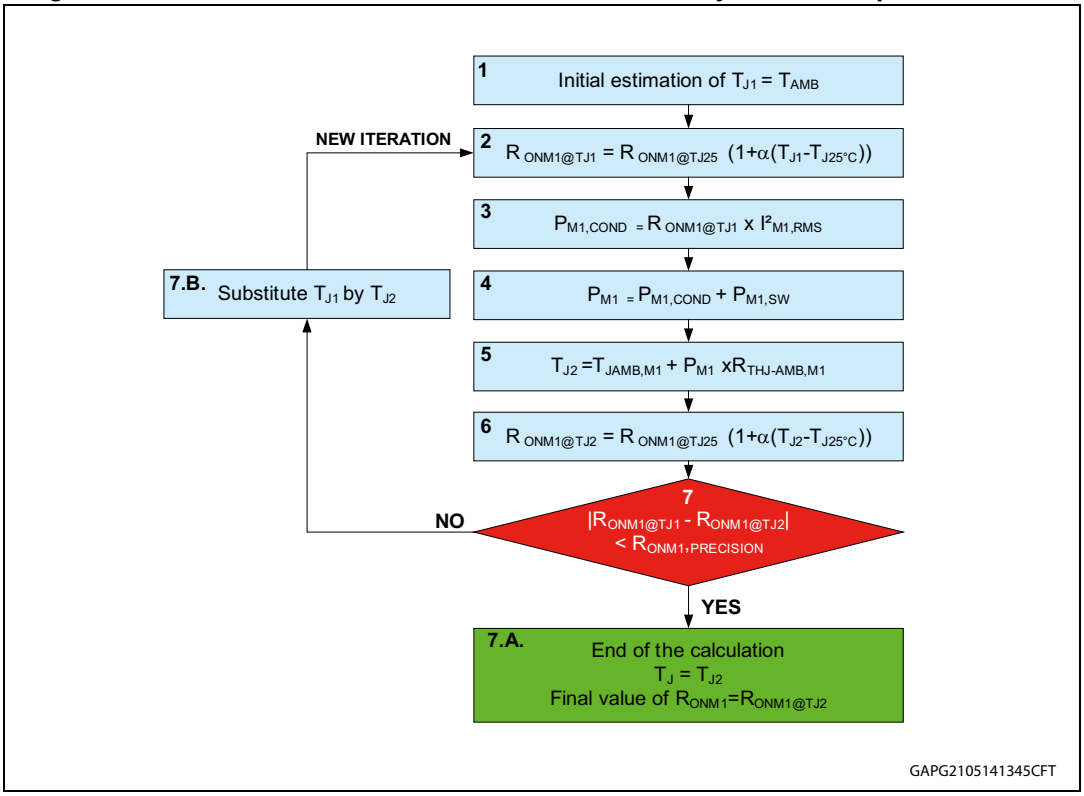
To have a first approximation of the junction temperature of M1, we consider that the switching losses are temperature independent.

The *Figure 18* sums up the proposed procedure:

1. Initial estimation of T_J (noted T_{J1}). For example, we can use T_{AMB} as a starting point.
2. Calculation of the corresponding Rdson, using the Rdson at 25°C and the thermal coefficient α
3. Calculation of the conduction losses in M1
4. Calculation of the total power losses in M1
5. Calculation of the resulting junction temperature (noted T_{J2})
6. Calculation of the R_{DSON} corresponding to T_{J2}
7. The calculation is finished if the difference between R_{DSON} at T_{J1} and T_{J2} is smaller than the required precision. The final R_{DSON} and junction temperature are respectively $R_{ONM1@TJ2}$ and T_{J2}.

If the difference between Rdson at T_{J1} and T_{J2} is higher than the required precision we substitute T_{J1} by T_{J2} and we restart a new iteration.

Figure 18. Flowchart for the iterative calculation of the junction temperature of M1



Appendix A Glossary

In this document, currents and voltages with lower case character refer to the instantaneous parameter whereas the higher case characters refer to the DC or average values. For example i_L represents the instantaneous inductor current and I_L designates the DC inductor current.

The subscripts $_{MIN}$ and $_{MAX}$ refer to the minimum and to the maximum values. The subscripts $_{PEAK}$ and $_{PP}$ refer to the peak and to the peak-to-peak value. The subscript $_{RMS}$ refers to the RMS value.

Table 4. Notations and abbreviations

Notation	Description
$I_{IN,}$ V_{IN}	Input current and voltage
I_{OUT},V_{OUT}	Output current and voltage
V_{FLED}	Single LED forward voltage
F_{SW}	Converter's switching frequency
T	Converter's period = 1/ F_{SW}
I_L, V_L	Inductor voltage and current
L	Inductor and inductance
$L_{BOUNDARY}$	Inductance operating at the boundary between CCM and DCM
$I_{L,PEAK}$	Inductor peak current
$I_{L,VALLEY}$	Inductor valley current
$I_{L,PP}$	Inductor peak-to-peak current
$I_{L,PP@VINMIN}$	$I_{L,PP}$ at $V_{IN,MIN}$
$t_{ON,}$ t_{OFF}	Converter's on-time, off-time
D	Converter's duty cycle = t_{ON}/T
V_{FDIODE}	Forward voltage of freewheeling diode
$R_{THJ-AMB,D}$	Junction-to-ambient thermal resistance of the freewheeling diode
$R_{THJ-AMB,M1}$	Junction-to-ambient thermal resistance M1
M1	Boost switching MOSFET
M2	Dimming MOSFET
I_{M1} ,V_{M1}	M1 drain-source current and voltage
$V_{M1,GS}$	M1 gate-source voltage
$v_{M1,GS,TH}$	Threshold voltage of $v_{M1,GS}$
$v_{M1,DS}$	Drain-Source voltage of M1
C_{GS}	M1 gate-source capacitor
C_{GD}	M1 gate-drain capacitor

Table 4. Notations and abbreviations (continued)

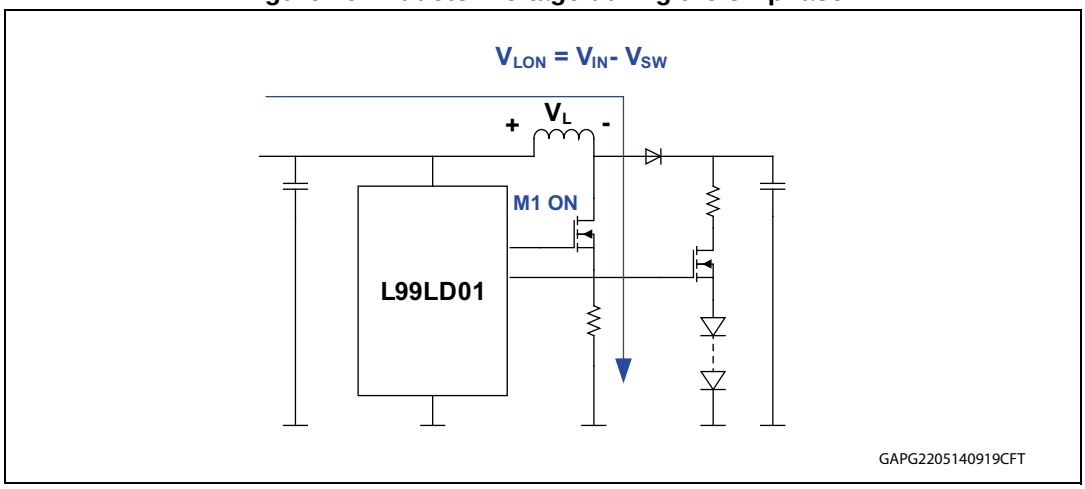
Notation	Description
C_{ISS}	M1 input capacitance
I_{COUT}, V_{COUT}	Output capacitor current and voltage
ESR_{COUT}	ESR output capacitor
$V_{COUT,ESR}$	Voltage across ESR_{COUT}
I_{CIN}, V_{CIN}	Input capacitor current and voltage
ESR_{CIN}	ESR input capacitor
$V_{CIN,ESR}$	Voltage across ESR_{CIN}
R_{ONM1}	Rdson of M1
R_{ONM2}	Rdson of M2
R_{SHUNT}	Shunt resistor
R_{SENSE}	Sense resistor
r	Inductor current ripple ratio
r_{MAX}	Max. allowed inductor current ratio
I_{OUT}	Output current (= LED current)
DCR	Inductor DC resistance
D_{50}	50% duty cycle
V_{G1}	Gate driver voltage
a	Thermal coefficient of R_{ONM1}
P_{DIODE}	Power dissipation of the diode
$P_{M1,SWON}$	Switching losses of M1 during the switch-on transition
$P_{M1,SWOFF}$	Switching losses of M1 during the switch-off transition
T_{AMB}	Ambient temperature
$P_{M1,COND}$	Conduction losses of M1
P_{M1}	Total power losses of M1
h	Converter’s efficiency
CCM	Continuous conduction mode
DCM	Discontinuous conduction mode
MLCC	Multilayer ceramic capacitor

Appendix B Calculation details

8.4 Calculation of the duty cycle in CCM

During t_{ON}, the inductor is magnetized by the voltage $V_{L,ON} = V_{IN} – V_{SW}$

Figure 19. Inductor voltage during the on-phase

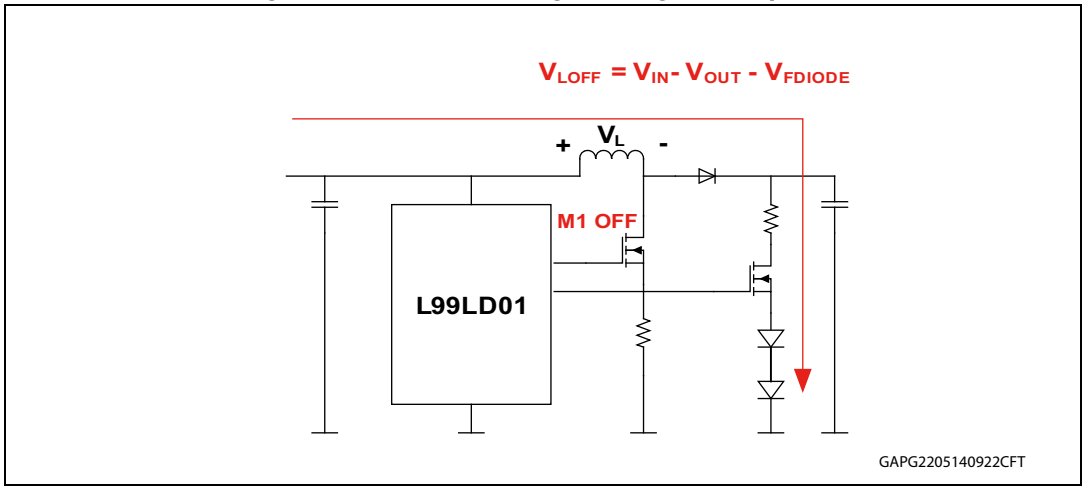


As the relation between the inductor's current and voltage is $V_L = L\frac{di_L}{dt}$, Integrating over t_{ON} gives:

$$\Delta I_{LON} = \frac{V_{L,ON}}{L} \times t_{ON} = \frac{V_{IN} - V_{SW}}{L} \cdot t_{ON} = \frac{V_{IN} - V_{SW}}{L} \times \frac{D}{F_{SW}}$$

During t_{OFF}, the inductor is demagnetized by the voltage $V_{L,OFF} = V_{IN} – V_{OUT} – V_{FDIODE}$

Figure 20. Inductor voltage during the off-phase



$$\Delta I_{LOFF} = \frac{V_{L,OFF}}{L} \times t_{OFF} =$$

$$= \frac{V_{IN} - V_{OUT} - V_{FDIODE}}{L} \times t_{OFF} = \frac{V_{IN} - V_{OUT} - V_{FDIODE}}{L} \times \frac{1 - D}{F_{SW}}$$

In steady state, by definition, the current starts at each switching cycle at the same value, which means that the current increase during t_{ON} is exactly compensated the current decrease during t_{OFF}.

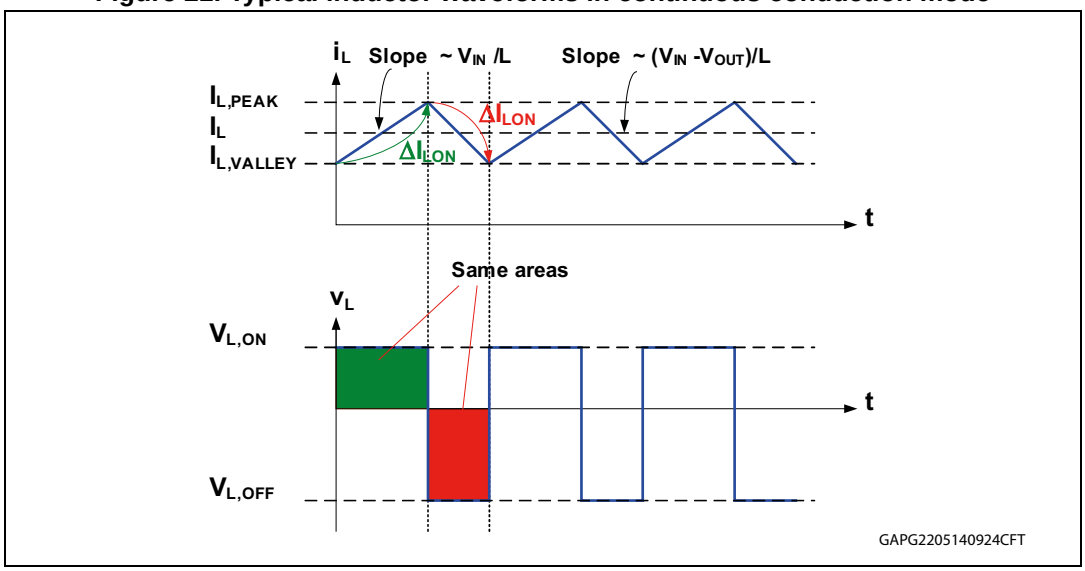
$$\Delta I_{L,ON} + \Delta I_{L,OFF} = 0$$

or

$$V_{L,ON} \times t_{ON} + V_{L,OFF} \times t_{OFF} = 0$$

This property is known as the volt-second balance. As a consequence, on the inductor voltage waveform (see *Figure 21*), the area under V_L in during t_{ON} is equal to the area above V_L during t_{OFF}.

Figure 21. Typical inductor waveforms in continuous conduction mode



$$(V_{IN} - V_{SW}) \times D = (V_{OUT} + V_F - V_{IN}) \times (1 - D)$$

Neglecting V_{SW} and V_F in the case of an ideal converter (without power loss and 100% efficiency) we have:

$$V_{IN} \times D = V_{OUT} \times (1 - D)$$

Extracting the duty cycle in the last equation:

$$D = \frac{V_{OUT} - V_{IN}}{V_{OUT}} = 1 - \frac{V_{IN}}{V_{OUT}}$$

8.5 Calculation of the mosfet RMS current

By definition:

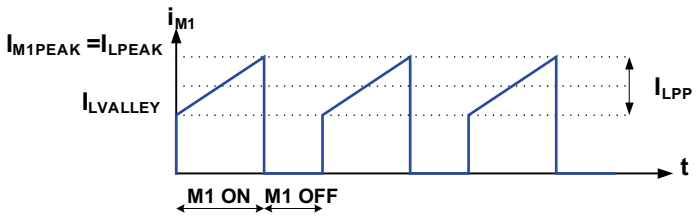
$$I_{M1,RMS}^{2} = \frac{1}{T}\int_{0}^{T} i_{M1}^{2} dt$$

With

$$i_{M1} = I_{L,VALLEY} + \frac{I_{L,PP}}{D \times T} \times t$$

for 0 < t < D x T and i_{M1} = 0 for D x T < t < T (see *Figure 15*)

Figure 22. Typical current waveform of M1 in CCM



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$$I_{M1,RMS}^{2} = \frac{1}{T}\int_{0}^{T} \left(I_{L,VALLEY} + \frac{I_{L,PP}}{D \times T} \times t\right)^{2} dt =$$

$$= \frac{D \times T}{3 \times T \times I_{L,PP}} \left[\left(I_{L,VALLEY} + \frac{I_{L,PP}}{D \times T} \times t\right)^{3}\right]_{0}^{DT}$$

$$I_{M1,RMS}^{2} = \frac{D}{3 \times I_{L,PP}} \left(\left(I_{L,VALLEY} + I_{L,PP}\right)^{3} - I_{L,VALLEY}^{3}\right)$$

$$I_{M1,RMS}^{2} = \frac{D}{3 \times I_{L,PP}} \left(\left(I_{L} + \frac{I_{L,PP}}{2}\right)^{3} - \left(I_{L} - \frac{I_{L,PP}}{2}\right)^{3}\right)$$

Using the factoring formula:

$$a^{3} - b^{3} = (a^{3} - b) \times (a^{2} - ab + b^{2})$$

$$I_{M1,RMS}^{2} = \frac{D}{3 \times I_{L,PP}} \left(I_{L,PP} \times \left(3I_{L}^{2} + \frac{I_{L,PP}^{2}}{4} \right) \right)$$

Equation 30:

$$I_{M1,RMS}^{2} = D \times \left(I_{L}^{2} + \frac{I_{L,PP}^{2}}{12} \right)$$

Since $I_L = \frac{I_{OUT}}{1-D}$, finally:

$$I_{M1,RMS} = \frac{I_{OUT}}{1-D} \times \sqrt{D \times \left(1 + \frac{1}{12} \left(\frac{I_{L,PP}}{I_L} \right)^{2} \right)}$$

8.6 Calculation of the freewheeling diode RMS current

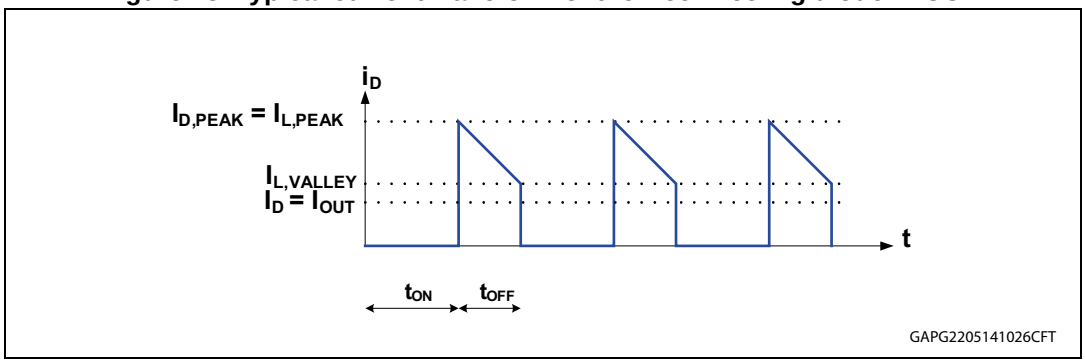
By similarity of the MOSFET (see *Figure 22*) and diode (see *Figure 23*) current waveforms: the diode RMS current can be derived from *Equation 30*.

The peak current and valley currents are the same. The diode conduct during (1 - D) x T in a switching cycle, whereas the MOSFET conducts during D x T. Therefore in the case of the diode, D must be replaced by 1 - D in *Equation 30*.

Equation 31:

$$I_{D,RMS}^{2} = (1-D) \times \left(I_{L}^{2} + \frac{I_{L,PP}^{2}}{12} \right)$$

Figure 23. Typical current waveform of the freewheeling diode in CCM



8.7 Calculation of the inductor RMS current

Figure 2 shows that:

$i_L = i_{M1}$ (and ($i_D = 0$)) for $0 < t < DT$

$i_L = i_D$ and ($i_{M1} = 0$) for $DT < t < T$

By definition,

$$I_{L,RMS}^2 = \frac{1}{T}\int_0^T i_L^2 dt = \frac{1}{T}\int_0^T i_{M1}^2 dt + \frac{1}{T}\int_0^T i_D^2 dt = I_{M1,RMS}^2 + I_{D,RMS}^2$$

Using *Equation 30* and *Equation 31*, we have:

Equation 32:

$$I_{D,RMS}^2 = I_L^2 + \frac{I_{L,PP}^2}{12}$$

Equation 33:

$$I_{L,RMS} = \sqrt{I_L^2 + \frac{I_{L,PP}^2}{12}} = \sqrt{\frac{I_{OUT}^2}{(1-D)^2} + \frac{I_{L,PP}^2}{12}}$$

or

Equation 34:

$$I_{L,RMS} = \frac{I_{OUT}}{1-D}\sqrt{1 + \frac{1}{12}\left(\frac{I_{L,PP}}{I_L}\right)^2}$$

Appendix C Document management

1. High efficiency constant current LED driver (L99LD01, DocID025319)
2. Switching power supplies A to Z. by S. Maniktala (ISBN13: 978-0-7506-7970-1)
3. Fundamental of Power Electronics by R. Erickson, D. Maksimovic (ISBN 0-7923-7270-0)
4. Switch-Mode Power Supplies by C. Basso (ISBN 978-0-07-150859-9)

Revision history

Table 5. Document revision history

Date	Revision	Changes
29-May-2014	1	Initial release.

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