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Buck converter for SSL applications

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Application note

Document information

Info	Content
Keywords	Buck, down, converter, driver, topology, AC/DC, DC/DC, SMPS, LED
Abstract	This document describes how to design a buck converter that can, for example, be used to drive an LED string and also illustrates the method of calculating components in the Boundary Conduction Mode (BCM).

Revision history

Rev	Date	Description
01	20091014	First release

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1. Introduction

The buck converter is one of the most common and often used Switch Mode Power Supply (SMPS) topologies. This topology is also called a down converter because of its main feature: The output voltage is always lower than the input voltage. A buck converter can be remarkably efficient (easily up to 95 % for integrated circuits) and self-regulating, which makes it useful for tasks such as converting the 12 V to 24 V typical battery voltage in a laptop down to the few volts needed by the processor. This topology can be used not only to convert voltage, but is also suitable to act as a current source, depending on the control method. A number of NXP LED drivers can operate in buck mode.

2. Scope

2.1 Scope

This application note discusses the general principles and considerations when designing a buck converter and especially a buck converter for LED's in boundary conduction mode with valley detect. It has a separate chapter for losses and key component calculation. This application note can be used when designing a buck converter using several of NXP's LED driver ICs, like the SSL1523, SSL1623, SSL2101, SSL2102 and UBA 3070. Dimmability and mains dimmability are not discussed within this section, as this is specific for each IC solution.

2.2 General philosophy of the application note

The setup of this document is made in such way, that a chapter on a related subject can be read with a minimum of cross-references to other documents or data sheets. This leads to some repetition as some of the information within this application note is also available in other, more dedicated application notes. In most cases, typical values are given to enhance the readability.

- [Section 3](#) discusses the theory of operation. It shows how voltages and currents flow during one converter cycle. It also gives a short overview of the trade-off between CCM and BCM/DCM modes.
- [Section 4](#) gives information on how to design key components, like the inductor value. It describes the resulting calculation of peak current when BCM, with valley detection, is used.
- [Section 5](#) shows some power calculations, to give the designer an insight into the loss mechanisms in the converter, and how his choices affect efficiency.
- [Section 6](#) briefly covers LED current tolerance and stability.

2.3 Related documents and tools.

Further information regarding design tools and the driver ICs mentioned in this document can be found on the product page for the specific IC (internet link), or are available through the local sales office.

3. Theory of operation

The operation of the buck converter is relatively simple, with an inductor and two switches that control the inductor input current. It alternates between connecting the inductor to source voltage to store energy in the inductor, and discharging the inductor into the load.

[Figure 1](#) shows a simplified application diagram of a buck converter, connected to a voltage supply and a load. For a basic understanding of the application, V_1 and V_o can be considered as DC. In a practical application, a MOSFET or bipolar transistor replaces the switch S1, and a diode replaces the switch S2.

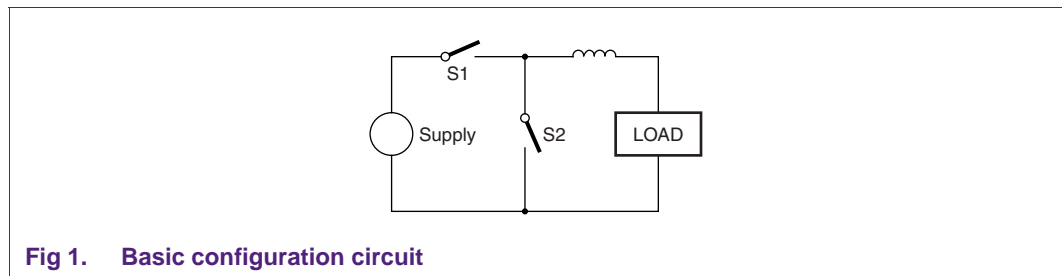


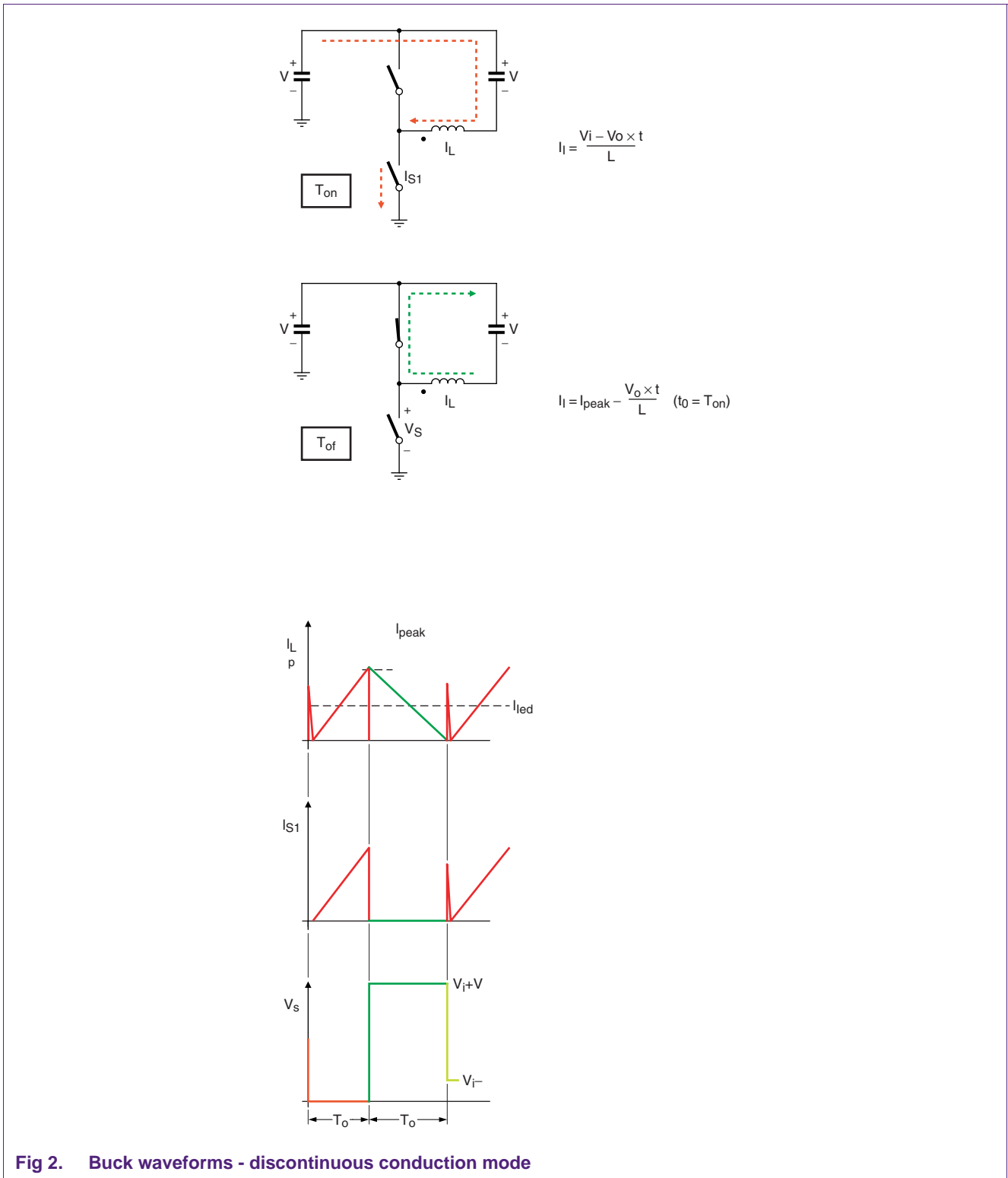
Fig 1. Basic configuration circuit

The circuit is defined by the state of the switches. With two switches there are four modes, but not all of them are applicable. Modes 1 and 2 are the most important and nearly always present while mode 3 is only present in Discontinuous Conduction Mode (DCM). Mode 4 must be prevented, since this would short circuit the supply. The state of the switches in modes 1 to 3 is displayed in [Table 1](#).

Table 1. Possible modes of operation

Mode	S1	S2	Duration
1	On	Off	$\delta_1 \times T$
2	Off	On	$\delta_2 \times T$
3	Off	Off	$\delta_3 \times T$

Operation of the buck-back converter is briefly explained on the next page. The figures show the equivalent circuit diagrams for the first two modes. Simplified waveforms for one complete switching cycle are shown also.



During the time $\delta_1 \times T$ (Mode 1) switch S1 is switched on and a current starts to flow through the inductor L. At the moment switch S1 is switched off the secondary switch S2 is closed and a current starts to flow towards the output. During the conduction time of switch S2, the energy in the conductor is reduced. δ_3 is entered when the current through

switch S2 has decreased to zero. The mode of operating just described is called Discontinuous Conduction Mode (DCM). The border between DCM and Continuous Conduction Mode (CCM) is reached when the time $\delta_3 \times T$ has become zero. This is called Boundary Conduction Mode (BCM)

Switch S2 is often replaced by a diode. It must be ensured that only Mode 3 is entered when the current through the inductor is zero. If both switches open when there is still current running through the inductor, the current will try and seek another path, and a very high voltage peak will be the result. This peak might damage the switches or the inductor. Using a suitable diode will prevent this.

Both Continuous mode and Discontinuous mode buck solutions are common, and each solution has the following specific advantages and disadvantages:

- The CCM converter has less input and output current ripple than the discontinuous mode version, so it requires less additional filtering.
- The CCM converter has lower core losses because less of the BH-curve is utilized. It must, however, have an inductor value inverse to the current ripple, which results in a much bigger core and more windings. This counters the lower core losses, and gives more wire losses.
- The CCM converter cannot be regulated to low values; the control margin is determined by the current ripple.
- The DCM converter has no hard current switching when S1 conducts, so a switching method that is optimized for low switch-off losses only can be used.
- The DCM converter makes full use of magnetic energy storage, so it can work with a smaller inductor.

The previous list illustrates that Discontinuous mode is the most effective solution for small form factor dimmable SSL solutions.

A BCM converter offers even more advantages, because Discontinuous mode has a dead time in which the inductor is not used. It offers the smallest size and the lowest switching losses, and full dimmability. The ripple current at both the input and output is however higher, so more buffering and filtering is needed to reduce this and to reach mains conducted emission standards like FCC15 and IEC55015.

4. Key components design procedure

This chapter guides you through the procedure for designing a Boundary Conduction mode buck converter for SSL applications:

4.1 Output current versus peak current

A typical minimal buck application circuit driving one string of LEDs is shown in [Figure 3](#): Starting parameter to design such a circuit is the required LED current and the LED voltage. Assuming the converter works exactly in boundary conduction mode, the relation between output current and inductor peak current is straight-forward:

$$I_{peak} = 2I_{led} \quad (1)$$

$$I_{peak} = \delta_1 \times \frac{V_I - V_o}{f \times L} \quad (4)$$

Combining 1 to 4 results in formula 5:

$$L_3 = \frac{I}{2 \times I_{led} \times f} \times \frac{V_o^2 - V_I \times V_o}{V_I} \quad (5)$$

Example:

f = 100 kHz, I_{led} = 700 mA, V_I = 200 V, V_O = 100 V. I_{peak} = 1.4 A, δ₁ = 50 %, L₃ = 357 μH.
t₁ = 5 μs, t₂ = 5 μs

and:

f = 100 kHz, I_{led} = 700 mA, V_I = 200 V, V_O = 10 V. I_{peak} = 1.4 A, δ₁ = 5 %, δ₂ = 9.5 %, L₃ = 67.8 μH, t₁ = 0.5 μs, t₂ = 9.5 μs

4.3 Valley detect

The next converter cycle can start just after t₂ has ended and the converter current has reached zero, but in doing this, the switch will switch on again with substantial voltage over it. There is a certain amount of capacitance over supply and switch, which is built up of several components:

- The parallel capacitance of the inductor.
- The reverse charge of the freewheel diode.
- The drain-gate capacitance of the switch.

When discharging this capacitance, the stored energy stored is dissipated in the switch:

$$P_{sw} = \frac{I}{2} \times C_p \times V_{sw}^2 \times f \quad (6)$$

Example: f = 100 kHz, V_{sw} = 200 V, C_p = 100 pF. P_{sw} = 200 mW.

As a result, the switch will heat up and the efficiency will go down. To overcome this, a feature has been built-in that is unique for NXP converters: This feature is called valley detect. There is special circuitry that senses when the voltage on the drain of the switch has reached its lowest value. Then, the next cycle is started. As a result, the switching losses can go down significantly.

There is however another effect: A time (t₃) is introduced in which there is little current running in the inductor. This time will last half the period of the resonant frequency:

$$t_{valley} = \pi \times \sqrt{L_p \times C_p} \quad (7)$$

Example: L_p = L₃ = 357 μH, C_p = 100 pF, t_{valley} = 0.594 μs

To be most effective, two conditions must be met:

- The excitation voltage (= V_O) must be close to half the input voltage.
- The L_pC_p combination must be under dampened.

$$V_O = \frac{I}{2} \times V_I \quad \text{and} \quad R_{ser}^2 \times C_p^2 - 4 \times L_p \times C_p \ll 0 \tag{8}$$

R_{ser} = the serial dampening resistor within the $L_p C_p$ circuit, and consists of coil resistance and magnetic losses.

Example: $V_I = 200 \text{ V}$, $V_O = 100 \text{ V} = 0.5V_I$, OK.

$R_{ser} = 1 \text{ } \Omega$, $C_p = 100 \text{ pF}$, $L_3 = 357 \text{ } \mu\text{H}$. $-1.43 \times 10^{-13} \ll 0$. OK

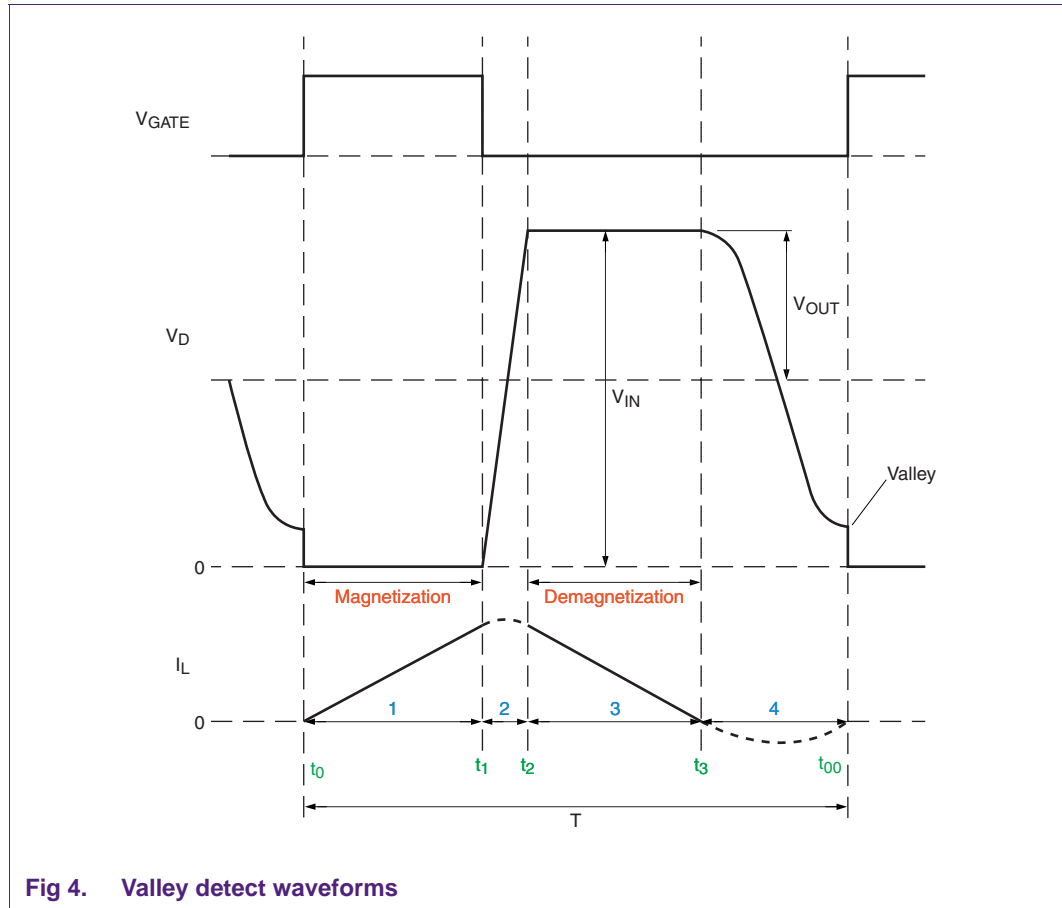


Fig 4. Valley detect waveforms

However, in order to reach the same LED current, the peak value must be adjusted, and this in turn will alter the converter frequency. The average current towards the output is given with [Equation 9](#), [Equation 10](#), [Equation 11](#) and [Equation 12](#):

$$I_{led} = I_{peak} \times \frac{t_1 + t_2}{2 \times (t_1 + t_2 + t_3)} \tag{9}$$

$$\varphi = \frac{V_o}{V_I + V_o} \tag{10}$$

$$t_2 = \frac{I_{peak} \times L_3}{V_o} \tag{11}$$

$$t_1 = \frac{I_{peak} \times L_3}{V_I - V_o} \tag{12}$$

Combining [Equation 9](#), [Equation 10](#), [Equation 11](#) and [Equation 12](#) results in [Equation 13](#):

$$2 \times I_{led} = \frac{\frac{I_{peak}^2 \cdot L_3}{V_o} \times (\varphi + 1)}{\frac{I_{peak} \times L_3}{V_o} \times (\varphi + 1) + t_3} \quad (13)$$

And when written out, gives [Equation 14](#)

$$0 = L_3 \times (\varphi + 1) \times I_{peak}^2 - 2 \times I_{led} \times (\varphi + 1) \times L_3 \times I_{peak} - 2 \times I_{led} \times t_3 \times V_o \quad (14)$$

This 2nd order function can be solved using the ABC formula:

$$a = L_3 \times (\varphi + 1) \quad (15)$$

$$b = -2 \times L_3 \times (\varphi + 1) \times I_{led} \quad (16)$$

$$c = -2 \cdot t_3 \cdot V_o \cdot I_{led} \quad (17)$$

$$I_{peak} = \frac{\pm \sqrt{b^2 - 4 \times a \times c}}{2 \times a} \quad (18)$$

Example: $\varphi = 1$, $a = 0.714 \times 10^{-3}$; $b = -1 \times 10^{-3}$, $c = -83.1 \times 10^{-6}$, $I_{peak} = 1.48$ A, $t_1 = 5.28$ μ s, $t_2 = 5.28$ μ s, $t_3 = 0.594$ μ s, $f' = 89.6$ kHz.

4.4 Peak current limit

In the example schematic, resistor R5 limits the peak current. When the voltage level over this resistor reaches a threshold, the cycle will stop, and the switch will stop conducting. This threshold can be used to control peak current. Using peak current control, the LED current is half the peak current in BCM mode. Also, the tolerance on this detection is proportional with the tolerance on LED current. If we call the threshold level V_{ocp} , then [Equation 19](#) can be used:

$$R_5 = \frac{V_{ocp}}{I_{peak}} \quad (19)$$

Example: $I_{peak} = 1.48$ A, $V_{ocp} = 0.52$ V, $R_5 = 0.35$ Ω .

4.5 Ripple current calculation

Component C5 filters the current through the LED's so this current will approach the average current through the inductor. The remaining alteration is called ripple, and can be expressed as percentage of the average current as indication. If the current waveform is symmetric, which will be the case with buck converters, the average will be half the sum of maximum and minimum current. Next formula gives an approximate result of the ripple current:

$$C_5 = \frac{1}{2 \times \pi} \times \frac{1}{f \times Ripple_{\%} \times R_{dyn}} \quad (20)$$

In the formula above, R_{dyn} is the differential resistance of the Led string at the rated average current. This value is derived by taking the tangent of the UI graph for the respective LED. It is not the division between voltage and current at the point of operation.

Example: 10 LED's are used in series at 100 mA. Each LED has a dynamic resistance of 1 Ω , so the total dynamic resistance is 10 Ω . At a ripple of 5 % and a frequency of 100 kHz, C5 will be 3.18 μ F. Take 3.3 μ F.

OR: 1 LED is used at 1 A. It has a dynamic resistance of 0.1 Ω . At a ripple of 1% and a frequency of 100 kHz, C5 will be 1.6 mF.

The value calculated within this formula is intended to filter ripple current caused by converter operation. This value is not intended to filter current variation due to input voltage fluctuation. Often, the input voltage ripple, especially when rectifying and buffering 50 Hz to 60 Hz mains voltage, has amplitude that does not allow the linear approximation as used in above formula. For mains buffer calculation use [Equation 21](#).

$$C3 + C4 = \frac{2 \times P_{tot} \times t_{dis}}{V_{mainspeak}^2 - V_{buff(min)}^2} \tag{21}$$

where:

$$P_{tot} = P_{in} + IC \text{ losses.}$$

4.6 Inductor design parameters

In buck converter designs, the importance of the main inductor L3 quality is often underestimated. To achieve a highly efficient solution, not only the inductance value, but also the ohmic losses, saturation current, proximity losses, core losses, parasitic capacitance and stray magnetic fields are important. Not understanding the functionality and implementing an optimized component, will result in either, inferior performance or an impractical design. Chapters [Section 4.6.1](#) to [Section 4.6.4](#) will give some guidelines.

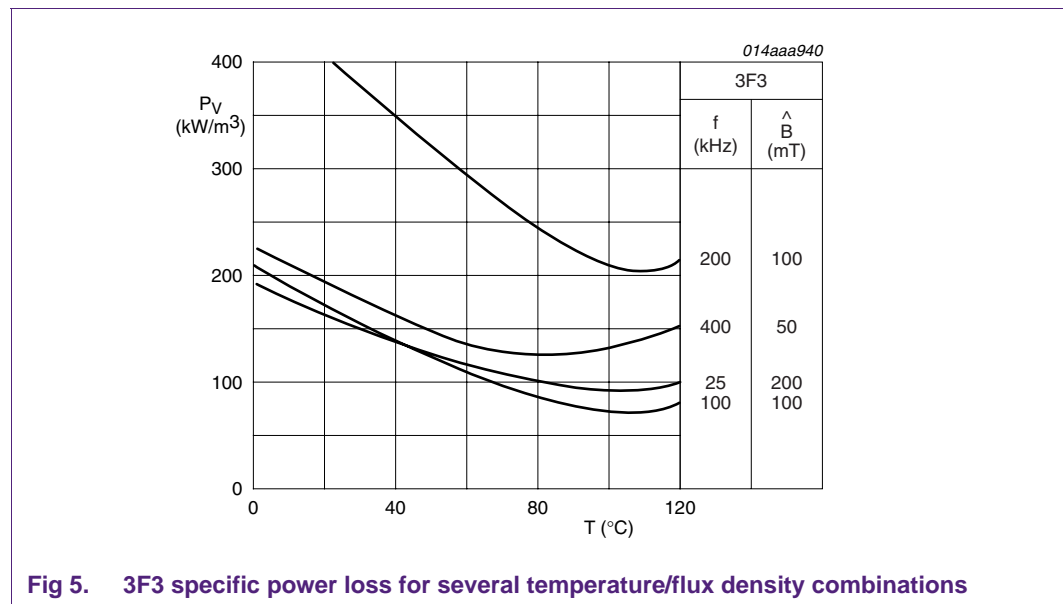


Fig 5. 3F3 specific power loss for several temperature/flux density combinations

For core material, each manufacturer has another code. For applications between 50 kHz and 200 kHz, 3F3 (Ferroxcube), N87 (Epcos) or TP4 (TDG), are recommended. Select the material that has optimum lowest loss at working temperature. A core material not suitable for the effective frequency of the converter will give high core losses.

Table 2. Ferrite Core Comparative Geometry Considerations

Aspect	Pot Core; RM Core	Double slab core	E core	Ec; ETD Cores	PQ Core	EP Core	Toroid
core costs	high	high	low	medium	high	medium	very low
bobbin costs	low	low	low	medium	high	high	none
winding costs	low	low	low	low	low	low	high
winding flexibility	good	good	excellent	excellent	good	good	fair
assembly	simple	simple	simple	medium	simple	simple	None
mounting flexibility	good	Good	good	fair	fair	good	poor
heat dissipation	poor	good	excellent	good	good	poor	good
shielding	excellent	good	poor	poor	fair	excellent	good

4.6.1 Core type selection

Core geometry depends on several factors, like cost, flexibility, shielding and utilization factors. A core can have an inner core that results in a round or square winding shape. The stray inductance can vary with core shape. Core size is determined by the maximum stored energy in the inductor together with the required air gap. A core with a large air gap can store more energy than a core with a small air gap. In practice, for Discontinuous mode converters, an optimum design is reached when core losses and winding losses (proximity and skin losses) are balanced. A compromise has to be made between high storable energy levels, low leakage inductance and small tolerances on the inductance. Using [Equation 22](#), the maximum energy stored in the inductor can be calculated.

$$E = \frac{1}{2} \times L_3 \times I_p^2 \quad (22)$$

Example: Core type: $L_3 = 357 \mu\text{H}$, $I_p = 1.48 \text{ A}$. $E = 0.39 \times 10^{-3} \text{ J}$.

[Figure 3](#) shows the RM core types that can be applied:

Table 3. Core selector

Core type	Material	Ag (μM)	Ue (N/A^2)	Le (mm)	Al (nH)	Ae (mm^2)
RM4	3H3-A100	160	154	20.9	100	11.0
RM4/I	3F3-A160	110	215	23.3	160	13.8
RM5	3H3-A250	110	201	21.2	250	21.2
RM5/I	3F3-A250	130	186	23.1	250	24.8
RM6S	3H3-A315	120	221	26.8	315	31.4
RM7/I	3F3-A250	240	135	30.0	250	44.1
RM8	3H3-A630	90	342	35.6	630	52.0
RM10/I	3H3-A1000	110	367	44.6	1000	96.6

4.6.2 Calculate windings

A_l is often specified in the data sheet of the core material. It relates to the inductive value of a single turn on the selected core. Using this figure, and knowing the inductance, the calculation of winding number is quite straightforward, as shown in [Equation 23](#):

$$N_{I3} = \sqrt{\frac{L_3}{A_l}} \quad (23)$$

Example: Core type: RM8 3H3-A630, $A_l = 630$ nH, $L_3 = 357$ μ H, $N = 24$

A practical value for N_p can be obtained by rounding the calculated value to its nearest integer. As double check, the maximum magnetic B-Field is determined by the magnetic material. Also be informed that the peak value of B-field reached during operation has substantial impact on core losses. We will not further discuss these losses, but as rule of thumb, state that the B-field in the magnetic material should remain below the specified B_{max} of the material. The B-Field can be estimated using [Equation 24](#):

$$B_{max} = u_e \times \frac{N_{I3} \times I_p}{l_e} \quad (24)$$

Example: Core type RM8 3H3-A630. $N = 24$, $I_p = 1.48$, $u_e = 342$, $l_e = 35.6$,
 $B_{max} = 342 \times 24 \times 1.48 / 35.6 = 338$ mT.

4.6.3 Auxiliary winding count

The auxiliary winding can be used for two purposes. Firstly, it can sense demagnetization of the inductor, and secondly, can generate the required voltage to power the controller. If the winding is generated for demagnetization only, the voltage generated can be much smaller than when using the winding to generate the V_{cc} . This affects the winding ratio. For demagnetization detection, both the negative and positive voltage should be larger than the threshold level. For V_{cc} generation using a single diode rectifier, it is most efficient to take the longest time of δ_1 and δ_2 and dimension the winding accordingly. For interval δ_2 , [Equation 25](#) applies:

$$V_{aux} = \frac{N_{aux}}{N_{I3}} \times V_{o(min)} \Rightarrow N_{aux} = \frac{V_{aux} \times N_{I3}}{V_{o(min)}} \quad (25)$$

Example: At $V_{o(min)} = 100$ V, $N_{I3} = 24$, $V_{aux} = 14$ V. $N_{aux} = 3.36$ rounded up to 4

Note that the voltage on the auxiliary winding should always higher than the minimum V_{cc} voltage as required by the IC. There is voltage loss over the inductor, the rectifier diode and there is ripple on the V_{cc} due to discharge. All these factors have to be taken into account. There is a direct relation between voltage on the auxiliary winding and converter output voltage. The output voltage depends of the sum of forward voltages of the attached LED's, so the minimum V_f should be taken as the starting point for checking if sufficient voltage is available on the auxiliary winding.

4.6.4 Select wire diameters

Wire diameter selection is a trade-off between available winding area, ohmic losses, proximity losses and skin losses. As rule of thumb, when using wire sizes below 0.6 mm diameter at operating frequencies below 200 kHz, the skin losses are negligible. Above 0.6 mm diameter, it is recommended that Litze wire or multiple strands are used. Skin depth can be calculated using [Equation 26](#):

$$\delta = \sqrt{\frac{2 \times \rho}{2 \times \pi \times f_{eff} \times u_r \times u_o}} \tag{26}$$

In which: $u_o = 0.4 \times \pi \times 10^{-6}$, $\rho = \text{resistivity} = 17 \times 10^{-9}$ (copper). U_r (copper) = 1.

Example: At 100 kHz sinusoidal current, using copper, the skin depth will be 0.21 mm.

The effective frequency does not correspond to the converter frequency, but to the harmonics of the applied waveform. For a triangular wave current, the amplitude and frequency of the waveforms can be deduced using Fourier analyses. The amplitude of the coefficients depends on the ratio between δ_1 and δ_2 . as can be seen in [Table 4](#).

Table 4. Harmonics amplitude coefficients

Ratio	1st	2nd	3rd	4th	5th	6th	7th
0.05	0.334	0.165	0.108	0.078	0.060	0.048	0.039
0.2	0.372	0.151	0.067	0.023	0	0.010	0.003
0.5.	0.405	0	0.045	0	0.016	0	0.008

A higher converter ratio will also give more high order harmonics, as well as increased core and proximity losses in the transformer. These harmonics must be filtered in order to be EMC compliant, so more, or better, input and output filtering is required.

The ohmic losses depend on the peak currents in the wires. They can be estimated by simply calculating the wire resistance, and calculating the average power dissipation in the wire. As an approximation, the current density should be between 300 and 500 CM (Circular mills)/Amp. [Table 5](#) shows wire sizes relative to current:

Table 5. Wire selection table

Dia (mm)	Nearest AWG	Area (mm ²)	Area (CM)	DC Res. Ohm/M	Typical Current I _l (Amp)
0.1	38	0.008	15	2.195	0.04
0.2	32	0.031	62	0.549	0.15
0.25	30	0.049	97	0.351	0.24
0.315	28	0.078	154	0.221	0.38
0.355	27	0.099	195	0.174	0.49
0.4	26	0.126	248	0.137	0.62
0.56	23	0.246	486	0.070	1.22
0.71	21	0.396	781	0.044	1.95
16 × 0.2	-	0.503	992	0.034	2.48
37 × 0.2	-	1.162	2294	0.015	5.73
61 × 0.2	-	1.916	3782	0.009	9.45

4.7 V_{CC} generation dimensioning.

When the auxiliary winding is also used for V_{CC} generation, one should take the following aspects into account:

- At startup, the converter is not working, so no voltage is generated in the auxiliary winding. There should always be a startup circuit present that provides sufficient current to the V_{CC} for the first few cycles.
- The voltage on the auxiliary winding depends on the output voltage, so one should use the worst case situations to calculate whether minimum power demands are met and if dissipation and current values are within limits.
- Voltage is only present during part of the cycle. The average current flowing towards the V_{CC} of the IC should be sufficient to drive the IC. Thus, the peak current flowing should be higher than the required average current.

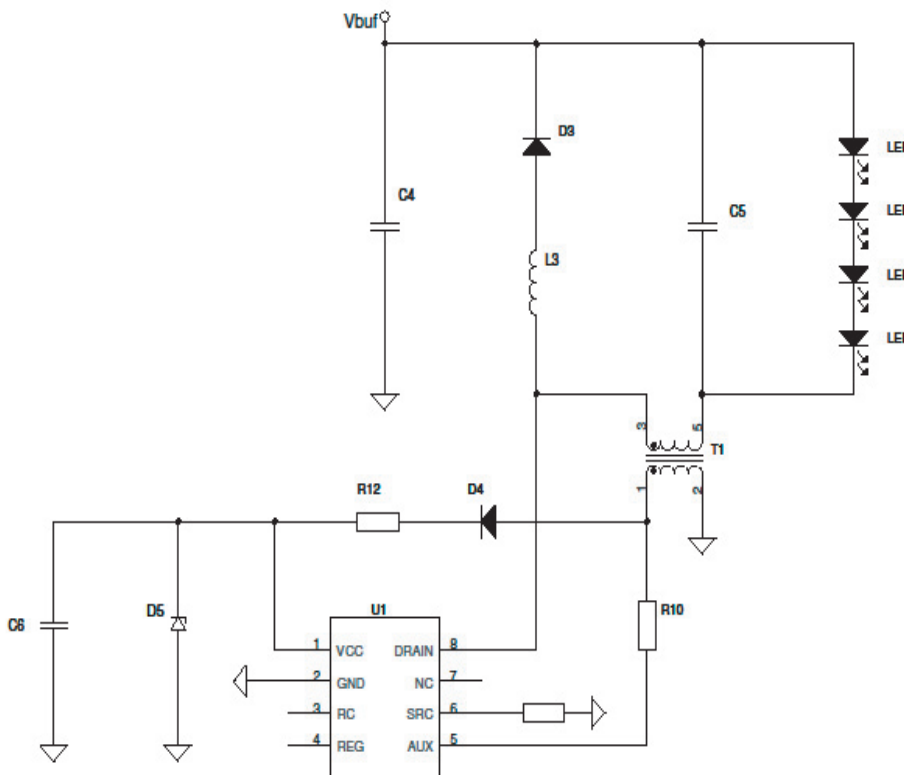


Fig 6. V_{CC} generation circuit

Example 1: At $V_{aux} = 14$ V, $I_{cc} = 2$ mA at 12 V, $\delta 2 = 46$ %. $V(R12) = 14 - 12 - 0.7 = 1.3$ V, $I(R12) = 2$ mA / 0.46 = 4.34 mA.

$R12 = 1.3$ V / 4.34 mA = 299 Ω . Round down gives 270 Ω .
 $P(R12) = I^2 \times R \times \delta 2 = 2.4$ mW.

Example 2: At $V_{aux} = 18$ V, $I_{cc} = 2$ mA at 12 V, $\delta 2 = 4$ %. $(R12) = 18 - 12 - 0.7 = 5.3$ V.
 $I(R12) = 2$ mA / 0.04 = 50 mA.

$R12 = 5.3$ V / 50 mA = 106 Ω . Rounding down gives 100 Ω . $P(R12) = 10$ mW.

If the circuit is dimmable, the calculation has to be redone for the worst case situation. Some ICs like the SSL1523 and SSL2101 have internal HV generation. If sufficient drain voltage is available, the IC is capable of providing its own supply. Notice that a smaller interval of current charge and a bigger tolerance leads to over-dimensioning of this circuit. It gives more losses in the serial resistor (R12), diode (D4) and inductor. The margin might be such, that additional protection against V_{cc} over-voltage might be necessary. A zener diode (D6) is included in the circuit for this purpose.

4.7.1 Buffer capacitance C6 calculation:

When V_{cc} is generated, the incoming current has to be buffered in order to provide a continuous and stable voltage. The voltage drop over C6 should be such that V_{cc} does not drop below the minimum voltage. [Equation 27](#) can be applied for this minimum capacitor value:

$$C_6 = \frac{I_{cc} \times \Delta t}{\Delta V} \quad (27)$$

Example: At $\Delta V = 1.3$ V, $I_{cc} = 2$ mA, $\Delta t = 6$ μ s, C6 will be at least 9.23 nF.

In practice, the value for C6 is chosen much higher to reduce noise and capacitive coupling with the surroundings. Between 1 μ F and 4.7 μ F are common values.

4.8 Demagnetization detection dimensioning

On several of the NXP ICs, there is a demagnetization detection functionality. This often uses a pin that has certain minimum and maximum threshold voltage. For the NXP range of LED drivers, this level is set at +100 mV and -100 mV. There is also a negative and positive clamping diode with threshold around 0.5 V. These clamping diodes can have a maximum current level, $I_{demag(max)}$. When using an auxiliary winding, the current should be limited to a level where the threshold voltages are reached and the maximum current not exceeded.

Example: At $V_{aux} = 14$ V, 100 μ A < I_{aux} < 5 mA. R_{aux} (R10) = $14 / 100 \times 10^{-6} = 140$ K Ω .

Be aware that the demagnetization detection is phase dependent: The winding direction should be opposite to the main inductor in order to start next cycle at low valley. Reversing the winding will result in switching at top detection.

5. Power calculations

The resulting efficiency of a buck converter is one of the critical specifications for the design. One of the things to consider, is that efficiency is always relative. Part of the losses of a buck converter, like the IC V_{CC} generation, are fixed and depend on the IC. Because of fixed losses, efficiency tends to go down with lower output power. The variable losses consist of a number of factors, and these are discussed in next sub-chapters. The formulas in these chapters will give the designer insight in the parameters that determine the losses for each component.

5.1 Resistive switch dissipation

Besides capacitive losses from [Equation 7](#), there are also ohmic losses in the switch. The main parameters that determine these losses are the resistance of the switch, which is expressed as R_{DSon} for MOSFET switches and peak current. There is a momentary peak dissipation and the average dissipation.

$$P_{peak} = I_{peak}^2 \times R_{DSon} \quad (28)$$

Over period t_1 , the total dissipation will be as shown in [Equation 28](#)

$$P = \frac{1}{3} \times I_{peak}^2 \times R_{DSon} \quad (29)$$

And over the total time period, the average ohmic switch dissipation will become as given in [Equation 30](#):

$$P = \frac{1}{3} \times t_1 \times f \times I_{peak}^2 \times R_{DSon} \quad (30)$$

Example: $f = 89.6 \text{ kHz}$, $R_{DSon} = 2.2 \text{ } \Omega$, $I_{peak} = 1.48 \text{ A}$, $t_1 = 5.28 \text{ } \mu\text{s}$. $P = 0.76 \text{ W}$.

Remark: The size of these losses is mostly determined by peak current and R_{DSon} of the switch.

5.2 Capacitive switch dissipation

The capacitive switch losses have already been discussed in [Section 4.3 Equation 6](#). It is important to note that these losses are independent of the peak current and thus also the LED current. By using valley detection and an output voltage that is half the input voltage, these capacitive switching losses can be avoided.

Without this option, the balance between switch size causing lower R_{DSon} losses, and capacitive switch losses will shift: A bigger switch will have lower R_{DSon} , but higher drain capacitance. In such a situation, the optimum has to be selected:

For the next examples, $I_{peak} = 1.48 \text{ A}$, $t_1 = 5.28 \text{ } \mu\text{s}$, $V_{sw1} = 100 \text{ V}$, $f = 89.6 \text{ kHz}$.

Example 1: $I_{\delta} = 1.5 \text{ A}$, $R_{DSon} = 5.5 \text{ } \Omega$, $C = 300 \text{ pF}$, $P_{RDSon} = 1.9 \text{ W}$, $P_{CSW} = 0.13 \text{ W}$, $P_{tot} = 2.03 \text{ W}$

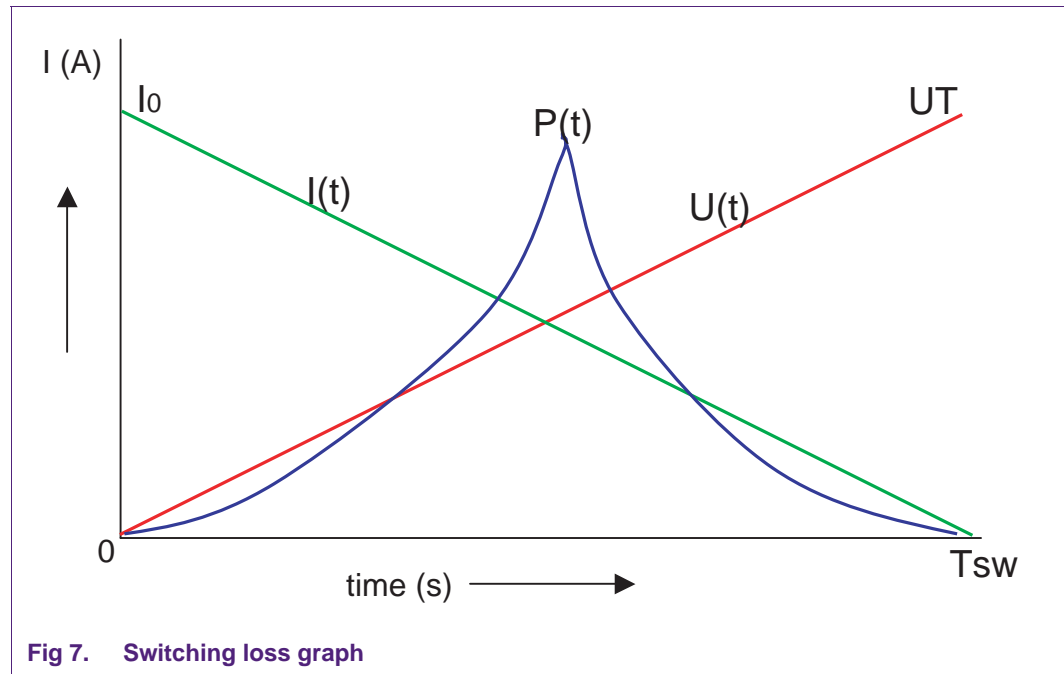
Example 2: $I_{\delta} = 3.5 \text{ A}$, $R_{DSon} = 2.2 \text{ } \Omega$, $C = 550 \text{ pF}$, $P_{RDSon} = 0.76 \text{ W}$, $P_{CSW} = 0.24 \text{ W}$, $P_{tot} = 1 \text{ W}$

Example 3: $I_0 = 13 \text{ A}$, $R_{DSon} = 0.42 \text{ } \Omega$, $C = 3.1 \text{ nF}$, $P_{RDSon} = 0.14 \text{ W}$, $P_{CSW} = 1.39 \text{ W}$, $P_{tot} = 1.53 \text{ W}$

Remark: From the three examples above, example 2 has best performance.

5.3 Switching losses

Besides the capacitive losses, there are also losses occurring due to hard switching of the current. This happens when the switch closes. The data sheet of IC or MOSFET specifies a switching slope at which the FET closes. During this time, there is an overlap of the current and voltage, and this overlap causes dissipation. Assuming the current drops and the voltage rise is linear within a fixed time 'T', the dissipation can be calculated:



$$I(t) = I_0 - \frac{I_0}{T} \times t = I_0 \times \left(1 - \frac{t}{T}\right) \quad U(t) = U_T \times \frac{t}{T} \tag{31}$$

$$P(t) = I(t) \times U(t) = I_0 \times \left(1 - \frac{t}{T}\right) \times U_T \times \frac{t}{T} \tag{32}$$

$$E = \int P(t) dt = \int_0^T I_0 \times \left(1 - \frac{t}{T}\right) \times U_T \times \frac{t}{T} dt \Leftrightarrow \tag{33}$$

$$E = \int_0^T \frac{I_0 \times U_T}{T^2} \times t dt \quad - \int_0^T \frac{I_0 \times U_T}{T^2} \times t^2 dt \Leftrightarrow \tag{34}$$

$$E = \frac{1}{2} \times \frac{I_0 \times U_T}{T} \times T^2 - \frac{1}{3} \times \frac{I_0 \times U_T}{T^2} \times T^3 = \tag{35}$$

$$\frac{1}{6} \times I_0 \times U_T \times T \quad P_{eff} = \frac{1}{6} \times I_0 \times U_T \times f \times T_{sw} \tag{36}$$

Example: $T = 100 \text{ nS}$, $I_o = 1.5 \text{ A}$, $U_T = 200 \text{ V}$, $f = 88 \text{ kHz}$. $P_{\text{eff}} = 0.44 \text{ W}$.

This dissipation increases with the switching time. When using valley detection, these losses are reduced at switch-on. They are still present at switch-off.

5.4 Freewheel diode losses

The freewheeling diode has two loss mechanisms: The forward losses and the reverse charge losses. The forward, or conductive, losses can be estimated using the time versus the current and voltage drop given in [Equation 37](#): and [Equation 38](#).

$$P_f = I_{led} \times V_f \times t_2 \times f \quad (37)$$

$$P_{rev} = \frac{1}{2} \times V_I^2 \times C_{rev} \times f \quad (38)$$

Example: $f = 89.6 \text{ kHz}$, $I_{led} = 0.7 \text{ A}$, $t_2 = 5.28 \text{ }\mu\text{s}$, $V_f = 0.7 \text{ V}$, $V_I = 200 \text{ V}$, $C_{rev} = 10 \text{ pF}$.
 $P_f = 230 \text{ mW}$, $P_{rev} = 18 \text{ mW}$.

The forward voltage of the diode can be lowered using a Schottky diode, but these diodes are often difficult to obtain with reverse voltages above 100 V. Also, care should be taken not to oversize this diode, as this does not appreciably lower the forward losses, but the reverse charge is often directly related to the maximum current of the diode.

5.5 Inductor losses

The inductor has several loss mechanisms. Calculation of these losses is very complex and there is much debate on the way these losses contribute to the total inductor losses. [Section 5.5](#) will simply illustrate a number of loss mechanisms within the inductor.

5.5.1 Ohmic losses

The combination of wire length and thickness causes ohmic losses. The calculation of the resistance and losses can be derived from [Equation 39](#) and [Equation 40](#):

$$R_{dc} = \rho \times \frac{l}{A} \quad (39)$$

$$P_{dc} = \frac{1}{T} \times \int_0^T I^2 \times R_{dc} dt = \frac{1}{3} \times I_p^2 \times R_{dc} \quad (40)$$

Example: Wire length 1 m at diameter 0.56 mm. $\rho_{Cu} = 17.2 \times 10^{-9}$.
 $A = \pi \times R^2 = 0.246 \times 10^{-6}$. $R_{dc} = 70 \text{ m}\Omega$. $I_p = 1.48 \text{ A}$. $P_{dc} = 51 \text{ mW}$

5.5.2 Proximity losses

For proximity losses, the full calculations are outside the scope of this application note. What should be made clear, however, is that they are closely related to the skin depth and number of windings. See [Figure 8](#):

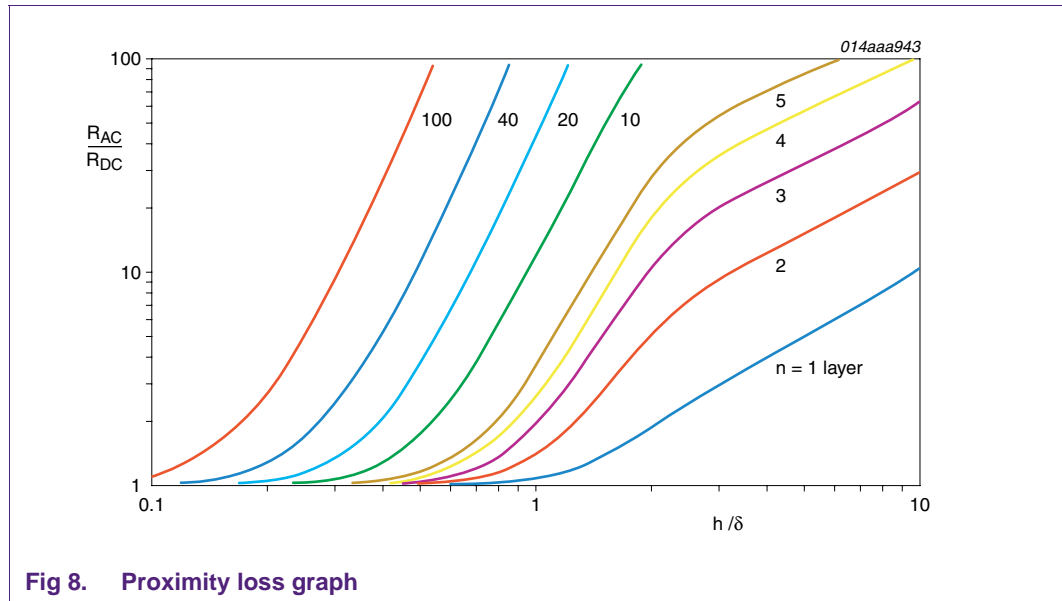


Fig 8. Proximity loss graph

Too many layers of wires with radius that is close to, or below skin-depth should be avoided. Normally, the proximity losses are calculated as a factor of the dc wire resistance: $R_{ac} = n \times R_{dc}$.

Keeping low resistive losses helps to lower proximity losses. This is another drawback of the CCM mode inductor. To achieve higher inductance, more windings are required, thus increasing DC and AC resistance and countering the lower core losses.

5.5.3 Core losses

The core losses in the magnetic material are determined by the magnetization curve and frequencies. At each converter cycle, the magnetic field in the core material is excited by the magnetic flux density producing a curve that is highly non-linear with the saturation level and hysteresis. The surface area enclosed by the variation in B-field strength at a certain frequency determines the losses. A bigger core, a higher B-field and a higher frequency increase these losses. The core material data sheet shows the loss per unit of volume at given frequencies.(see [Figure 10](#))

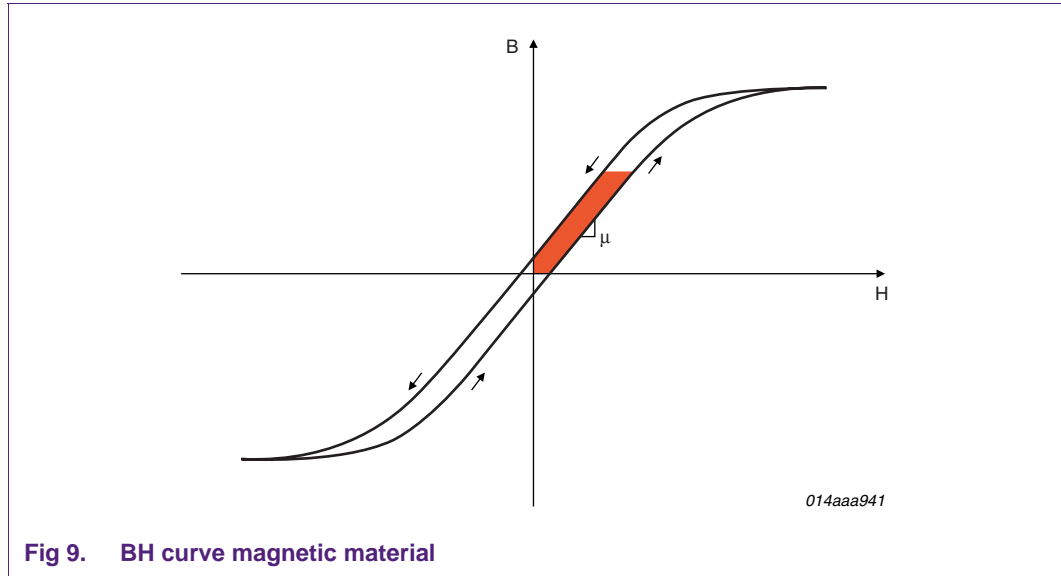


Fig 9. BH curve magnetic material

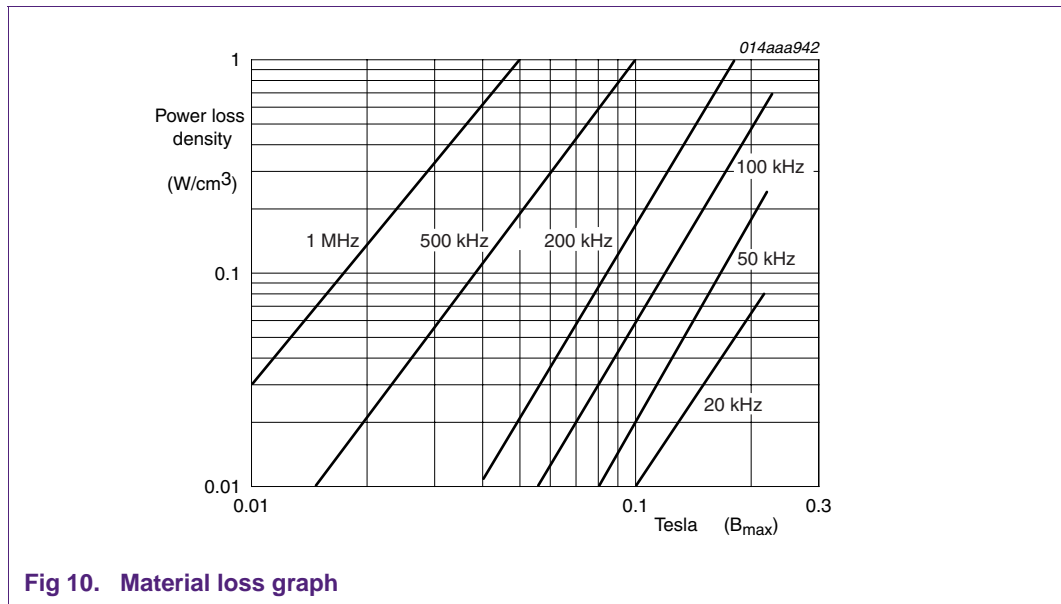


Fig 10. Material loss graph

A simple empirical formula that calculated core loss is called a Steinmetz equation as shown in [Equation 41](#):

$$P_h = K_h \times f^\alpha \times B_{max}^\beta \times V_{core} \tag{41}$$

K_h and α are dependent of core material. This formula can be improved by including the harmonics of a square waveform as shown in [Equation 42](#):

$$P_{nse} = K_h \times (2f)^\alpha \times B_{max}^\beta \times (D^{1-\alpha} + (1-D)^{1-\alpha}) \times V_{core} \tag{42}$$

In [Equation 42](#), D is duty cycle, B_{\max} the peak flux density, f is the frequency of the fundamental and V_{core} the volume of the core. We can see that a higher frequency, higher flux density, smaller duty-factor and bigger volume all increase core losses. A bigger core might not always reduce core losses; if the B-field is already low, the increase in volume will counter the lower losses due to reduce flux density.

Example: At a duty cycle of 50 %, $K_h = 0.05$, $f = 80$ (kHz), $\alpha = 1.8$, $B_{\max} = 1.00$ mT, $\beta = 3$ and $V_{\text{core}} = 2.4$ cm³, the loss will be $0.05 \times (160)^{1.84} \times (0.1)^3 \times 3.58 \times 2.4^{-6} = 1.36$ W.

5.6 Sense resistor losses.

For the sense resistor losses, [Equation 30](#) can be applied. R_{DSon} is replaced with sense resistor value.

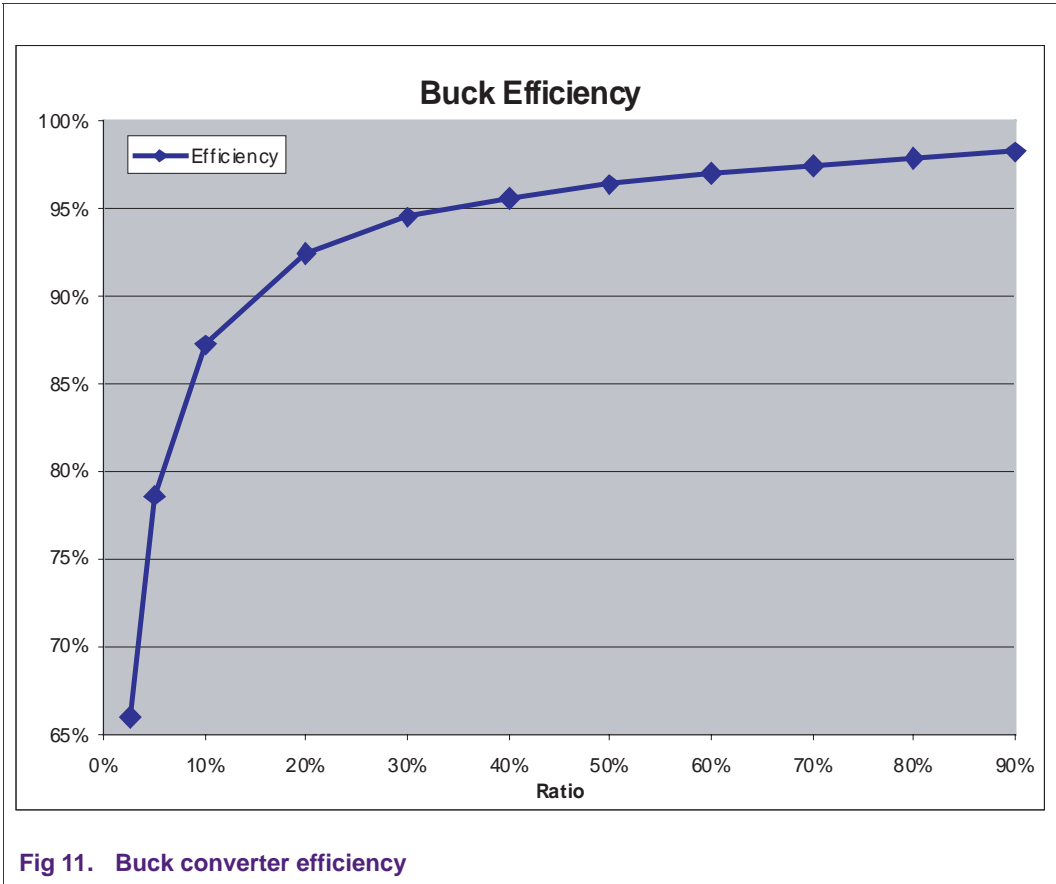
Example: $I_{\text{peak}} = 1.48$ A, $R_{\text{sense}} = 0.5 / 1.49 = 0.33$ Ω , $P_{\text{sense}} = 80$ mW

5.7 Total system losses

When undertaking buck power calculations, it is vital to consider the impact of each individual loss on the total converter efficiency. To illustrate this aspect, we will take a single driver with set current, a 200 V input voltage, and we will vary the output voltage in steps. The relation between the input and the output voltage is plotted on the X-axis as a ratio. Each loss mechanism will be calculated, and the resulting driver efficiency plotted.

As shown in [Figure 11](#), the converter efficiency starts to drop at a low ratio. This is not caused by excessive increase in losses, but by the relative decrease of useful output power. Some losses, like the magnetic core losses, and the ohmic losses in the switch, are reduced. There is an increase in the forward losses in the freewheel diode, the capacitive switching losses and the losses due to hard switching. This makes sense, since the conduction time t_2 of the freewheel diode is large and the voltage drop over the switch is also large.

It can be seen that a very low resistance switch will be more helpful at a large ratio, like 50 % to 90 %. Low switch capacitance, low forward voltage of the freewheel diode and fast switching are more effective at a small ratio, 5 % to 20 % for example.



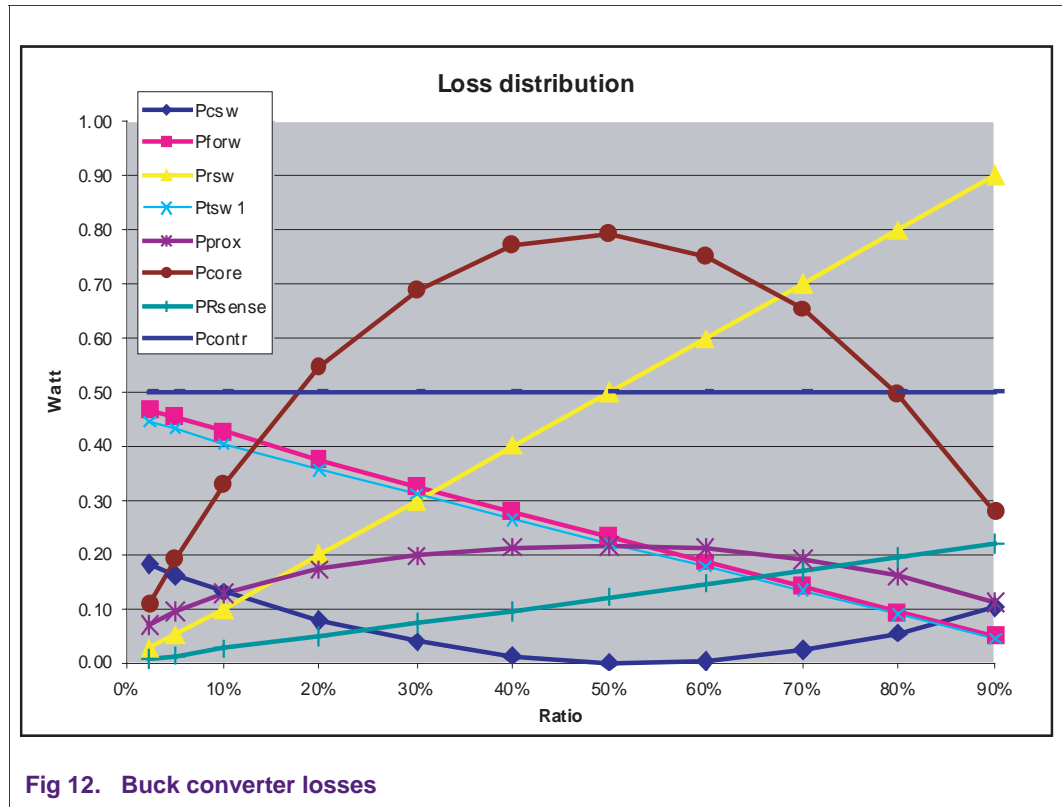


Fig 12. Buck converter losses

6. Current tolerance and stability

6.1 Current tolerance

In essence, there are only two main components that determine current tolerance: the spread on detection voltage and the tolerance sense resistor. This can be derived from [Equation 43](#):

$$\Delta I_{led} = \Delta I_{peak} = \Delta V_{ocp} + \Delta R_6 \tag{43}$$

Example: $V_{ocp(min)} = 0.48 \text{ V}$. $V_{ocp(avg)} = 0.50 \text{ V}$ $V_{ocp(max)} = 0.52 \text{ V}$. $\Delta V_{ocp} = \pm 4 \%$.
 $\Delta R_6 = \pm 1 \%$. $\Delta I_{led} = \pm 5 \%$

There is some influence possible by variation of C_p and L_p with valley detection, but in practice, the time influenced is much smaller than the total cycle time.

Example: $\Delta L_p = 10 \%$. $\delta 3 / T = 0.052$. $\Delta I_{led} = 0.5 \times \Delta L_p \times 0.05 = 0.25 \%$.

6.2 Current stability

For the buck using peak current control, stability is seldom an issue because the current is controlled per cycle. It is intrinsically stable. If some other means to stabilize the current is used, like current mirror detection, accuracy might increase but the loop response must be calculated. The main component that determines response at peak current control is

output capacitor C5. It has to be charged and discharged. At switch-on, the discharged capacitor will have to reach working voltage first before any current flows through the LED's and light is produced. This time is equal to the charge time of formula 21.

Example: At $\Delta V = 100 \text{ V}$, $I = 700 \text{ mA}$, $C6 = 3.3 \text{ }\mu\text{F}$, Δt will be at least $471 \text{ }\mu\text{s}$.

At turn-off, the diode characteristic of the LED will come in play. Instead of a sudden drop, there will be an exponential drop of current, starting with the nominal current. The LED current will slowly fade until not visible. In practice, this might take several seconds. Since the LED's are placed in a self-rectification loop with the freewheel diode, any capacitive coupling on the drain side, or inductive coupling over the loop with an AC source will induce a current through the LEDs. Even a small current of for instance $100 \text{ }\mu\text{A}$ might be visible. If large, ungrounded objects like heat-sinks connected to phase are in close proximity of the LEDs, this might happen.

7. Summary

This document gives an overview of operations and calculation of the buck converter in discontinuous conduction mode. It explains why valley detection is a key feature, and it shows how a number of key components can be calculated. It closes by a description of loss mechanisms in the converter, and how they contribute to driver efficiency.

8. Abbreviations

Table 6. Abbreviations

Acronym	Description
BCM	Boundary Conduction Mode
CCM	Continuous Conduction Mode
CM	Circular Mills
EMC	Electro Magnetic Compatibility
DCM	Discontinuous Conduction Mode
IC	Integrated Circuit
LED	Light Emitting Diode
MOSFET	Metal Oxide Semiconductor Field Effect Transistor
SMPS	Switch Mode Power Supply
SSL	Solid State Lighting

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