AN2928
Application note

Modified buck converter for LED applications

Introduction

The use of high power LEDs in lighting applications is becoming increasingly popular due to rapid improvements in lighting efficiency, longer life, higher reliability and overall cost effectiveness. Dimming functions are more easily implemented in LEDs, and they are more robust and offer wider design flexibility compared to other light sources.

Applications suitable for the use of LEDs include lighting for streets, stadiums, fairs and exhibitions, shops, interiors, as well as for decorative lighting, outdoor wall lighting and consumer lighting such as lamps and ballasts. Therefore, LED use for lighting is likely to represent an increasingly large proportion of the lighting market in the future. To assist engineers in their design approach, the STEVAL-ILL013V1 80 W offline PFC LED driver demonstration board has been developed. This application note describes, step-by-step, all the principles and calculations used for a modified buck converter intended for high brightness LED applications.

The converter is designed as a constant current source to achieve the best lighting performance from the LEDs. A “modified buck" topology was chosen because the power switch is connected to ground rather than the high side switch, as in a standard buck topology, so with this solution it is easier to control the switch. The design uses a fixed off-time (FOT) network operating in continuous conduction mode (CCM), rendering the overall solution simple and cost-effective. The modified buck converter described in this document can be used for lighting applications from low power and low voltage, to high power and high voltage. This allows designers to cover a wide range of different LED systems using a single topology.

Additionally, in lighting applications where the input active power is higher than 25 W and a high power factor is required, the high PF converter can be connected as the first stage, before the modified BUCK converter. The STEVAL-ILL013V1 shows this design concept.

The STEVAL-ILL013V1 demonstration board is an 80 W offline dimmable LED driver with high power factor (PF) intended for 350 mA, 700 mA and 1 A LEDs, and is based on STMicroelectronics’ L6562A transition-mode PFC controller. The design is complaint with standard EN61000-3-2 (limits for harmonic current emissions). The order code is STEVAL-ILL013V1 and the complete design, including schematic diagram, bill of material, calculations, measurements, etc. is described in user manual UM0670 (see *Section 3: Reference and related materials*).

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1 Modified buck converter in constant current mode

As stated in the introduction, the aim of this application note is to describe a modified buck converter working in FOT and CCM. The basic principle of the design using the L6562A controller is shown in *Figure 1* and *Figure 2*. *Figure 1* represents the stage when the power MOSFET Q_1 is turned on. As shown by the red arrow, the current flows from the DC voltage input (V_{IN}) through the load (LEDs), the inductor (L), the power MOSFET Q_1 and the sensing resistor. Capacitor C_4 is charged via diode D_2 and resistor R_5, since the transistor Q_1 is open and its gate voltage is around 10 V. During the t_{ON} time, the load current increases and stops as soon as the voltage on the current sense resistor reaches the internal threshold on the CS pin of the L6562A. The current sense of the L6562A is clamped at 1.08 V (typ). *Figure 2* shows the t_{OFF} time, when the power MOSFET is switched off. The inductor keeps the current flowing in the same direction and the circuit is closed through diode D_1. The load current is decreasing and the minimum current is set by the fixed off-time network (t_{OFF} time is always constant), because capacitor C_4 is discharged to the resistor R_4. The voltage on capacitor C_4 is connected to the ZCD (zero current detector) pin of the L6562A. As soon as the capacitor is discharged and its voltage falls below 0.7 V (the ZCD threshold), the L6562A switches the power MOSFET again and the load current is increased. This process repeats cycle-by-cycle, as shown in the timing diagrams in *Figure 1* and *Figure 2*.

Figure 1. Modified buck converter - t_{ON} time

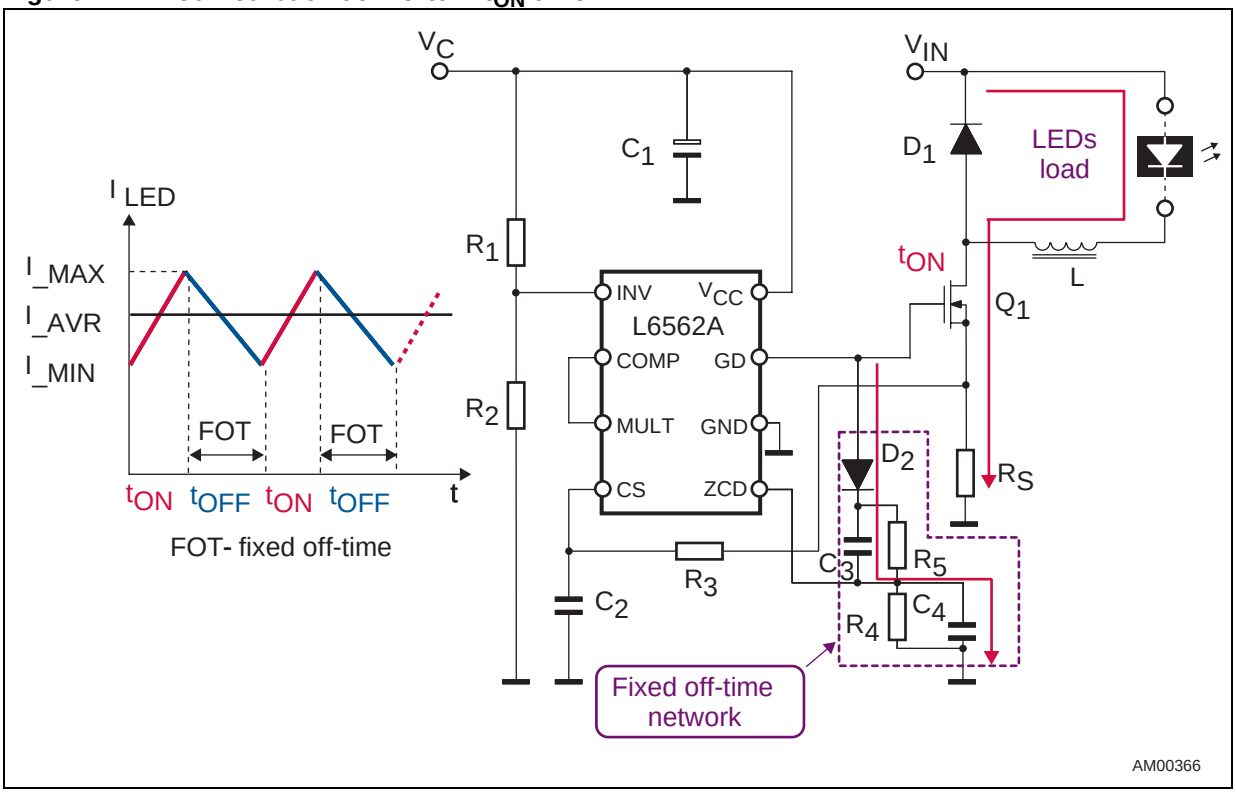
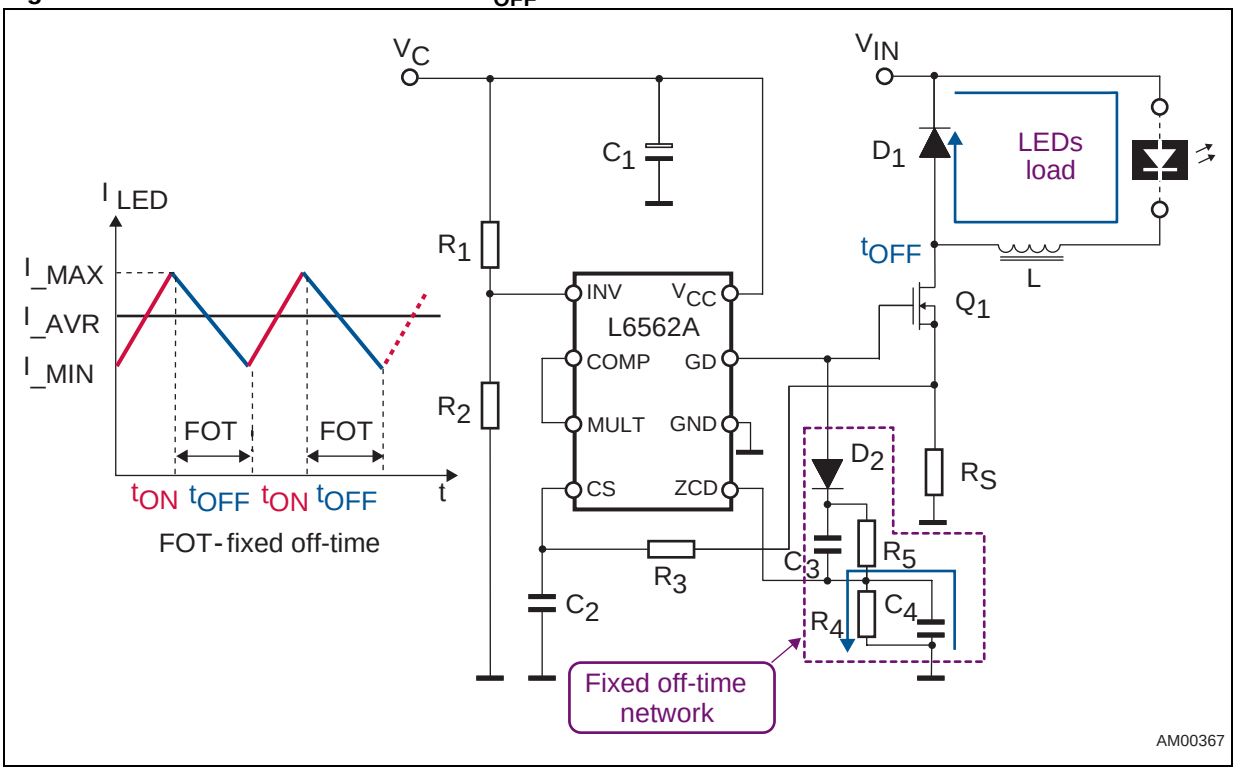


Figure 2. Modified buck converter - t_{OFF} time



2 Design equations for the modified buck converter

This section provides all the calculations required for a designer to develop an application with the modified buck converter working in FOT and CCM. The equations are described step-by-step, following an application design procedure. First, the basic equations for this type of converter are shown, then the components for the t_{OFF} time are calculated, the proper power diode and power MOSFET is selected, and finally the power inductor calculation is demonstrated.

2.1 Basic equations for the modified buck converter

Figure 3 shows basic circuit stage during t_{ON} and t_{OFF} time, with indicated voltage and component references used in the equations.

The voltage across the inductor L is calculated using the following equation:

Equation 1

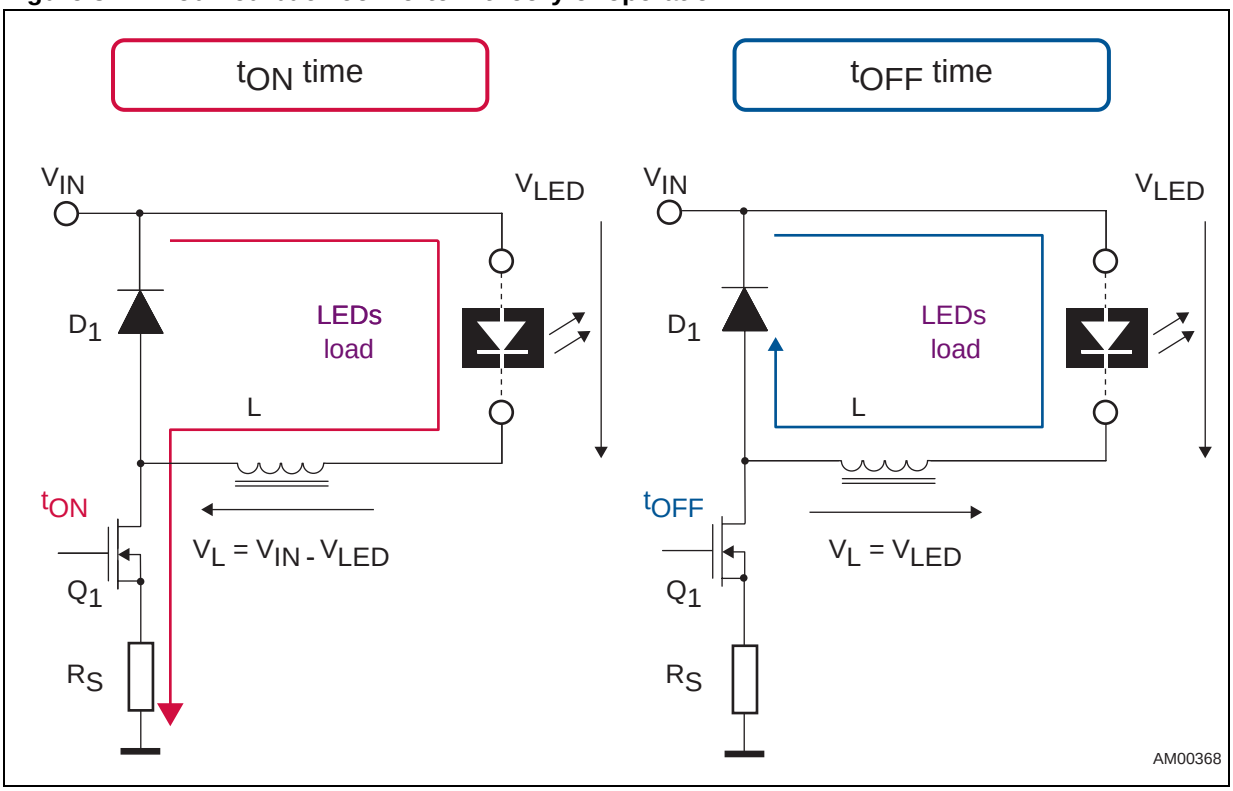
$$V_L = L \cdot \frac{dI_L}{dt}$$

V_L= inductor voltage (V)

L = inductance (H)

I_L = inductor current (A)

Figure 3. Modified buck converter - theory of operation



Using *Equation 1*, it is possible to calculate an inductor current change during t_{ON} and t_{OFF} time:

Equation 2

$$\Delta I_{L_ON} = \int_{0}^{t_{ON}} \frac{V_L}{L} dt = \frac{(V_{IN} - V_{LED}) \cdot t_{ON}}{L}$$

Equation 3

$$\Delta I_{L_OFF} = \int_{t_{ON}}^{t_{ON}+t_{OFF}} \frac{V_L}{L} dt = -\frac{V_{LED} \cdot t_{OFF}}{L}$$

ΔI_{L_ON} = inductor current change during t_{ON} time (A)

ΔI_{L_OFF} = inductor current change during t_{OFF} time (A)

V_{IN} = input voltage (V)

V_{LED} = LED (load) voltage (V)

t_{ON} = turn-on time (s)

t_{OFF} = turn-off time (s)

In CCM, the inductor current change during t_{ON} and t_{OFF} time is the same:

Equation 4

$$\Delta I_{L_ON} = \Delta I_{L_OFF}$$

Using *Equation 2* and *Equation 3*, it is possible to create following equations:

Equation 5

$$\frac{(V_{IN} - V_{LED}) \cdot t_{ON}}{L} = \frac{(V_{LED} - t_{OFF})}{L}$$

Equation 6

$$V_{IN} \cdot t_{ON} - V_{LED} \cdot t_{ON} = V_{LED} \cdot t_{OFF}$$

Equation 7

$$V_{IN} \cdot t_{ON} = V_{LED} \cdot (t_{OFF} + t_{ON}) = V_{LED} \cdot T$$

The duty cycle for the modified buck topology (also valid for a standard buck topology) converter is calculated:

Equation 8

$$D = \frac{t_{ON}}{T} = \frac{V_{LED}}{V_{IN}}$$

2.2 Fixed off-time network calculation

The basic idea for this type of converter is to obtain a constant off-time when the power MOSFET is turned off. This design approach is quite simple and cost-effective, because the constant off-time is easily set by the RC network. Its calculation is described in this section.

The duty cycle is defined by the switching frequency and turn-off time:

Equation 9

$$D = \frac{t_{ON}}{T} = \frac{T - t_{OFF}}{T} = 1 - \frac{t_{OFF}}{T} = 1 - f \cdot t_{OFF}$$

f = switching frequency in CCM (Hz)

T = period in CCM (s)

D = duty cycle (-)

From *Equation 8* and *Equation 9*, the turn off-time can be calculated (the switching frequency is selected):

Equation 10

$$1 - f \cdot t_{OFF} = \frac{V_{LED}}{V_{IN}}$$

Equation 11

$$t_{OFF} = \frac{\left(1 - \frac{V_{LED}}{V_{IN}}\right)}{f}$$

As stated above, the modified buck converter uses a FOT network. The off-time is set by resistor R_4 and capacitor C_4, as shown in *Figure 2*. During the on-time, the gate voltage of the power MOSFET is high, diode D_2 is opened and the voltage at the ZCD pin is internally clamped at $V_{ZCD_CLAMP} \approx$ 5.7 V. During the off-time, the gate voltage of the power MOSFET is low, diode D_2 is closed and the voltage at the ZCD pin decreases based on an exponential law:

Equation 12

$$V_{ZCD} = V_{ZCD_CLAMP} \cdot e^{\left(\frac{1}{R_4 \cdot C_4}\right)}$$

The voltage at the ZCD pin decreases until it reaches the internal triggering limit, which causes switching to the turn-on stage. The trigger voltage for the L6562A is 0.7 V. The time needed for the ZCD voltage to go from V_{ZCD_CLAMP} to $V_{ZCD_TRIGGER}$ defines the duration of the off-time t_{OFF}:

Equation 13

$$t_{OFF} = R_4 \cdot C_4 \cdot \ln\left(\frac{V_{ZCD_CLAMP}}{V_{ZCD_TRIGGER}}\right) = R_4 \cdot C_4 \cdot \ln\left(\frac{5.7}{0.7}\right) = R_4 \cdot C_4 \cdot 2.1$$

V_{ZCD} = voltage on the ZCD pin of the L6562A (V)

V_{ZCD_CLAMP} = clamp voltage on the ZCD pin of the L6562A (V)

$V_{ZCD_TRIGGER}$ = trigger voltage on the ZCD pin of the L6562A (V)

Capacitor C_4 can be selected, and the resistor is easily calculated using *Equation 14*, or the inverse can be calculated with *Equation 15*.

Equation 14

$$R_4 = \frac{t_{OFF}}{2.1 \cdot C_4}$$

Equation 15

$$C_4 = \frac{t_{OFF}}{2.1 \cdot R_4}$$

Resistor R_5 (shown in *Figure 2*) is designed to limit the charge current flowing to capacitor C_4 during the t_{ON} time, and must be within the following range (as described in application note AN1792. See *Section 3: Reference and related materials*).

Equation 16

$$\left[\frac{V_{GD_MAX} - V_{ZCD_CLAMP} - V_F}{I_{ZCD_MAX} + \frac{V_{ZCD_CLAMP}}{R_4}}\right] < R_5 < R_4 \cdot \left(\frac{V_{GD_MIN} - V_{ZCD_CLAMP} - V_F}{V_{ZCD_CLAMP}}\right)$$

V_F = diode D_2 forward voltage (V) (typically 0.7 V)

V_{ZCD_CLAMP} = clamp voltage on the ZCD pin of L6562A (V) (5.7 V)

V_{GD_MAX} = output high - maximum gate driver voltage (V) (15 V)

V_{GD_MIN} = output high - minimum gate driver voltage (V) (9.8 V)

I_{ZCD_MAX} = maximum sink capability for the ZCD pin (A) (0.01 A)

If the t_{ON} time is very short (light load or low output voltage), then capacitor C_4 cannot be quickly charged via resistor R_5. Therefore it is recommended to connect capacitor C_3 in parallel with resistor R_5. The maximum size of capacitor C_3 can be calculated with the following equation:

Equation 17

$$C_3 < C_4 \frac{V_{ZCD_CLAMP}}{(V_{GD_MAX} - V_{ZCD_CLAMP} - V_F)}$$

2.3 LED current calculation

The preceding equations (*Equation 12* through *17*) are used to calculate the FOT network using resistors R_4 and R_5 and capacitors C_3 and C_4. At this point, focus shifts to the power circuit because the output LED current and inductor must be calculated. For the inductor current change, the following equation is used:

Equation 18

$$\Delta I_L = I_{MAX} - I_{MIN} = 2 \cdot (I_{MAX} - I_{AVR})$$

ΔI_L = inductor (LED) current change (A)

I_{MAX} = maximum (LED) inductor current (A)

I_{MIN} = minimum (LED) inductor current (A)

I_{AVR} = average (LED) inductor current (A)

Combining *Equation 18* and *Equation 3*, considering the current change as an absolute value (i.e. positive), it is possible to derive:

Equation 19

$$\Delta I_L = \frac{V_{LED} \cdot t_{OFF}}{L} = 2 \cdot (I_{MAX} - I_{AVR})$$

where t_{OFF} is calculated using *Equation 11*.

From *Equation 11* and *Equation 19* the equation for deriving the inductor size can be formulated (maximum and average LED current is selected):

Equation 20

$$L = \frac{V_{LED} \cdot (1 - D)}{2 \cdot (I_{MAX} - I_{AVR}) \cdot f} = \frac{V_{LED} \cdot t_{OFF}}{2 \cdot (I_{MAX} - I_{AVR})}$$

Sense resistor R_S can be easily calculated because the voltage threshold on the CS pin for the L6562A is 1.08 V and therefore the resistor size is following:

Equation 21

$$R_S = \frac{V_{CS}}{I_{MAX}} = \frac{1.08}{I_{MAX}}$$

V_{CS} = current sense threshold (V)

Although the modified buck converter using the FOT network and working in CCM that is described here works as a constant current source, a limitation is the current dependency on the output voltage (number of LEDs). To understand this limitation it is necessary to derive the average inductor current (which is the LED current) from *Equation 20*. The result is shown in *Equation 22*, which provides the information listed.

Equation 22

$$I_{AVR} = I_{MAX} - \frac{V_{LED} \cdot (1-D)}{2 \cdot L \cdot f} = I_{MAX} - \frac{V_{LED} \cdot t_{OFF}}{2 \cdot L}$$

- I_{MAX} is constant and set by the resistor R_S
- t_{OFF} is constant and set by the FOT network
- The average inductor (LED) current is independent of the input voltage
- The average inductor (LED) current depends slightly on the voltage across the LEDs (i.e. number of LEDs) and therefore the design shows the best results using a fixed number of LEDs. A variable number of LEDs results in less current precision.

2.4 Power MOSFET calculation

The power MOSFET is chosen based on maximum stress voltage, maximum peak MOSFET current, total power losses, maximum allowed operating temperature and the driver capability of the L6562A.

Maximum stress voltage on the power MOSFET (drain-source voltage) for this modified buck converter is equal to the input voltage. The power MOSFET must be selected with some voltage margin. For example, if the input voltage is maximally 400 V, then maximum drain-source voltage should be 450 V or higher.

Maximum peak MOSFET current was selected in order to calculate the inductor size in *Equation 20*. Also in this case, the power MOSFET must be chosen with some current margin.

Total power losses on the power MOSFET must be calculated, due to the importance of designing a proper heat sink to avoid temperature stress on the power MOSFET. Basically, total power losses on the power MOSFET occur through conduction losses (depending on the $R_{DS(ON)}$), switching losses and gate charge loss caused by charging up the gate capacitance and then discharging this capacitance to ground. The gate charge loss is very small compared to the conduction and switching losses, so it is not used for further calculations. For total power MOSFET loss, a valid equation is:

Equation 23

$$P_{TOT} = P_{CON} + P_{SW}$$

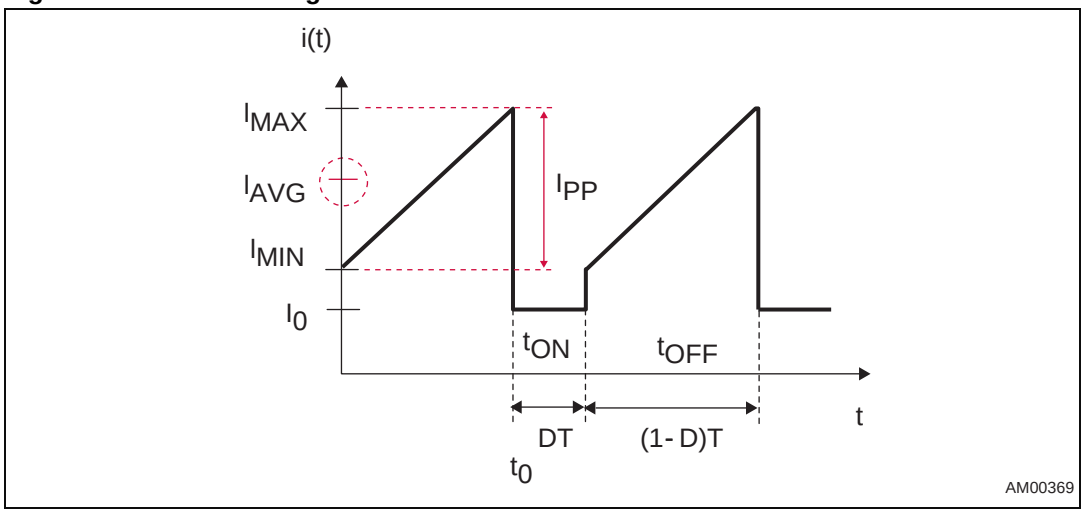
P_{TOT} = total power losses on the power MOSFET (W)

P_{CON} = conduction losses on the power MOSFET (W)

P_{SW} = switching losses on the power MOSFET (W)

The power MOSFET conduction loss is represented by the continuous conduction current flowing through the MOSFET during the on-time stage. Therefore, the power loss depends on its static drain-source resistance ($R_{DS(ON)}$). In order to calculate conduction loss properly, it is necessary to calculate the drain current RMS value. *Figure 4* shows the sawtooth signal for which the RMS value was calculated in *Equation 24*. Note that in this case, average current I_{AVR} is defined as the average value of the sawtooth portion ($(I_{MAX}+I_{MIN})/2$).

Figure 4. Sawtooth signal



Equation 24

$$I_{RMS}^2 = \frac{1}{T}\int_{t_0}^{t_0+T} i^2(t)dt = \frac{1}{T}\left[\int_{t_0}^{t_0+DT} i^2(t)dt + \int_{t_0+DT}^{t_0+T} i^2(t)dt\right] = D \cdot I_0^2 + (1-D) \cdot \left[\left(\frac{I_{MAX}+I_{MIN}}{2}\right)^2 + \frac{I_{PP}^2}{12}\right]$$

I_{RMS} = root-mean squared drain current (A)

I_{PP} = peak-to-peak current (A) (difference between $I_{_MAX}$ and $I_{_MIN}$)

The real drain current waveform is given in *Figure 5*. As can be observed, the signal is quite similar to the signal in *Figure 4*, except that there is no current during the off-time stage (I_O = 0) and the on and off-times are reversed. Therefore, for real current *Equation 24* is modified to become *Equation 25*.

Square RMS drain current is calculated using the equation:

Equation 25

$$I_{RMS}^2 = D \cdot \left[\left(\frac{I_{MAX}+I_{MIN}}{2}\right)^2 + \frac{I_{PP}^2}{12}\right] = D \cdot \left[I_{AVR_ON}^2 + \frac{I_{PP}^2}{12}\right]$$

I_{AVR_ON} = average current during t_{ON} time (A) (see *Figure 5*).

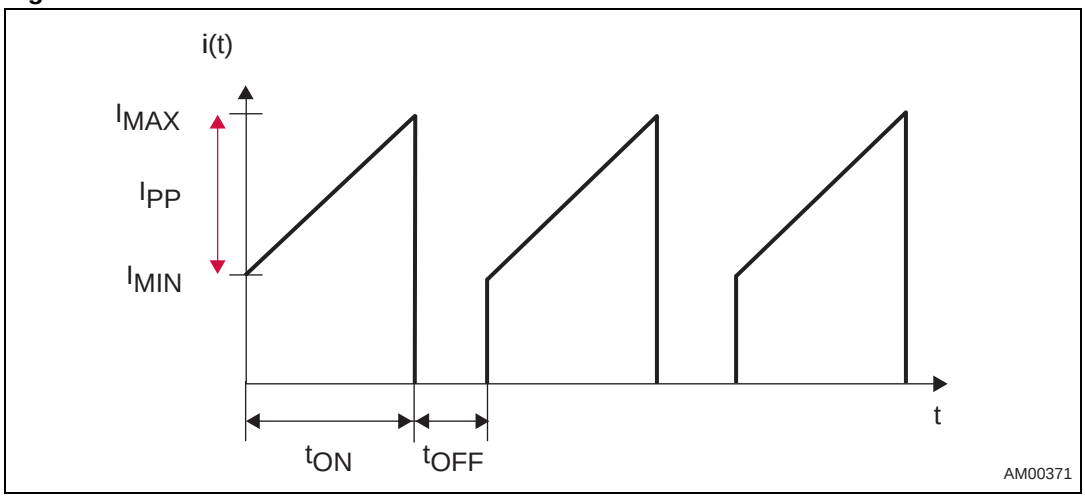
And finally it is possible to calculate the continuous conduction losses:

Equation 26

$$P_{CON} = I_{RMS}^2 \cdot R_{DS(ON)}$$

$R_{DS(ON)}$ = static drain-source on resistance for working MOSFET temperature (Ω).

Figure 5. Real drain MOSFET current



The second part of the MOSFET losses is switching losses, which depend on the (on and off) switching time, drain MOSFET current, drain-source voltage and the switching frequency. The switching time, rise time and fall time is a function of the gate-to-drain “Miller” charge of the MOSFET ($Q_{_GD}$), the internal resistance of the driver, the threshold voltage ($V_{GS(TH)}$), and the minimum gate voltage which enables the current through the drain source of the MOSFET. As the correct calculation of switching power losses is complex due to non linear behavior of the switch, it is not possible to obtain an exact equation for the calculation of switching losses. Moreover, the switching behavior is also influenced by the performance of the driver and layout design (leakage inductance and parasitic capacitors). To arrive at an estimation of the switching power losses, *Equation 27* can be used.

Equation 27

$$P_{SW} = \frac{V_{IN} \cdot I_{MAX} \cdot t_{OFF_SW} \cdot f}{2}$$

t_{OFF_SW} = switch off-time (s) (typically tens of ns).

For example, switched off-time measured on the STEVAL-ILL013V1 using the STP9NM50N power MOSFET (400 V input voltage) is 120 ns.

Total power ($P_{_TOT}$) is lost in the power MOSFET and its heat sink, so it is simple to calculate:

Equation 28

$$P_{TOT} = \frac{T_{JMAX} - T_A}{R_{thJC} + R_{thCH} + R_{thHA}}$$

T_{JMAX} = maximum junction temperature (°C)

T_A = ambient temperature (°C)

R_{thJC} = junction-to-case thermal resistance (°C/W)

R_{thCH} = case-to-heat sink thermal resistance (°C/W) (usually between 0.35 and 0.8 for the insulating washer)

R_{thHA} = heat sink-to-ambient thermal resistance (°C/W)

Finally, if the heat sink and its thermal resistance is known, it is possible to calculate maximum static drain-source on resistance from *Equation 26*, *Equation 27* and *Equation 28* for easy power MOSFET selection.

Equation 29

$$R_{DS(ON)} < \frac{T_{JMAX} - T_A}{(R_{thJC} + R_{thCH} + R_{thHA}) \cdot I_{RMS}^2} - \frac{V_{IN} \cdot I_{MAX} \cdot t_{OFF_SW} \cdot f}{2 \cdot I_{RMS}^2}$$

2.5 Power diode selection

The power diode (D_1 from *Figure 1*) is chosen based on its maximum stress voltage, its maximum peak current and total power losses. The power losses are lower for a larger duty cycle and vice-versa, because the diode is opened (connected) during off-time.

Maximum voltage stress across the diode is equal to the input voltage $V_{IN,}$ and therefore the power diode must be selected with some voltage margin. For example, if the input voltage is maximally 400 V, then maximum repetitive peak reverse voltage (V_{RRM}) should be 450 V or higher.

Maximum peak diode current is selected in order to calculate the inductor size in *Equation 20*. Also in this case, the power diode must be selected with some current margin.

Power losses are generally calculated with the following equation:

Equation 30

$$P_{LOSS_D} = \frac{1}{T}\int_0^T i_D(t) \cdot u_D(t)dt$$

P_{LOSS_D} = power diode losses (W)

i_D = power diode current (A)

u_D = power diode voltage (V)

And assuming a constant voltage drop over the diode it is possible to approximately calculate the power losses on the diode (switching losses are not included) with the following equation:

Equation 31

$$P_{LOSS_D} = I_{AVR_D} \cdot V_F$$

I_{AVR_D} = power diode average current (A)

V_F = power diode forward voltage for calculated average diode current (V)

where the average diode current is shown in *Figure 6* and can be calculated using:

Equation 32

$$I_{AVR_D} = (I - D) \cdot \left(\frac{I_{MAX} + I_{MIN}}{2}\right)$$

And finally the junction diode temperature without using the heat sink can be calculated from the following equation (ambient temperature is chosen):

Equation 33

$$T_J = P_{LOSS_D} \cdot (R_{thJC} + R_{thCA}) + T_A$$

T_J = power diode junction temperature (°C)

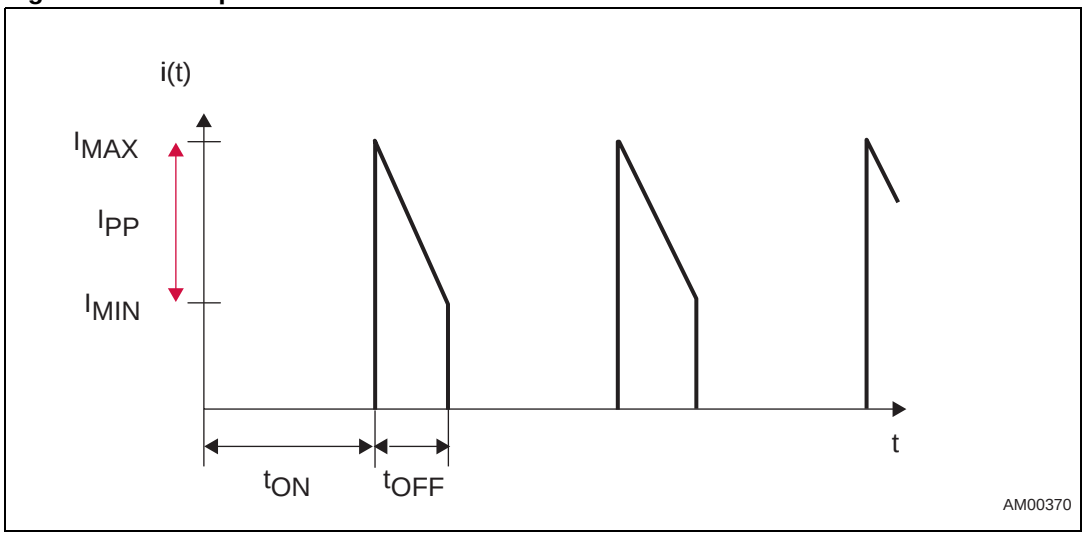
R_{thCA} = case-to-ambient thermal resistance (°C/W) (for example, the TO-220 package has a typical thermal resistance of 60 °C/W)

The calculated power diode junction temperature must be lower then maximum diode junction temperature T_{JMAX}. For proper design, it is recommended to keep the junction temperature much lower than its maximum in order to avoid temperature stress on the power diode.

Equation 34

$$T_J < T_{JMAX}$$

Figure 6. Real power diode current



2.6 Inductor calculation

All components for the design are calculated, but the final step in the design procedure still remains, since it is necessary to calculate the inductor (L). The calculations that follow are valid for the inductor used in the STEVAL-ILL013V1 demonstration board, but for applications with (for example) lower voltages, some standardized inductors for DC-DC can also be used.

First, the inductor core size must be selected, for which it may be helpful to calculate the minimum area product using application parameters. The minimum required core area product (AP), where the flux swing is limited by core saturation is:

Equation 35

$$AP_{MIN} = \left(\frac{L \cdot I_{PEAK} \cdot I_{RMS}}{B_{MAX} \cdot CI}\right)^{\frac{4}{3}}$$

AP_{MIN} = minimum area product (cm^4)

I_{PEAK} = inductor peak current (A)

I_{RMS} = inductor RMS current (A)

B_{MAX} = saturation limited flux density (T) (power ferrites like N27 or N67 have typically 0.3 T)

Equation 36

$$CI = J_{MAX} \cdot C_R \cdot 10^{-4}$$

J_{MAX} = maximum current density (A/cm^2) (That commonly used for natural convection cooling is 420 A/cm^2)

C_R = ratio of the total copper area to the window area (-)

The constant C_R gives an estimation of how effectively the wires are placed on the core. For example, if the input voltage of the modified buck converter is 400 V (such as when a power factor preregulator is used), then the wires must be well-isolated and the ratio between the copper and window area is about 0.5, which in the end means that there is 50% of the wire on the inductor core. The exact calculation using the equation listed here can be found in the user manual UM0670 (see *Section 3: Reference and related materials*).

Once the minimum area product is calculated, the designer should then select the right inductor core and ferrite material according to the higher AP value. The AP value is calculated from the winding cross section and core cross section. For example, the ETD29 core with ferrite material N27 from EPCOS has a winding cross section of 97 mm^2 and a core cross section of 71 mm^2. Maximum flux density for the N27 is 0.3 T.

Equation 37

$$AP = A_N \cdot A_{MIN}$$

A_N = winding cross section (mm^2)

A_{MIN} = minimum core cross section (mm^2)

For a proper inductor core, the calculated area product must be higher than the AP_{MIN} calculated in *Equation 35*. If the condition derived from *Equation 38* is not fulfilled, the designer must select a bigger inductor core.

Equation 38

$$AP_{MIN} < AP$$

A simple way how to calculate the number of inductor turns is shown in *Equation 39*, since manufacturers also include the inductance factor $A_{L,}$ which depends on air gap, in the product datasheets. For example, the ETD29 core with 1 mm gap and N27 ferrite material has an inductance factor of 124 nH.

Equation 39

$$L = N^2 \cdot A_L$$

A_L = inductance factor (H)

N = inductor number of turns (-)

The number of turns for the inductor is:

Equation 40

$$N = \sqrt{\frac{L}{A_L}}$$

After the number of turns is calculated, it is also necessary to calculate the wire diameter. The maximum power loss can be calculated using the maximum inductor temperature and core thermal resistance with the following equation:

Equation 41

$$P_{MAX_LOSS} = \frac{T_{MAX} - T_A}{R_T}$$

P_{MAX_LOSS} = maximum power loss in the inductor (W)

T_{MAX} = maximum inductor temperature (°C)

R_T = thermal resistance of the inductor core used (°C/W) (for example, thermal resistance for the E25 core used in the STEVAL-ILL013V1 design is 40 °C/W).

The loss in the core is:

Equation 42

$$P_{CORE} = P_V \cdot W \cdot 10^{-3}$$

P_{CORE} = loss in the inductor core (W)

P_V = core loss defined in the datasheet (mW/g)

W = core weight (g) (for example, the ETD29 has a weight of 28 g)

Maximum power loss in the wire is simply:

Equation 43

$$P_{WIRE} = P_{MAX_LOSS} - P_{CORE}$$

P_{WIRE} = maximum power loss in the wire (W)

The maximum wire resistance derives from the following equation:

Equation 44

$$R_{MAX_WIRE} = \frac{P_{WIRE}}{I_{RMS}^2}$$

R_{MAX_WIRE} = maximum wire resistance (Ω)

Winding resistance depends on the diameter, and is defined using the following formula:

Equation 45

$$R = p \cdot \frac{l}{S} = p \cdot \frac{l_N \cdot N}{\pi \cdot d}$$

R = wire resistance (Ω)

p = resistivity of the copper (Ω • cm) (1,76 • 10^{-6} for temperature 25 °C)

S = conductor cross-section area (cm^2)

l = wire length (cm)

l_N = average length of turn (cm)

d = wire diameter (cm)

The wire diameter is properly selected if the total wire resistance is lower than the maximum wire resistance.

Equation 46

$$R < R_{MAX_WIRE}$$

3 Reference and related materials

1. STEVAL-ILL013V1, 80 W offline PFC and LED driver demonstration board with dimming based on the L6562A; data brief
2. AN1792, Design of fixed-off-time-controlled PFC pre-regulators with the L6562; application note
3. L6562A, Transition-mode PFC controller; datasheet
4. AN1059, Design equations of high-power-factor Flyback converters based on the L6561; application note
5. UM0670, 80 W off-line LED driver with PFC; user manual

Note: The reference and related materials listed above are available on the STMicroelectronics web site at www.st.com.

4 Revision history

Table 1. Document revision history

Date	Revision	Changes
24-Mar-2009	1	Initial release.

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