AN10925

Flyback converter for non-dimmable SSL152x and SSL1623 applications

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Document information

Info	Content
Keywords	SSL1523, SSL152x, SSL1623, LED driver, mains supply, AC/DC conversion, power factor
Abstract	This application note provides guidelines for creating an efficient AC/DC conversion function using the SSL152x and SSL1623 flyback LED driver IC. This document gives a general description of SSL152x and SSL1623 controllers

Revision history

Rev	Date	Description
01	20100923	Initial version

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1. Introduction

This document provides an insight into the operation and application of the mains isolated SSL152x and SSL1623 flyback LED drivers. It is intended for anyone interested in this product and flyback LED drivers.

Remark: This document is also applicable to other Switched Mode Power Supply (SMPS) products as shown in Table 1.

This introductory section describes the contents of this Application note and the purpose of each section. Each section is a self contained piece of information which in most cases can be read independently from the other section(s). Specific references to other sections are included which contribute to an even better comprehension of the subject.

The first part of this application note is background information about flyback converters using a transformer with only one output, and especially about the SSL152x and SSL1623 flyback LED driver itself; see Ref. 1 on page 21. The second part illustrates the SSL152x and SSL1623 flyback LED driver demo board design.

The basic theory and operation of the flyback topology is explained in Application note AN10754; see Ref. 2 on page 21. The SSL152x and SSL1623 flyback LED drivers are also able to operate in a Buck converter configuration. This type of topology is described in AN10876; see Ref. 4 on page 21. More details of the exact operation of flyback or Buck converters can readily be found in electronic reference books.

Section 2 “Functional description” serves as background information about the SSL152x and SSL1623 flyback LED driver IC features in general.

The actual application design is covered in Section 3 “General Step-by-step design procedure” which provides details of how to build the flyback converter.

In Section 3.3 “Short-circuit shutdown latch” a short-circuit shutdown circuit is described. This can be connected to the SSL152x and SSL1623 based circuits, which are supplied by the auxiliary winding.

Table 1. Applicable documents: cross reference

SSL product types	Applicable TEA product types
SSL152x (SSL1522, SSL1523)	TEA152x (TEA1522, TEA1523)
SSL1623	TEA1623

2. Functional description

This section serves as background information. It describes the features and control mechanism of the SSL152x and SSL1623 LED driver. Most features can be identified in Figure 1.

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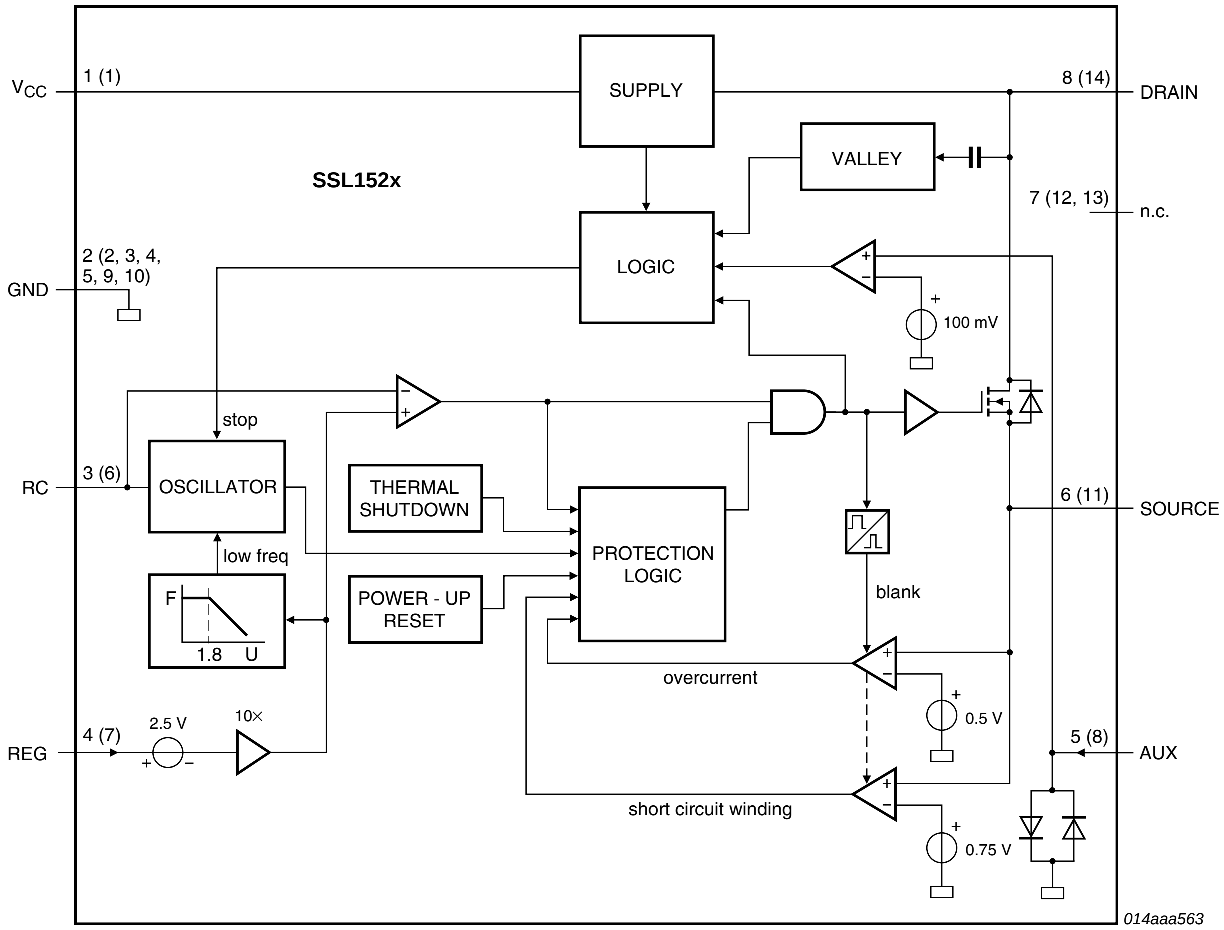

Pin numbers without parenthesis refer to DIP8 packages and within parenthesis refer to SO14 packages.

Fig 1. SSL152x LED driver block diagram

2.1 Start-up and under voltage lockout

Start-up of the SSL152x is initialized by an accurate, high voltage, start-up current source (no dissipative external start-up resistor required). When the voltage on the DRAIN pin is high enough, a start-up current will flow towards the V_{CC} pin. The SSL1623 has no intelligent internal high voltage start-up option so it must be externally supplied at start-up and during operation. The SSL152x and SSL1623 LED driver starts switching as soon as the voltage on the V_{CC} pin passes the V_{CC} start level.

For high efficiency operation, the supply drawn from the DRAIN pin of the SSL152x IC is stopped as soon as the V_{CC} voltage is high enough, and the auxiliary winding of the transformer takes over.

In the event that the auxiliary supply is insufficient, the internal high voltage supply of the SSL152x also supplies the IC. When the voltage on the V_{CC} pin drops below the $V_{th(UVLO)}$ level, the IC stops switching and restarts from the rectified mains voltage.

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2.2 Power MOS transistor

The SSL152x and SSL1623 LED driver has an on-board power switch which is capable of withstanding 650 V on the drain. The devices are not avalanche rugged so sufficient device breakdown protection measures must be taken. The drain-source on-state resistance (R_{DSon}) of the MOS transistor depends on the type of IC selected.

2.3 Oscillator

The switching frequency (f_{sw}) of the SSL152x and SSL1623 LED driver is set by connection of a capacitor and a resistor in parallel with the RC pin. The capacitor is rapidly charged to the $V_{RC(max)}$ level and, starting from a new primary stroke, it discharges via the resistor to the $V_{RC(min)}$ level. When the $V_{RC(min)}$ level has been reached, the capacitor is recharged. The switching frequency is calculated in Equation 1

$$\frac{1}{f_{sw}} = t_{charge} + R_{osc} \times C_{osc} \times 1n\left(\frac{V_{RC(max)}}{V_{RC(min)}}\right) \tag{1}$$

where:

t_{charge} = charge time

R_{osc} = oscillator resistance

C_{osc} = oscillator capacitance

2.4 Control mechanism

The SSL152x and SSL1623 LED driver uses current mode control. An optocoupler transfers the LED current information to the primary side of the circuit. The conduction time of the internal MOS transistor is modulated. The average output current to the LEDs is controlled by setting the primary peak current through the transformer. The implementation is shown in Figure 2.

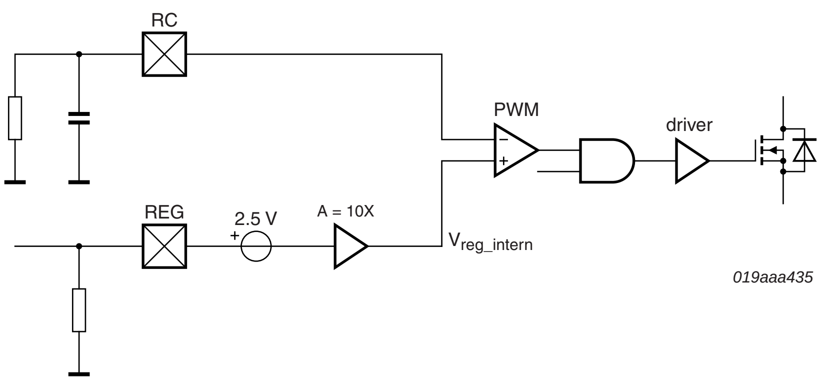

Fig 2. SSL152x and SSL1623 LED driver regulation mechanism

2.4.1 On-time Control

The internal regulation voltage (V_{reg_intern}) is equal to the difference between the external regulation voltage and the internal voltage source (2.5 V) multiplied by a factor of 10. This internal regulation voltage is compared with the oscillator voltage. When the oscillator

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voltage is lower than the internal regulation voltage, the power MOS transistor is turned off. The higher the external regulation voltage, the lower the conduction time of the MOST transistor. The MOST's controlling mechanism is shown in Figure 3.

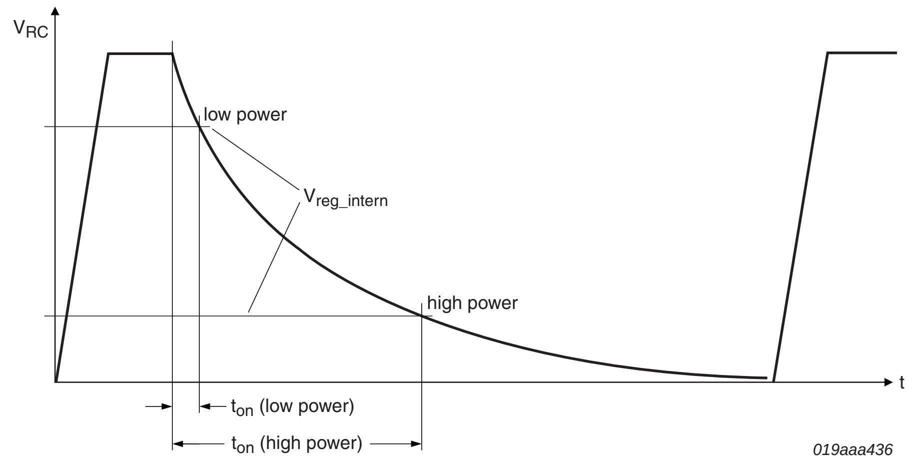

Fig 3. Regulation mechanism

2.4.2 Maximum duty cycle

The power MOS transistor will always be switched off as soon as the maximum duty cycle is reached. The maximum duty factor is set to 75 % (typical value at 100 kHz oscillation frequency).

2.4.3 Minimum duty cycle

The minimum duty cycle is equal to 0 %. This is achieved when the internal regulation voltage is equal to or higher than the maximum oscillator voltage. In this case the power MOS transistor is not switched on.

2.4.4 Advantage exponential oscillator

The use of an exponential oscillator has the advantage that the relative sensitivity of the duty cycle to the regulation voltage at low duty cycles is almost equal to the relative sensitivity at high duty cycles. The result is a more constant gain over the duty cycle range compared to a PWM system with a linear sawtooth oscillator. A small variation in the regulation voltage, see Figure 4 results in a variation of the MOS transistor's conduction time. This variation is smaller at low duty cycle levels than at high duty cycle levels. For a sawtooth oscillator, the variation is equal over the full duty cycle range. The variation in the conduction time of the MOS transistor results in a variation of transferred power. For an exponential oscillator the variation in transferred power at a low duty cycle level is lower with respect to the linear oscillator. This guarantees stable operation at low duty cycle levels.

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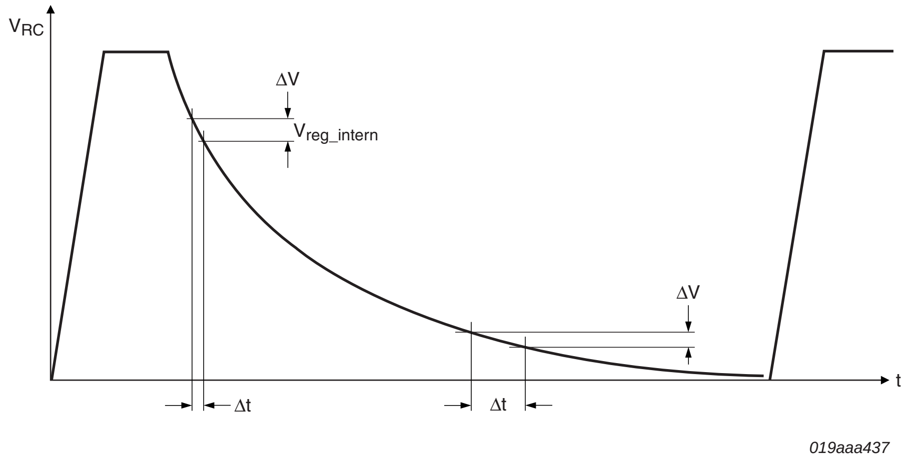

Fig 4. Regulation mechanism

2.5 Demagnetization

The SSL152x and SSL1623 LED driver always operates in Discontinuous Conduction Mode (DCM) or Boundary Conduction Mode (BCM). The auxiliary winding of the transformer is connected to the SSL152x and SSL1623 LED driver AUX pin with a resistor. Current will flow to or from the AUX pin via the two antiparallel diodes. The direction in which the current flows depends on the transformer's auxiliary winding voltage and direction.

When the secondary diode is conducting, the voltage of the auxiliary winding is positive. This supplies a current to the AUX pin. As a result, the AUX pin voltage is clamped to a positive diode voltage. When the AUX pin voltage is higher than 100 mV, the oscillator will not start a new primary stroke.

Demagnetization recognition is suppressed during the t_{suppr} time. This time starts at the moment of switching off the integrated power MOS transistor. This might be necessary in order to prevent false demagnetization detection, especially for applications with low output voltages and transformers with a large leakage induction.

2.6 Valley switching

In order to increase the efficiency of the SSL152x and SSL1623 LED driver, dedicated valley switching circuitry is built in. Minimizing the power MOS transistor's switch-on losses increases the efficiency of the converter; see Figure 5, Figure 6 and Figure 7. When the internal power MOS transistor is switched-on, a new primary stroke is started.

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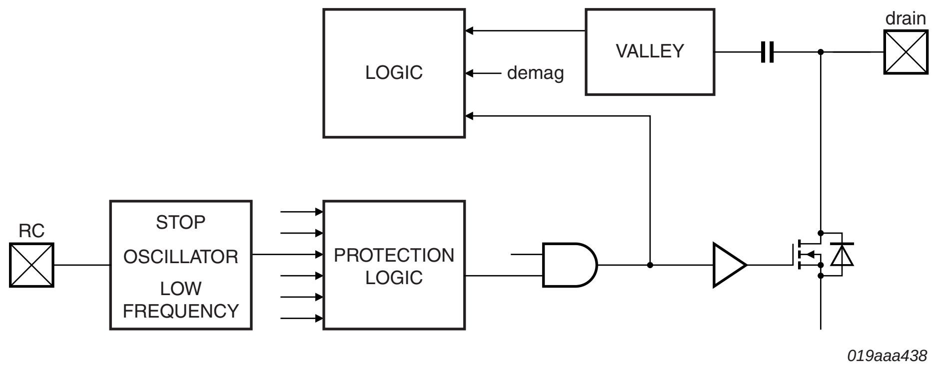

Fig 5. Valley switch circuit

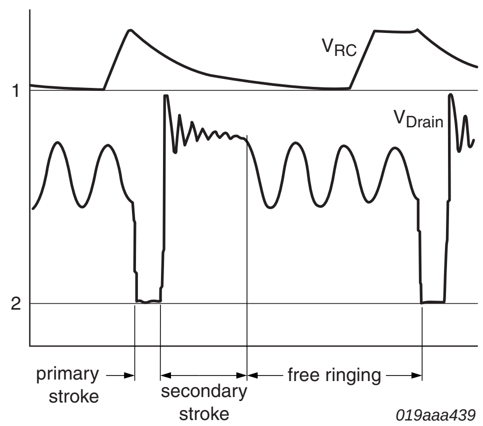

Fig 6. Valley switch graph

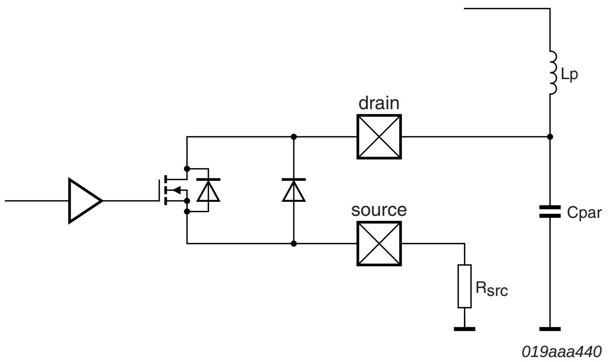

Fig 7. Components at the DRAIN pin

After a period of time determined by the oscillator voltage (V_{RC}) and the internal regulation voltage (V_{reg_intern}), the power switch is turned off, see Section 3.1.4 "Oscillator". At this moment, the secondary stroke is started. The duration of the secondary stroke is determined by the energy stored in the transformer and the output voltage. The SSL152x and SSL1623 LED driver detects the end of the secondary stroke with the

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demagnetization function. Due to the primary transformer's inductance and a parasitic capacitance on the DRAIN pin, the voltage on the DRAIN pin shows an oscillation. The frequency of this oscillation is calculated with Equation 2. As soon as the oscillator is ready ($V_{RC} = V_{(RC(max))}$) and the secondary stroke has ended (V_{aux} < 100 mV), the oscillator waits for a low drain voltage before a new primary stroke is started. The voltage, the value of the parasitic capacitor and the switching frequency determine the switch-on losses; see Equation 3.

$$f_{ring} = \frac{1}{2 \times \pi \times \sqrt{L_p \times C_p}} \quad (2)$$

where:

f_{ring} = ringing frequency

L_p = primary self-inductance

C_p = parasitic capacitance on drain node

$$P_{switch-on} = \frac{1}{2} \times C_p \times V_{DRAIN}{}^2 \times f_{sw} \quad (3)$$

where:

$P_{switch\text{-}on}$ = switch-on losses

V_{DRAIN} = actual voltage on pin DRAIN at moment of switching on

f_{sw} = switching frequency

The power MOS transistor can be switched on just before the actual valley (at low ringing frequencies) or just after it at high ringing frequencies. For a flyback application with a reflected output voltage ($n \times V_o$ = 80 V), a typical curve is shown in Figure 8.

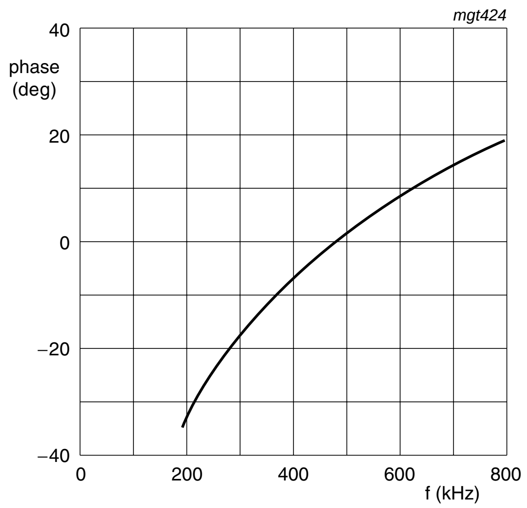

Fig 8. Typical phase of drain ringing at switch-on (at $N \times V_o$ = 80 V)

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Figure 8 shows that for a ringing frequency of 480 kHz, the power MOS transistor is switched on exactly in the valley, thus at the minimum drain voltage. This reduces the switch-on losses to a minimum. At a ringing frequency of 200 kHz the MOS transistor is switched on approximately 33 degrees before the actual valley and the switch-on losses are still significantly reduced. The impact of valley switching on the switching losses is greatest when the excitation voltage that caused the ringing, is half that of the drain voltage.

2.7 Current protections

The current through the internal power MOS transistor is converted into a voltage via the external source resistor and supplied to two comparators. With these two comparators, two types of current protection are implemented in the SSL152x and SSL1623 flyback LED driver. Both protections will be discussed in Section 2.7.1 and Section 2.7.2 as follows:

2.7.1 Overcurrent Protection (OCP)

The voltage on the SOURCE pin is measured cycle by cycle and compared to the $V_{ocp\text{-}max}$ maximum level. The power MOS transistor is switched off as soon the voltage on the SOURCE pin exceeds the $V_{ocp\text{-}max}$ level (typical 0.5 V). To prevent a false OCP detection at switch on of the power MOS transistor, the comparator is disabled during the t_{leb} time (typical 350 ns). When OCP is detected, the actual switching cycle is terminated until the next cycle starts.

2.7.2 Short Winding Protection (SWP)

The voltage on the SOURCE pin is measured and compared to the $V_{swp\text{-}max}$ maximum level. The power MOS transistor is switched off as soon the voltage on the SOURCE pin exceeds the $V_{swp\text{-}max}$ level (typical 0.75 V). To prevent a false SWP detection at switch on of the power MOS transistor, the comparator is disabled during the t_{leb} time (typical 350 ns). When SWP is active, the IC oscillator is disabled until the IC is reset by removing V_{CC}.

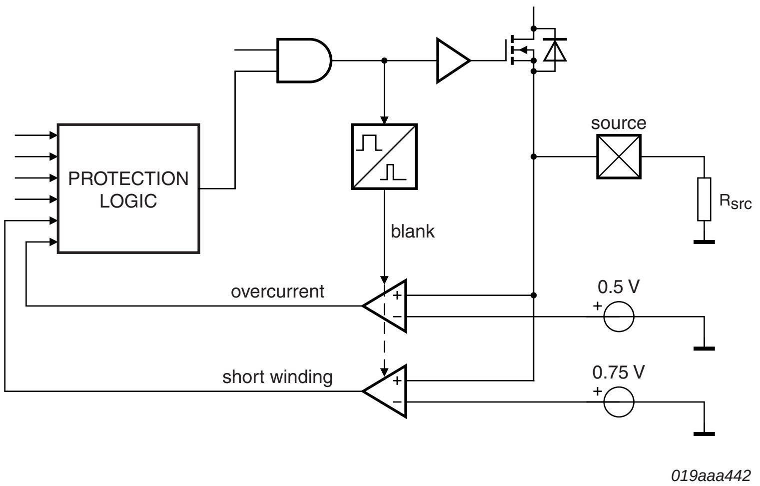

Fig 9. Current protections

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2.8 OverTemperature protection (OTP)

An accurate temperature protection is provided with the SSL152x and SSL1623 flyback LED driver. When the junction temperature exceeds the thermal shutdown temperature, $T_{pl(max)}$, the IC will stop switching and the supply current is reduced to the start-up current level for the SSL152x. As a result, the internal junction temperature will decrease. The SSL152x and SSL1623 flyback LED driver resumes operation as soon as the temperature has sufficiently decreased; $T_{pl(max)} - T_{pl(hys)}$. Should the temperature rise higher than the $T_{pl(max)}$ level again, switching is stopped and the supply current is reduced. Blinking between on and off state therefore occurs

3. General Step-by-step design procedure

This section guides you through the procedure for designing a basic flyback converter with the SSL152x and SSL1623

3.1 Designing the basic SSL152x and SSL1623 application

Figure 10 shows a basic application using the SSL152x and SSL1623 for universal mains and high power factor. This application behaves like a primary regulated voltage source.

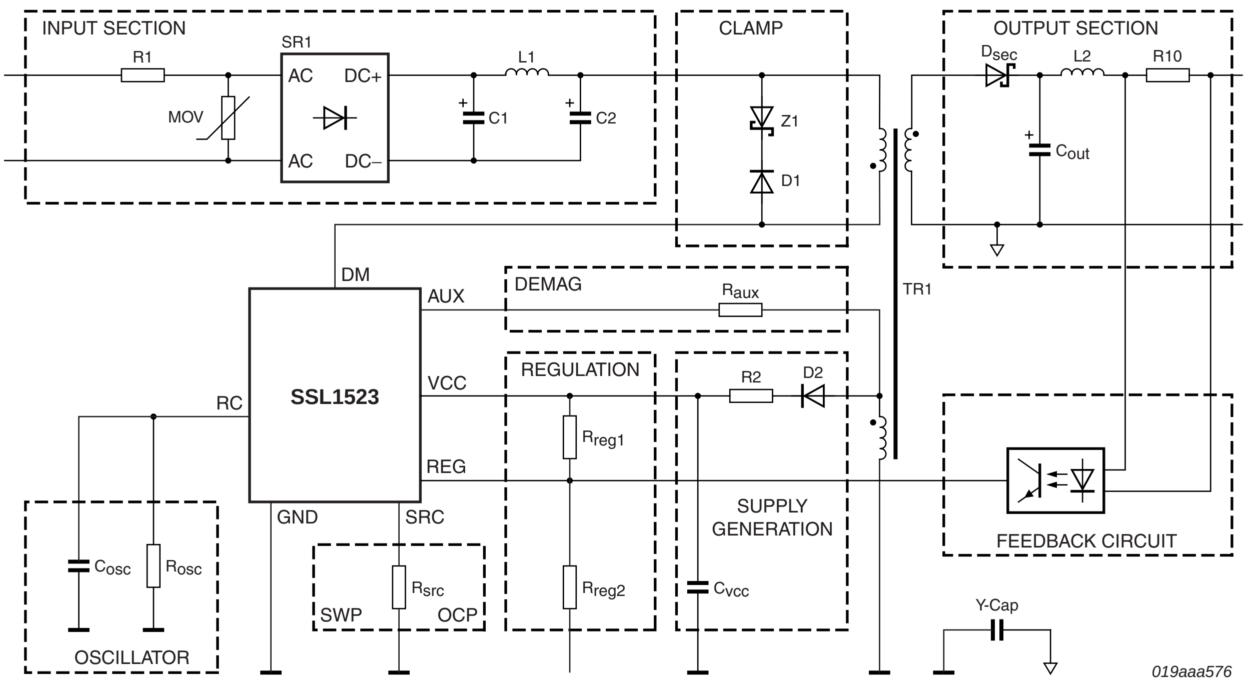

Fig 10. Basic SSL152x and SSL1623 application

The mains voltage is rectified, buffered and filtered in the input section and connected to the primary winding of the transformer. The following blocks can be identified from the SSL152x and SSL1623:

- Oscillator
- OverCurrent Protection (OCP) and ShortWinding Protection (SWP)

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- Regulation
- Demag (demagnetization detection)
- Supply generation
- Feedback circuit

In the output section, the transferred energy is stored in a capacitor and filtered before it becomes available on the output pins. The current feedback circuit is described in Section 3.2 “Feedback circuit dimensioning”

A clamp is added across the primary winding of the transformer to prevent too high a voltage overshoot on the DRAIN pin of the SSL152x and SSL1623, when the internal power MOS transistor is switched off.

3.1.1 Input section

3.1.1.1 Determine system requirements

In order to calculate the input section, the following system parameters must be identified:

- Minimum and maximum AC input voltage, see Table 2

Table 2. Input voltage range

Input voltage (AC)	Minimum	Maximum
universal mains	100 V	254 V

- Mains frequency, see Table 3

Table 3. Input frequency range

Input frequency	Minimum	Maximum
universal mains	45 Hz	65 Hz

- Required output power and voltage
- Estimated efficiency

Efficiency loss due to output voltage

The voltage drop across the output diode affects the efficiency of the whole converter. The increased voltage drop across the output diode and high output current will lower the converter's efficiency.

If the output voltage is below about 25 V and high efficiency is required, a Schottky Barrier diode is recommended or alternatively a fast PN diode can be used.

The efficiency loss due to the output diode is calculated with Equation 4.

$$P_{loss,\,Dout} = \frac{V_{f,\,Dout}}{V_o} \times 100 \times \delta 2 \qquad (4)$$

where:

$P_{loss,\,Dout}$ = output diode power loss

$V_{f,Dout}$ = output diode forward voltage

V_o = output voltage

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PN diode = $V_{f,Dout}$ = 0.7 V.

Schottky diode = $V_{f,Dout}$ = 0.5 V.

Efficiency loss due to snubber/clamp circuit

A snubber network on the DRAIN pin or a clamp circuit across the transformer's primary winding is required in order to keep the drain voltage below the breakdown voltage of the integrated MOS transistor. The estimated efficiency loss due to a snubber or clamp circuit is shown in Table 4

Table 4. Clamp/snubber efficiency loss

Method	Power range	Efficiency loss
RC snubber	$P_o < 3$ W	20 %
RCD clamp	full range	15 %
Zener clamp	full range	10 %

Efficiency loss due to other components

Efficiency loss due to other components in the application is estimated to be approximately 5 %.

Efficiency of the whole converter

The estimated efficiency of the whole converter is calculated with Equation 5.

$$\eta = \frac{P_{in} - P_{loss,\,Dout} - P_{loss,\,clamp} - P_{loss,\,additional}}{P_{in}} \quad (5)$$

where:

P_{in} = input power

$P_{loss,\,Dout}$ = output diode power loss

$P_{loss,\,clamp}$ = clamping power losses

Additional losses ($P_{loss,\,additional}$) are typically: inrush, transformer, switching, IC, wiring etc

Remark:

1. These are estimated values and can differ in a practical application.
2. The peak clamp might not be necessary for the 100 V to 120 V only applications.

3.1.1.2 Calculate the inrush resistor value (R1)

The inrush resistor (R_{inrush}) limits the maximum peak current through the diode bridge rectifier. For most of the diode bridge rectifiers suitable for this application, the I_{FSM} parameter is approximately 20 A. The minimum value for this resistor is calculated with Equation 6.

$$R_{inrush} = \frac{\sqrt{2} \times V_{AC,\,max}}{I_{FSM}} \quad (6)$$

where:

$V_{AC,\,max}$ = maximum AC voltage

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3.1.1.3 Calculate the input buffer capacitance

Before the input buffer capacitance (C_{Buf}) can be calculated, two additional parameters must be defined:

1. The required power factor of the mains input current. The power factor of the mains input current depends on the combination of the value of the output power, the total capacitance of the buffer capacitors, and the inrush resistor. The minimum value is defined in Section 3.1.1.2.
2. The mains input voltage and/or the mains input voltage range

In Table 5 the R_{inrush} and the C_{Buf} multiplicators are given for an application with a power factor of 0.9 of the mains input current.

Table 5. Inrush resistor and buffer capacitance multiplicators

Input voltage range	R_{inrush} ($\Omega \times$ W)	C_{Buf} (nF/W)
110 V	$\geq$ 40	$\leq$ 240
230 VC	$\geq$ 160	$\leq$ 60
universal mains	$\geq$ 160	$\leq$ 60

Taking into account the system efficiency results in the correction factors of Equation 7 and Equation 8:

$$R_{inrush,\,tot} = \frac{\eta}{P_o} \times R_{inrush} \qquad (7)$$

where:

$R_{inrush,\,tot}$ = total inrush resistance

$$C_{Buf,\,tot} = \frac{P_o}{\eta} \times C_{Buf} \qquad (8)$$

where:

$C_{Buf,\,tot}$ = total input buffer capacitance

3.1.2 Input voltage clamp

A MOV or a Transorb at the input of the circuit is required to protect the circuit against lighting surges. The input buffer capacitor is too limited to absorb the energy of lighting surges.

3.1.3 Clamp on DRAIN pin

The maximum clamping voltage ($V_{Clamp,\,max}$) can be determined if Equation 9 is applied. In this equation $V_{(BR)DSS}$ is the breakdown voltage of the SSL152x and SSL1623's integrated power MOS transistor. Since the power MOS transistor is not avalanche rugged, a small safety margin (V_{margin}) is added (25 V is sufficient).

$$V_{Clamp,\,max} = V_{(BR)DSS} - V_{DC,\,max} - V_{margin} \qquad (9)$$

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3.1.4 Oscillator

Before the oscillator components can be calculated, the operating frequency must be chosen. The switching frequency of the SSL152x and SSL1623 can be set between 10 kHz and 200 kHz. A commonly used switching frequency is 100 kHz.
The oscillator frequency is set by two parallel components, a resistor (R_{osc}) and a capacitor (C_{osc}). The capacitor is rapidly charged to the $V_{RC(max)}$ level (typical 2.5 V) and discharged via the resistor to the $V_{RC(min)}$ level (typical 75 mV). The discharge takes 3.5 RC times (RC = oscillator time constant = $R_{osc} \times C_{osc}$).

The oscillator time constant is calculated with Equation 10. The oscillator charge time (t_{charge}) is derived from the SSL152x and SSL1623 specification (t_{charge} = 1μs).

$$RC = \frac{1}{3.5} \times \left(\frac{1}{f_{sw}} - t_{charge}\right) \tag{10}$$

The values for both R_{osc} and C_{osc} can now easily be extracted from the RC time constant. Using an oscillator capacitor less than 220 pF is not recommended as the drain voltage might distort the oscillator voltage. From an efficiency point of view, a high value C_{osc} capacitor is not preferred at high operating frequencies (at 200 kHz and C_{osc} = 10 nF, 12.5 mW of power is dissipated in the oscillator).

For a switching frequency of 100 kHz, an oscillator time constant of 2.57 seconds is required. This time constant is achieved by the parallel connection of a 7.5 kΩ resistor and a 330 pF capacitor.

3.1.5 OCP resistor

The OCP resistor (R_{src}) sets the transformer's primary peak current and thus also the maximum transferred output power.

As the power factor is assumed to be 0.9, the mains current is almost sinusoidal and in phase with the mains voltage.

The OCP resistor defines the dimming curve and regulates the maximum power delivered to the LEDs. It regulates the peak current through the inductor and thus the maximum power level. It also provides overcurrent protection to the converter. This technique removes part of the dependency between the output power and the mains voltage. The built-in overcurrent protection circuit triggers at 0.52 V. If the secondary losses and frequency are known, R_{src} can be calculated with Equation 11 as follows:

$$R7 = \sqrt{\frac{f_{conv} \times L_p}{8 \times P_{in}}} \tag{11}$$

where:

f_{conv} = converter frequency

P_{in} is the transformer input power at the peak of the mains input voltage including snubber losses. The transformer power is modulated by the rectified mains input voltage shape. As the peak transformer input power is twice the average power, twice the average transformer power must be used for P_{in}. The snubber losses must be added to this number.

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Example: For a 5 W application running at a switching frequency of 100 kHz and an efficiency of 80 %, the transformer input peak power will be 5 W × $^{2}/_{0.8}$ = 12.5 W. The snubber losses must be added. The R_{src} resistor is set to 1.2 Ω, limiting the peak current to approximately 400 mA.

3.1.6 Transformer

An SSL152x or SSL1623 flyback application requires a transformer with three windings; a primary input winding (N_p), a secondary output winding (N_s) and an auxiliary winding (N_a). The number of turns will be calculated for all three windings.

For transformer calculations see Application note: Ref. 2 "AN10754". The energy transferred through the transformer has a sinusoidal shape, so the peak power is twice the average power.

As result, the figures for P_{in} in Ref. 2 "AN10754" must be multiplied by 2.

3.1.7 Regulation components

Easy interfacing with secondary regulation is possible. Additional secondary electronics drives the photodiode of an optocoupler. In this case, the optocoupler's transistor is directly connected to input pin REG of the SSL1523 IC.
To prevent distortion on the regulator pin due to in coupling of high voltage signals it is recommended to keep the lower regulator resistor (R_{reg2}) below 10 kΩ.

3.1.8 Demag

The auxiliary resistor (R_{aux}) limits the current to the AUX pin of the SSL152x and SSL16223. According to the specification, the maximum current into or out of the AUX pin is respectively 5 mA and 10 mA. These values are far beyond the current that is necessary for detecting demagnetization. A good approximation for the resistance value for R_{aux} is given in Equation 12.

$$R_{aux} \approx 7 \times V_{aux}(k\Omega) \tag{12}$$

where:

V_{aux} = auxiliary voltage

3.1.9 Supply generation

Due to the fact that the integrated start-up current source (SSL152x only) is switched off when the auxiliary winding provides enough energy to supply the IC, only a small supply capacitor (C_{Vcc}) of maximum 1 µF is required (470 nF will fit practically all applications). The SSL1623 requires an external startup resistor.

The diode which connects the supply to the auxiliary winding is of the General Purpose PN type. The required breakdown voltage of this diode ($V_{br,\,Daux}$) is calculated with Equation 13.

$$V_{br,\,Daux} = \frac{N_a}{N_p} \times V_{dc,\,max} + V_{aux} \tag{13}$$

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The transformer parameters N_a and N_p are determined in Ref. 2 “AN10754”. A resistor is placed in series with the input buffer capacitor. The function of this resistor is to prevent overcurrent at startup. The value for this resistor is given in Table 5 of Section 3.1.1.3

3.1.10 Output section

3.1.10.1 Output diode

The output section starts with the output diode. The type of diode required (PN or Schottky) is decided in Section 3.1.10 (4.1.1). Equation 14 can be used to determine the maximum peak current ($I_{pk, Dsec}$) for the diode.

$$I_{pk,Dsec} = \frac{N_p}{N_s} \times I_{pk,prim} \tag{14}$$

where:

$I_{pk, prim}$ = primary peak current

Further detail can be found in Ref. 2 “AN10754”, *Section 6.2*.

3.1.10.2 Output capacitor

The converter is designed for a power factor of the mains input current of 0.9. As result the input capacitor has a relative low value, so the buffered rectified mains voltage is almost zero at the zero crossings of the mains voltage. As consequence the converter cannot generate a constant output current. A high mains-current related ripple current will be generated at the output of the flyback converter. In order to prevent this low frequency ripple current of 100 Hz to 120 Hz flowing through the LED load an output capacitor must be applied.

The estimated value of the output capacitor (C_{output}) is given in Equation 15 based on a ripple current of 25 %: Further detail can be found in *Section 6.2, AN10754*.

$$C_{output}(\mu F) = 36000 \times \frac{I_{LED}(A)}{n_{LEDs}} \tag{15}$$

where:

n_{LEDs} = number of series connected LEDs

Example: for six series connected LEDs at a current of 0.25 A, an output capacitor of 1500 μF is required.

When using an electrolytic capacitor for the output capacitor, it is recommended that a low ESR ceramic or foil capacitor be connected in parallel to improve EMC filtering and reduce dissipation. This capacitor should be mounted closer to the output diode than the electrolytic output capacitor.

To fulfill the lifetime requirements it is recommended to use a capacitor developed for high temperatures such as 105 °C or 130 °C.

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3.1.10.3 Output filter

The cut-off frequency of the output filter must be set to a frequency below the minimum operating frequency. The minimum operating frequency of the SSL152x and SSL1623 application can be as low as 0 Hz, but this is not a practical value. With Equation 16 and Equation 17, an output filter section can be calculated which has a resonance frequency of $^1\!/_{20}$ th of the switching frequency of 100 kHz.

$$LC = \frac{100}{(\pi \times f_{sw})^2} \tag{16}$$

$$L_{filter} = \frac{100}{C_{filter} \times \pi^2 \times f_{sw}^{\ 2}} \tag{17}$$

3.2 Feedback circuit dimensioning

The developed demo board is designed with a current feedback circuit.

The actual LED current is measured by a sense resistor R10 and a current mirror (R11, R12, Q10, R14 and R15); see Figure 11.

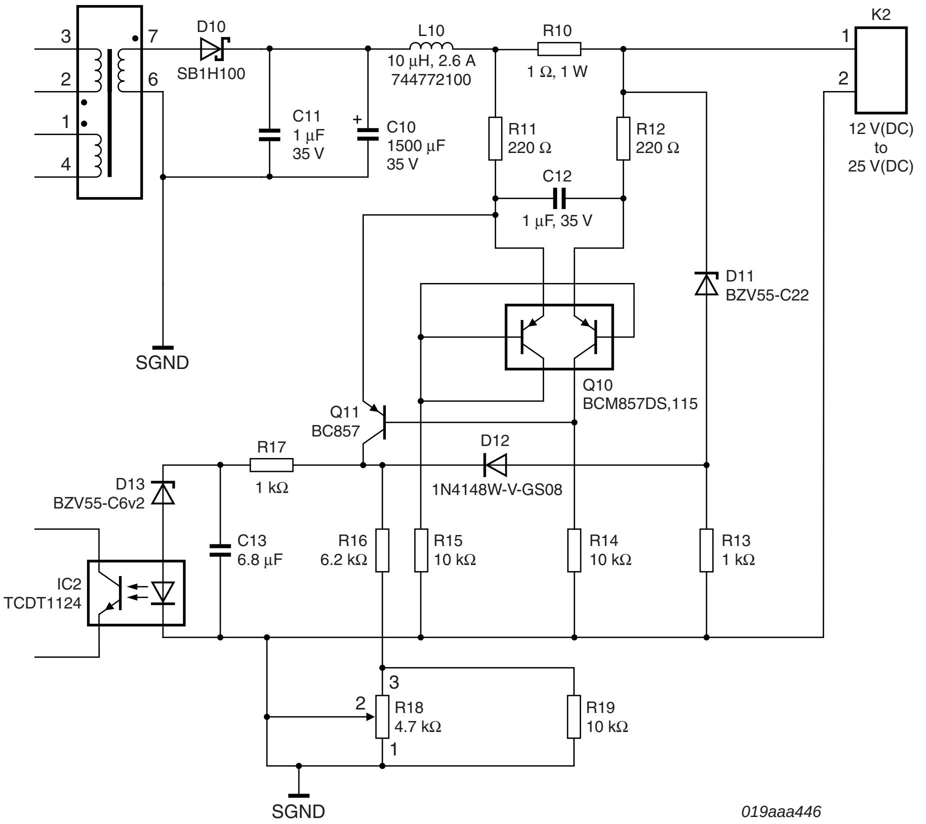

Fig 11. Schematic of output current measurement

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The reflected current is measured as voltage across resistors R16, R18, R19. When this voltage exceeds the Zener voltage of D13 and the forward voltage of the diode in IC2, the optocoupler will be activated. The setting of the SSL152x IC (IC1) on the REG pin will be increased in order to stabilize the output current at the desired value. The current through R16 is defined by the ratio between R10, R11 and the output current; see Equation 18

$$I_{R16} = I_{R10} \times \frac{R10}{R11} \tag{18}$$

Example:

With an output current of 250 mA, the current through R16 will be approximately 1.1 mA, generating a voltage of approximately 9 V across R16, R18 and R19. This voltage is higher than the Zener voltage of D13 added by the forward voltage of the diode in IC2, resulting in a current to flow through R17, D13 and IC1.

3.3 Short-circuit shutdown latch

The developed demo board is designed with auto-restart functionality. During overload mode of operation the output current is limited to a safe value and if an output short-circuit occurs, the circuit is operating in a Hiccup mode

For some applications the requirements demand a complete switching off of the circuit with a short-circuit or overload condition.

An additional circuit will add this functionality to the SSL152x and SSL1623 based applications. This principle can also be applied to the SSL210x based applications.

3.3.1 Description of the short-circuit shutdown schematic

The schematic diagram of the short-circuit shutdown protection is shown in Figure 12.

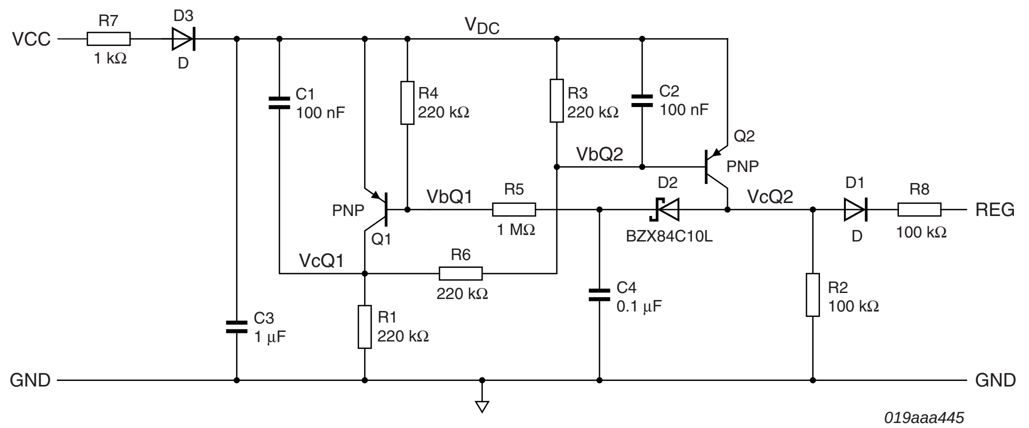

Fig 12. Schematic of the short-circuit shutdown protection

This protection principle can be applied in combination with the SSL LED driver as shown in Ref. 1 "Data sheet SSL152x". During normal conditions the voltage from the auxiliary winding of the converter to the V_{CC} pin of the IC is above 14 V. During overload and/or short-circuit of the LED output connection, the auxiliary voltage will drop and the V_{CC} voltage will drop below 14 V.

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The latching signal is supplied to the REG pin of the SSL152x or SSL1623 IC. During the latched status, the REG pin voltage rises above 2.5 V forcing the SSL152x or SSL1623 IC to the off-state condition.

The circuit is based on a latching configuration with two PNP transistors Q1 and Q2.

The latch is activated when the V_{CC} voltage drops below a predefined value set by the value of D2. A time delay function is added to the latch to prevent false latching and to temporarily disable the latch function during start-up of the SSL152x, or SSL1623 LED driver

During normal operation of the LED driver the V_{CC} voltage is above 14 V. A base current activates transistor Q1, which is in conduction mode. As a consequence transistor Q2 remains in the off-state condition.

During an overload or a short-circuit condition the V_{CC} voltage will decrease below 14 V. The base current, which activated transistor Q1 will reduce to zero, resulting in transistor Q1 switching off. As a consequence transistor Q2 will switch on and the REG pin of SSL152x and SSL1623 IC will increase above 2.6 V, resulting in the LED driver switching off. The internal supply of SSL152x and SSL1623 IC will now be activated and the V_{CC} voltage will stabilize at a lower value of approximately 10 V to 12 V. The transistors Q1 and Q2 remain in the latched condition.

The latch can only be released if the V_{CC} voltage is switched off for at least one second. In order to switch off the V_{CC}, the mains voltage must be removed from the input of the complete driver unit. However the V_{CC} voltage remains active as long as the rectified mains voltage across the input buffer capacitors has a value above 60 V DC.

After discharging the buffer capacitors to a level below 60 V DC, the capacitors C1, C2, C3, and C4 will discharge. The total time for releasing the latch is also highly dependent on the discharge time of the rectified mains buffer capacitors.

4. Abbreviations

Table 6. Abbreviations

Acronym	Description
AC	Alternating Current
DC	Direct Current
BCM	Boundary Conduction Mode
DCM	Discontinuous Conduction Mode
ESR	Equivalent Series Resistance
LED	Light Emitting Diode
MOV	Metal-Oxide Varistor
MOS	Metal-Oxide Semiconductor
MOST	Metal-Oxide Semiconductor Transistor
OCP	OverCurrent Protection
OTP	OverTemperature Protection
PWM	Pulse Width Modulation
SWP	Short Winding Protection

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5. References

[1] **Data sheet SSL152x —** SMPS ICs for mains LED drivers

[2] **AN10754 —** How to design an LED driver using the SSL2101 or SSL2102

[3] **UM10406 —** SSL1523 high power factor 5 W LED driver for universal mains

[4] **AN10876 —** Buck converter for SSL applications

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