A simple cost-effective PSR LED driver without auxiliary winding

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Abstract: A simple cost-effective LED driver based on the primary-side-regulation (PSR) scheme without auxiliary winding is proposed and designed. The driver operates in discontinuous conduction mode (DCM). By means of controlling the discharging and charging of an internal timing capacitor, the conduction time T_{on_s} of the secondary-side current is monitored and the ratio of T_{on_s} to the switching period T_s is kept constant, resulting in a constant output current. The output current zero cross detecting (ZCD) signal is sensed by the negative change of power MOSFET's source voltage due to the resonance of the primary winding with the parasitic capacitor C_{dss} of MOSFET. The driving IC and 650 V power MOSFET chips are packaged together in SOP-8. The proposed driver is implemented with few external components and its size is minimized. Experimental results show that the current variation is within $\pm 3\%$, and the efficiency exceeds 80% at the output power of 3–5 W under universal input voltage.

Keywords: LED driver, primary-side-regulation, cost-effective, without auxiliary winding

Classification: Integrated circuits

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1 Introduction

White light emitting diode (LED) has attracted a lot of attention from both industry and resident for its high efficiency, ease to drive, environmental friendliness, and long lifespan. The LED lamps are being used to replace the incandescent bulbs and fluorescent lamps [1, 2, 3].

The isolated AC-to-DC converters have been widely used among the LED driving circuits due to the safety and simplicity. In order to regulate the output current of the driving circuits precisely, the primary-side-regulation (PSR) and the secondary-side-regulation (SSR) scheme are usually adopted. PSR technique increasingly draws people's attention because its elimination of opto-coupler, which results in lower cost and simpler control circuit. Moreover, the life and temperature issues related to the opto-coupler are also eliminated [2, 3].

The conventional solution for PSR scheme is often to adopt an additional auxiliary winding for sensing the output current. At the time that the secondary winding current reaches zero, the zero cross detecting (ZCD) signal can be sensed by the auxiliary winding. In order to avoid the backflow of auxiliary winding current, the rectifier diode is often required [3, 4, 5, 6].

The efficiency, lifetime and robustness are important for LED drivers. Besides, the low cost, compact volume, and simplicity are the key factors in low power applications. For these reasons, we attempt to develop a simple cost-effective PSR LED driver. In contrast to the conventional PSR LED driver, it does not need an auxiliary winding for sensing the ZCD signal of output current, but it still maintains a good LED current accuracy and a high efficiency under universal input voltage. When the secondary-side current reaches zero, the ZCD signal is sensed by the negative change of the power MOSFET's source voltage, due to the resonance of the primary winding with the parasitic drain-to-source capacitance C_{ds} of MOSFET. We combine the 650 V power MOSFET switch chip with controlling IC chip in SOP-8 to make the converter small. An LED lighting board is implemented with minimized external component count and size. At LED-load power of 3–5 W, its current variation is within $\pm 3\%$ and its power conversion efficiency exceeds 80% under universal input voltage.

2 Design and operation principle

Based on a flyback converter, the converter is composed of bridge rectifier, bulk capacitor C_b , transformer T, bypass capacitor C_{Vcc} , startup resistor R_{st} , snubber circuit, LED driving IC (including built-in power MOSFET), current sense resistor R_{cs} , output rectifier diode D_o and capacitor C_o . The LED driving IC consists of control parts and protection parts. The latter parts include LEDs open/short circuit protection, over-temperature protection etc. The control parts of the driving IC include the charging current circuit I_1 and discharging current circuit I_2 , the main switch control circuit and ZCD signal sense circuit. Fig. 1 shows the block diagram of the circuit implementation along with driving IC.

The proposed LED driver uses a cascade configuration for the power switch. The gate of power MOSFET is held at a DC voltage V_{cc} (14 V), which is obtained by a linear shunt regulator. It employs an internal

Fig. 1. Block diagram of the circuit implementation along with driving IC

bandgap reference V_{ref0} (2.5 V), an error amplifier, a shunt NMOS and the external resistor R_{st} . V_{cc} is also used to drive low voltage driver and the power MOSFET. A DC voltage V_{dd} (5 V) provided by a low dropout voltage regulator is adopted to power other parts.

Fig. 2 shows the timing diagram of relevant signals for the proposed scheme.

Fig. 2. The timing diagram of relevant signals

1) Interval 1: $t_0 \sim t_1$

In this time interval, the power switch is turned on. The primary current i_{Lp} increases linearly (Fig. 2) and the converter stores its energy in its transformer. At the moment when the power transistor begins to turn on, the voltage across the timing capacitor V_{timing} is at a fixed voltage V_c . During interval 1, the analog switch U_1 is turned on and U_2 is turned off. The current I_1 charges the timing capacitor, resulting in the increase of capacitor voltage V_{timing} (Fig. 2). With the increase of primary current, the voltage V_1 across current sense resistor R_{cs} develops.

2) Interval 2: $t_1 \sim t_2$

Once V_1 reaches the setting point $V_{\text{ref}1}$, interval 2 begins. As the output of

comparator changes from low to high level, the flip-flop SR1 will be reset to “0” and the flip-flop SR2 will be triggered to turn off the power transistor. The converter releases its stored energy to its secondary winding, and the magnetizing current i_{L_s} decreases linearly (Fig. 2). Meanwhile, the analog switch U_1 is turned off and U_2 is turned on. The timing capacitor begins to discharge by the current I_2 , resulting in the decrease of the timing capacitor voltage V_{timing} (Fig. 2).

3) Interval 3: $t_2 \sim t_3$

As the secondary winding has run up energy, the MOSFET’s drain and source voltages begin to resonate. Once the source voltage V_{out} of the MOSFET decreases to a certain value, the ZCD sense circuit will generate a logic high output, and the reasons will be described in detail in the next section. Thus the flip-flop SR1 will be changed from “0” to “1”. At same time, the analog switch U_1 becomes ON and U_2 becomes OFF. Then the timing capacitor is charged by the current I_1 . After the voltage of timing capacitor turns bigger than V_c , the power MOSFET turns on, indicating that interval 1 begins again.

The converter operates in discontinuous conduction mode (DCM), and:

$$I_{LED} = I_s \frac{T_{\text{on}_s}}{T_s} = \frac{I_{sp} T_{\text{on}_s}}{2T_s} \quad (1)$$

$$I_{sp} = \frac{N_p}{N_s} I_{pp} \quad (2)$$

$$V_{LED} = L_s \frac{I_{sp}}{T_{\text{on}_s}} = L_p \left(\frac{N_s}{N_p} \right)^2 \frac{I_{sp}}{T_{\text{on}_s}} \quad (3)$$

Where N_p and L_p are the primary turns and magnetizing inductance, while N_s and L_s are the secondary turns and magnetizing inductance of the transformer, respectively. I_s is the average current on the secondary-side during T_{on_s} . I_{LED} and V_{LED} are the output current and voltage, respectively. T_{on_s} is the secondary-side discharge time, T_s is the switching period time. I_{pp} and I_{sp} are the peak currents on the transformer primary-side and the secondary-side respectively.

Because I_1 is designed to be equal to I_2 , T_{on_s} is equal to the sum of T_{on_p} and T_{off} . T_{on_p} is the time that the power MOSFET is ON, T_{off} is the dead time neither primary-side MOSFET nor secondary-side diode is ON.

$$T_s = T_{\text{on}_s} + T_{\text{on}_p} + T_{\text{off}} = 2T_{\text{on}_s} \quad (4)$$

Combining (1), (2) and (4), the output current I_{LED} can be expressed as

$$I_{LED} = \frac{I_{pp} N_p}{4N_s} \quad (5)$$

The peak current I_{pp} is determined by the external current sense resistor R_{cs} and internal V_{ref} . As the peak I_{pp} and turns ratio N_p/N_s are fixed, the output current I_{LED} will keep constant. From (1), (3) and (4), the frequency f can be obtained.

$$f = \frac{1}{T_s} = \frac{1}{2T_{\text{on}_s}} = \frac{1}{8} \left(\frac{N_p}{N_s} \right)^2 \frac{V_{LED}}{L_p I_{LED}} \quad (6)$$

According to (6), the frequency f is developed with the increase of the V_{LED} . When the input voltage V_{in} increases, the frequency will keep unchanged. This is because that the power MOSFET ON time T_{on_p} will decrease, but the dead time T_{off} will increase, and their sum will keep constant according to (4).

For the lowest input voltage, the MOSFET ON time T_{on_p} is the longest. And the voltage across the timing capacitor V_{timing} experiences the maximum value. When the operation condition is with maximum input voltage and minimum load, the dead time T_{off} is maximized. The voltage V_{timing} experiences the minimum value. So the reference voltage V_c is set at 2.4 V to limit the voltage of V_{timing} located between 0 V and V_{dd} (5 V).

3 ZCD sense circuit, operation principle and design consideration

As shown in Fig. 3 (a), ZCD sense circuit consists of a resonant and a blanking circuit. The comparator in the blanking block senses the source voltage V_{out} of power VDMOS across a resistor divider network R_3 and R_4 with a reference voltage V_{ref2} . V_{ref2} is the divider voltage of V_{cc} and equals to $V_{cc} * R_2 / (R_1 + R_2)$. V_{ZCD} is the divider voltage of V_{out} and equals to $V_{out} * R_4 / (R_3 + R_4)$. In order to get the proper DC voltage of V_{out} before resonance, switches K_1 and K_2 are adopted.

In the interval 1 ($t_0 \sim t_1$), K_1 and K_2 are OFF.

At $t = t_1$, the low voltage driver is turned off. Due to the leakage inductance of the primary winding and the parasitic capacitance, the ringing occurs. The ringing may be large enough to be misunderstood as a ZCD signal. To avoid unintended zero crossing detection, a blanking pulse is applied to the NOT gate to prevent the AND gate from turning to HIGH during blanking period. The fixed blanking time T_{blank} is set to be about 2 μ s. At same time, K_1 and K_2 are turned on.

At $t = t_1$, the voltage across the parasitic gate-to-source capacitor C_{gss} of power MOSFET is around V_{cc} . The carriers in the C_{gss} are discharged through R_g and R_6 , as shown in Fig. 3 (b). Since $(R_g + R_6) * C_{gss} \ll T_{blank}$ (2 μ s), the voltage across C_{gss} decreases from V_{cc} (14 V) to 0 V and V_{out} develops from 0 V to V_{cc} (14 V) during blanking period.

After the blanking time T_{blank} , K_1 is ON and K_2 is OFF. C_{gss} is charged by V_{cc} through R_g , R_3 and R_4 , as shown in Fig. 3 (c). V_{out} decreases gradually and experiences a voltage change ΔV_{on_s} , which is given by

$$\Delta V_{on\_s} = \frac{R_5 + R_6}{R_3 + R_4 + R_5 + R_6} V_{cc} (1 - e^{-t/\tau}) \quad (7)$$

Where the time constant $\tau = [R_g + (R_5 + R_6) // (R_3 + R_4)] * C_{gss}$, and $t = T_{on\_s} - T_{blank}$.

At $t = t_2$, $V_{out} = V_{cc} - \Delta V_{on\_s}$. At this time, the divider voltage V_{ZCD} must be higher than V_{ref2} . Otherwise the ZCD sense circuit will output a “1” signal to indicate a ZCD signal, although there is no ZCD signal in fact.

In the interval 3 ($t_2 \sim t_3$), the inductance of the primary winding and the parasitic capacitors begin to resonate. The parasitic capacitor C_{dss} of the power MOSFET is put in use. The equivalent resonant circuit is shown in Fig. 3 (d). As the values of R_5 and R_6 are big enough, the resonant circuit can be simplified, which is shown in Fig. 3 (e). It is a RLC circuit

Fig. 3. (a) ZCD sense circuit, (b) Discharging circuit of C_{gss} during blanking period, (c) Charging circuit of C_{gss} during the remaining time ($T_{on_s} - T_{blank}$) of interval 2 after blanking period, (d) Equivalent resonant circuit, (e) Simplified equivalent resonant circuit, (f) Waveform of V_{out} in different intervals

which is made up of L_p , C_d and R_g . C_d is the total capacitance at the power MOSFET drain node. As C_{dss} and C_{gss} are in a series connection, and $C_{gss} \gg C_{dss}$, resulting in $C_d \approx C_{dss}$.

At the beginning of resonant period ($t = t_2$), the drain voltage of power MOSFET is $(V_{in} + (N_p/N_s) \cdot V_{LED})$. Here V_{in} is the rectified input voltage. A damped oscillation voltage at drain node of power MOSFET can be expressed as

$$\Delta V_{D(t)} = \frac{N_p}{N_s} V_{LED} e^{-\alpha t} \cos(2\pi f_r t) \quad (8)$$

$$\text{Where } \alpha = \frac{R_g}{2L_p}, f_r = \frac{1}{2\pi\sqrt{L_p C_d}}.$$

During resonant period, R_g can be ignored because of its smaller value compared with the impedance of C_{gss} and C_{dss} . So the source voltage V_{out} experiences a voltage change $\Delta V_{off(t)}$ during resonant period which can be simply given by

$$\Delta V_{off(t)} = \frac{C_{dss}}{C_{dss} + C_{gss}} \Delta V_{D(t)} \approx \frac{C_{dss} N_p}{C_{gss} N_s} V_{LED} e^{-\alpha t} \cos(2\pi f_r t) \quad (9)$$

In the negative half-cycle of the damped oscillation, V_{out} can be expressed as

$$V_{out} = V_{cc} - \Delta V_{on-s} - \Delta V_{off(t)} \quad (10)$$

Where $|\Delta V_{off(t)}|$ is the absolute value of $\Delta V_{off(t)}$.

V_{out} decreases to its negative peak, resulting in the decrease of divider voltage V_{ZCD} . As indicated in Fig 3. (a), when V_{ZCD} is equal to V_{ref2} , ZCD sense circuit will output a “1” signal to SET SR1, indicating that the ZCD signal is sensed.

When $\cos(2\pi f_r t) = -1$, the valley of the drain resonance occurs. The drain and source voltages of power MOSFET experience absolute minimum values. It is obvious that the minimum V_{out} (i.e. the maximum $|\Delta V_{off(t)}|$) is affected by C_{dss}/C_{gss} , N_p/N_s , output voltage V_{LED} and the decay factor α . These parameters should be carefully considered to guarantee that the divider voltage V_{ZCD} should fall below V_{ref2} for effective sensing of ZCD for different output voltage V_{LED} under universal input voltage.

Based on the aforesaid analysis, the key parameters and values for ZCD sense circuit are designed and shown in Table I. The divider voltage $V_{ZCD} = 2V_{out}/13$, $V_{ref2} = V_{cc}/7$. When $V_{ZCD} = V_{ref2}$, a ZCD signal is sensed. From Eq. (10), it can be derived that

$$\Delta V_{on-s} + \Delta V_{off(t)} = \frac{V_{cc}}{14} \quad (11)$$

Table I. Key values for ZCD sense circuit ($V_{cc} = 14V$,
 $R = 50k\Omega$, $T_{blank} = 2\mu s$)

Symbol	R_1	R_2	R_3	R_4	R_5	R_6	R_g	V_{ref2}	V_{ZCD}
Value	1800k Ω (36R)	300k Ω (6R)	550k Ω (11R)	100k Ω (2R)	35k Ω	1k Ω	0.5k Ω	$\frac{V_{cc}}{7}$	$\frac{2V_{out}}{13}$

The switching frequency f is set to be 50 kHz, and C_{gss} is about 100 pF for the selected power MOSFET. Thus $T_{on\_s} = 1/(2f) = 10 \mu s$, $t = T_{on\_s} - T_{blank} = 8 \mu s$. From Eq. (7), $\Delta V_{on\_s} = 0.66 V$. And from Eq. (11), $|\Delta V_{off(t)}| = 0.34 V$. It means that when the negative change $\Delta V_{off(t)}$ for the power MOSFET's source voltage during resonant period is $-0.34 V$, ZCD signal is sensed.

4 Experimental results

The driving IC is implemented in CSMC 0.35 μm 5–40 V CMOS process. The key components and parameters of this prototype are as listed in Table II.

Table II. Key components and parameters of the prototype

Parameters	AC input voltage	Output power	Primary-side inductance	Transformer core
Symbol	V_{in}	P_o	L_p	
Value	85-265V/50Hz	3-5W	1.4mH	EE13/PC40/TDK
Parameters	Transformer turns ratio	Startup resistor	Current sense resistor	Bulk capacitor
Symbol	N_p/N_s	R_{st}	R_{cs}	C_b
Value	114:30	860k Ω	1.43 Ω	4.7 μF /450V
Parameters	Bypass capacitor	Output capacitor	Output rectifier diode	
Symbol	C_{Vcc}	C_o	D_o	
Value	1 μF /50V	10 μF /50V	ES1D	

The internal reference voltage V_{refl} is designed to be 0.45 V, and the external current sense resistor is set to be 1.43 Ω . The primary peak current is $I_{pp} = V_{refl}/R_{cs} \approx 315$ mA, and from Eq. (5), the output current $I_{LED} = I_{pp} * N_p / (4N_s) = 299$ mA. Lower V_{refl} leads to a lower sense resistance R_{cs} thus a lower conduction loss for a fixed I_{pp} . However, if the values of V_{refl} and R_{cs} are too small, a minor change of V_{refl} or R_{cs} will result in a significant change for the output current.

Driving IC chip and a 650 V VDMOS chip are packaged together in SOP-8. The wire bond micrograph is shown in Fig. 4 (a). $\Phi 25 \mu m$ Au bondwires are used. The die sizes of driving IC and VDMOS are 1.05 mm $\times$ 1.0 mm and 1.5 mm $\times$ 1.5 mm, respectively. The prototype of the proposed converter in Fig. 4 (b) is designed and implemented in a PCB whose dimension is 27 mm $\times$ 15 mm.

The experimental results are shown in Fig. 5. In Fig. 5 (a), for a fixed output power, the frequency is kept almost unchanged when input voltage is increased from 85 to 265 V. The frequency develops from 42 to 64 kHz when the output power is increased from about 3 to 5 W (i.e. output voltage is increased from 9.6 to 16.0 V). The results agree with the Eq. (6). In Fig. 5 (b), the current variation is within $\pm 3\%$ and the efficiency of the prototype exceeds 80% when the output power ranges from about 3 to 5 W under universal input voltage. Although the system's auxiliary winding and its relevant rectifier diode, divider resistors are eliminated, the current variation and the efficiency are kept at the same level, as compared with other low power LED drivers ([5] and [6]).

Fig. 4. (a) Wire bond micrograph of driving IC and VDMOS, (b) Photograph of the driver prototype

Fig. 5. (a) Measured frequency, (b) LED output current and efficiency verse input voltage and output power

5 Conclusions

We propose a PSR LED driver with auxiliary winding free. The driving IC is realized in $0.35\ \mu\text{m}$ 5–40 V CMOS process and packaged in SOP-8 together with 650 V power VDMOS. The ZCD signal is sensed by the negative change of power MOSFET's source voltage due to the resonance of the primary winding with the parasitic capacitor C_{dss} of MOSFET. And the auxiliary winding, relevant rectifier diode, divider resistors are eliminated. The LED lighting board is implemented with few external components, and its size is minimized. At the LED-load power ranged from about 3 to 5 W, its current variation is within $\pm 3\%$, and the efficiency exceeds 80% under universal voltage (85 to 265 V).