An Energy Channelling LED Driver Technology to Achieve Flicker-Free Operation with True Single Stage Power Factor Correction

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***Abstract*—A single-stage LED driver usually produces a significant twice line frequency ripple current when a high power factor is achieved. The ripple current further produces a flicker through LEDs. In this paper, an energy channeling single-stage LED driver is proposed to achieve flicker-free LED driving while maintaining high power factor correction. It also achieves true single stage power conversion and low component cost. A 8.5W, 50V/0.17A Buck-Boost experimental prototype has been built and tested to verify the proposed method.**

I. INTRODUCTION

A single-stage LED driver can usually achieve a high efficiency and a low component cost. However, the drawback of a single-stage LED driver is significant flicker it produces when driving LEDs. It has been reported in [1] that excessive flicker at twice the line frequency has great adverse effect to human eyes. A two-stage LED driver, even it is much higher on cost and lower on efficiency, becomes necessary for applications that have a strict requirement on flicker.

A lot of new LED driving methods were proposed with the attempts to reduce component cost, flicker and improve efficiency. An active energy storage method was proposed in [2-5]. It can achieve DC LED current driving while reducing output capacitor. However, a great amount power is converted back and forth in the system, which increases power loss and requires a complicated, expensive extra circuitry. A harmonic currents injecting method was proposed in [6-7]. The LED ripple current can be reduced at the cost of reducing a power factor. It is a design trade-off between reducing a ripple current and achieving a high power factor. A two stages combing method was proposed in [8-9]. The power MOSFETs are shared between the first stage PFC and the second stage DC-DC converter. It achieved a reduced component cost. However, it is difficult to achieve an optimized operation for both the first stage and the second stage with shared MOSFETs.

A ripple cancellation method was proposed in [10-12]. A low power converter is used in addition to an original single-stage LED driver. The low power converter produces an opposite twice line frequency ripple voltage to cancel the ripple voltage from the main output. As a result, a DC LED voltage is applied to LEDs to achieve DC LED current driving.

The energy channelling method proposed in this paper also uses ripple cancellation concept to produce a DC LED voltage and a DC LED current. However, it distinguishes itself from the method in [10-12] by not using an additional low power converter. Instead, it only needs an additional low voltage rating MOSFET. Therefore, the component cost can be further reduced. In addition, the proposed LED driver achieves a true single stage power conversion. It combines all the desirable features from both the single-stage and the two-stage LED driver. A 8.5W, 50V/0.17A experimental prototype has been built and tested to validate the proposed method.

The paper is arranged as follows. The detailed operating principle of the proposed energy channelling LED driver is discussed in section II. The critical circuit analysis is discussed in section III. The experimental result is shown in IV and the paper is concluded at section V.

II. OPERATING PRINCIPLE OF THE ENEGY CHANNELLING LED DRIVER

In this section, the operating principle of the proposed LED driver is discussed.

A. Power Stage

Fig. 1 illustrates the concept of the proposed energy channeling single-stage LED driver. Because the proposed LED driver is designed with high power factor, the input power contains a twice line frequency ripple. The input power is split to produce two outputs, V_{o1} and V_{o2}. The key point is controlling the input power splitting so that the sum of V_{o1} and $V_{o2,}$ which is the voltage applied to LEDs, is a DC voltage. The strategy of channeling the input power is as follows. The majority of the input power is channeled to feed

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V_{o1}. Therefore, a twice line frequency ripple voltage is produced on V_{o1}. The remaining small portion of the input power is channeled to feed V_{o2}. This part of power is carefully controlled in every switching cycle to produce an opposite twice line frequency ripple voltage on V_{o2}. This way, the ripple voltages from V_{o1} and V_{o2} are cancelled and a DC LED voltage is obtained to drive LEDs to produce a DC LED current.

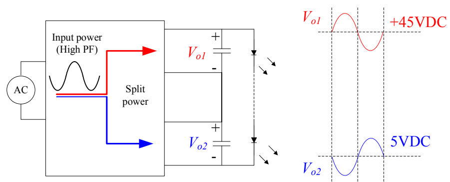


Fig. 1 Concept of the proposed energy channeling LED driver

Fig. 2 shows the power stage implementation of the proposed LED driver. It is based on a two-output Buck-Boost converter with an additional low voltage rating MOSFET Q2. It is by controlling Q2 to achieve precisely control on how much energy is fed to V_{o2} in every switching cycle.

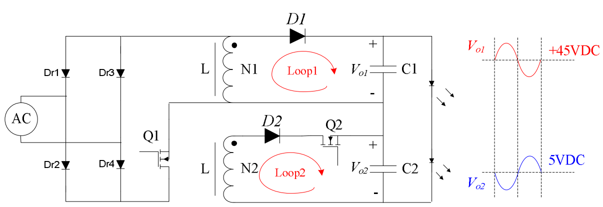


Fig. 2 Buck-Boost implementation of the proposed LED driver

B. The way to achieve energy channeling

Fig. 3 shows the operating principle of the proposed LED driver and an example key waveforms when the proposed LED driver operates at DCM.

Q1 is turned on at the time t_0. The inductor L is charged through winding N1 and both D1 and D2 are off. The state of Q2 doesn't influence the circuit. The inductor current $I_{L(N1)}$ starts increasing from zero. After a fixed turn on time, Q1 is turned off at t_1. Immediately after Q1 is turned off, Q2 remains off. The inductor current keep circulating through D1. The inductor is discharged in $[t_1\text{-}t_2]$ and the inductor energy is transferred to V_{o1}. The MOSFET Q2 is turned on at t_2 and the inductor current commutes from winding N1 to winding N2. The remaining inductor energy only transfers to V_{o2} after t_2. The way to achieve this objective is explained as follows.

In the proposed design, the turns ratio N1:N2 and the voltage ratio V_{o1}:V_{o2} is deliberately mismatched. For example,V_{o1} is regulated at 45V and V_{o2} is regulated at 5V (V_{o1}:V_{o2}=9:1). The turns ratio N1: N2 is made to be 8:2. When Q2 is on, the voltage across winding N2 is 5V and the voltage across winding N1 is 20V. Therefore, D1 is reverse biased and the inductor current only circulates through D2 and Q2 when Q2 is on.

The inductor current continues decreasing and becomes zero at time t3. A new switching cycle doesn't start until t4 to ensure DCM operation.

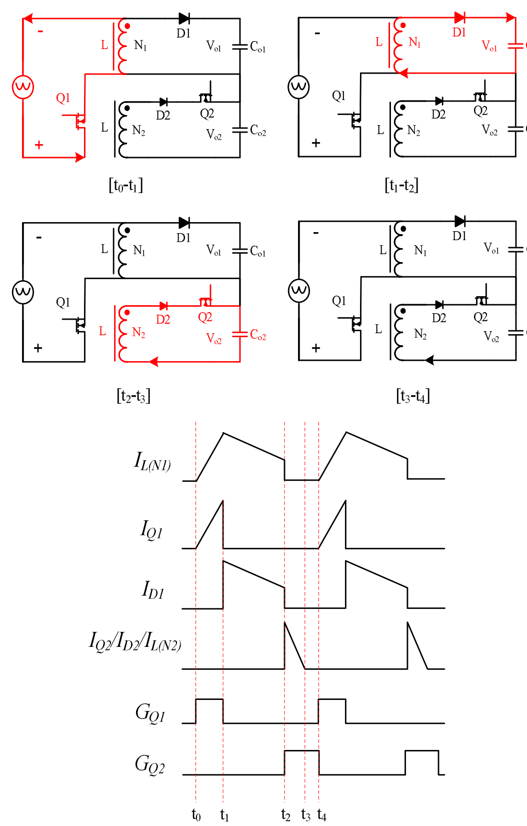


Fig. 3 Operating principle of the proposed LED driver and an example waveforms

Fig. 3 also clearly demonstrated that the proposed LED driver achieves a single stage power conversion. From $[t_0\text{-}t_1]$, the inductor is charged for one time. During [t1-t2], the inductor is discharged and the first portion of its energy is transferred to V_{o1}. During $[t_2\text{-}t_3]$, the inductor is continually discharged and transfers its remaining energy to V_{o2}. In a switching cycle, the inductor only experiences one time charging and one time discharging while transferring the power from the input to the output. Therefore, the proposed LED driver achieves a true single stage power conversion.

C. Control scheme

Fig. 4 shows the control diagram of the proposed LED driver. There are two control loops in the proposed LED driver. One control loop is used to achieve V_{o2} regulation and another control loop is used to achieve LED current regulation.

In order to achieve ripple cancellation, V_{o2} need to produce an opposite twice line frequency ripple voltage to that of V_{o1}. By sensing and reversing the twice line frequency ripple of V_{o1} and adding to a bias voltage, the reference voltage for V_{o2}, V_{o2_ref}, is produced. With a well designed feedback loop, V_{o2} can closely follow its reference voltage and produce an opposite ripple to that of V_{o1}.

V_{ctrl} is the compensation signal in the voltage feedback loop of V_{o2}. V_{ctrl} compares with the sawtooth waveform to produce the gate driving signal G_{Q2} under trailing edge modulation. When V_{o2} is higher than its reference, the compensation signal V_{ctrl} will be higher, the duty cycle of G_{Q2} of will be reduced and results a reduced current flow in V_{o2}, which will gradually bring V_{o2} down to its reference.

In order to achieve LED current regulation, the LED current is sensed and compared with current reference. The result is compensated in the PFC controller and leads to the RMS input current and V_{o1} voltage adjustment. V_{o1} settles to the value that produces an LED current that is exactly equal to the current reference. It should be noted that the DC voltage of V_{o2} is a constant. It is the DC voltage change of V_{o1} that leads to the DC LED current change.

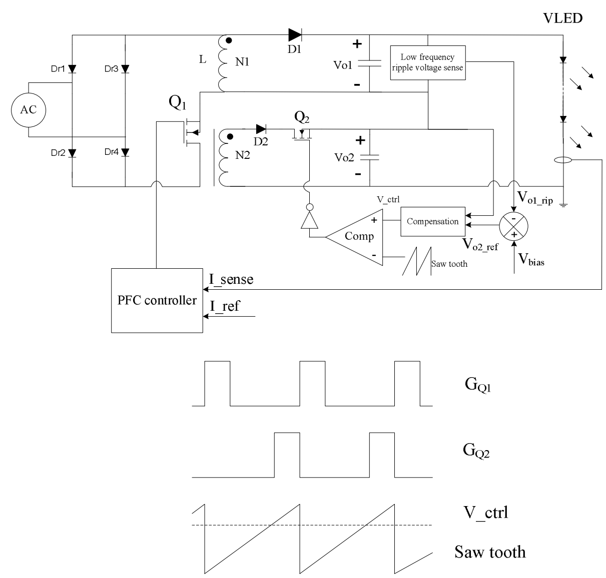


Fig. 4 Control diagram of the proposed LED driver

D. Special operating condition

It was discussed that by controlling the amount of energy feeding V_{o2} in every switching cycle, V_{o2} regulation is achieved. This is true only when there is enough energy to maintain V_{o2} from the AC input source. Fig. 5 shows the waveforms of the input power and the required power to maintain V_{o2} of the experimental prototype.

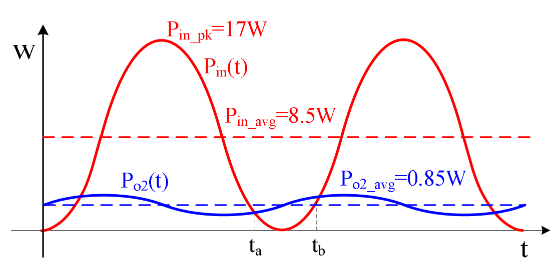


Fig. 5 Input power and output power P_{o2} waveforms

It is demonstrated that during time period [t_a-t_b], the input power P_{in} is lower than the output power P_{o2}. Under this condition, even all the inductor energy is transferred to V_{o2} while zero energy is transferred to V_{o1} in every switching cycle, the inductor energy is still not enough to maintain V_{o2}.

A simple modification is made to the original circuit and is presented in Fig. 6. An auxiliary voltage, V_{aux}, is produced by another winding N_{aux}. D3 is used to separate V_{aux} from the rectified AC input voltage. When V_{aux} is higher than the absolute value of the AC input voltage, D3 will conduct and the input rectifier will be reveres biased. V_{aux} will provide the energy to the output during this interval. A proper sized capacitor C_{aux} is used to keep V_{aux} relative constant. Because the input current follows the input voltage, the input current also becomes constant. Therefore, a constant power can be provided to the output during this time interval.

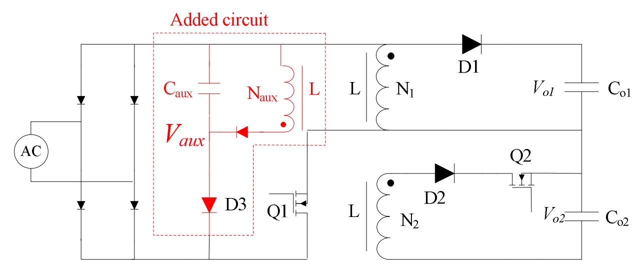


Fig. 6 Modification on the proposed LED driver

Fig. 7 shows the input power waveform with the modified circuit. During time period [t_a-t_b], V_{aux} provides the power to the output and the power is constant and higher than the instantaneous required power to maintain V_{o2}.

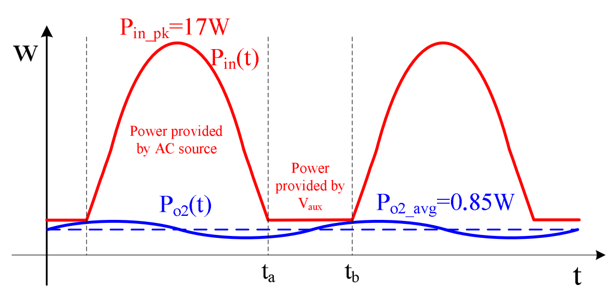


Fig. 7 Input power and output power P_{o2} with modified circuit

III. Critical Circuit Analysis

In this section, the critical circuit parameters of the proposed LED driver are discussed.

A. Q2 Current stress

In order to reduce the current stress for Q2, trailing edge modulation is used to generate the gate driving signal for Q2. As a result, the tail section of the inductor current flows in Q2. In every switching cycle, the peak switching current in Q2, I_{Q2_peak} ,under trailing edge modulation can be expressed as

$$I_{Q2\_peak}(t) = \frac{V_{o2}(t) \times T_{dis\_Q2}}{L_{N2}} \quad (1)$$

In Eq. (1), T_{dis_Q2} and L_{N2} represents the inductor current discharge time after Q2 is turned on and the inductance of the winding N2 respectively. On the other side, the current flowing in Q2 provides the LED current and causes V_{o2} voltage change in a switching cycle. Another relationship can be established as

$$\frac{I_{Q2\_peak}(t) \times T_{dis\_Q2}}{2T_s} = I_{LED} + \frac{\Delta V_{o2}(t) \times C_{o2}}{T_s} \quad (2)$$

In Eq.(2), T_s represents the length of one switching period, C_{o2} represents the output capacitor of V_{o2} and $\Delta V_{o2}(t)$ represents the voltage change on C_{o2} in that switching cycle. The left side of the Eq. (2) represents the average current flowing in Q2 in that switching cycle. The right side represents the LED current plus the current causing voltage change, $\Delta V_{o2}(t)$, on C_{o2}. The last term in Eq. (2) can be ignored since it is much smaller than the LED current, I_{LED}. Combining Eq. (1) and Eq. (2) yields

$$I_{Q2\_peak} = \sqrt{\frac{2 \times V_{o2}(t) \times I_{LED} \times T_s}{L_{N2}}} \quad (3)$$

$$T_{dis\_Q2} = \sqrt{\frac{2 \times L_{N2} \times I_{LED} \times T_s}{V_{o2}(t)}} \quad (4)$$

As shown in Eq. (3), since I_{LED}, T_s, and L_{N2} are fixed values in a design, the maximum peak switching current occurs when V_{o2} is at its maximum. For example, V_{o2} swings to 8V as maximum, I_{LED} is 0.17A, T_s is 50µS and L_{N2} is 50µH. The circulated maximum peak switching current for Q2 is equal to 1.65A and the rms current for Q2 in that switching cycle is 0.43A.

B. *Q2 voltage stress*

When Q1 is on, Q2 does not experience a voltage stress as the reverse voltage is applied on D2. When Q1 and Q2 is off while D1 is conducting, there is a voltage across the drain and the source terminals of Q2. This voltage can be described as

$$V_{ds\_Q2} = V_{o1}(t) \times \frac{N1}{N2} - V_{o2}(t) \quad (5)$$

The maximum voltage cross Q2 occurs when V_{o1} swings to its maximum and V_{o2} swings to its minimum. For example, V_{o1} swings to 48V as maximum and V_{o2} swings to 2V as minimum and the turns ratio N1:N2 is 8:2. The maximum voltage across Q2 is calculated to be 10V.

C. *Requirement on V_{aux}*

As shown in Fig. 5, the input power is less than the output power P_{o2} during time period $[t_a\text{-}t_b]$. At the mean time, V_{o2} swings to the vicinity of the average value of V_{o2}. Therefore, the power required to maintain V_{o2} during $[t_a\text{-}t_b]$ is approximately equal to the averaged output power of V_{o2}, P_{o2_avg}. On the other hand, the instantaneous input power can be express as

$$P_{in}(t) = \left(\frac{V_{in}(t)}{V_{in\_rms}}\right)^2 \times P_{LED} \quad (6)$$

The boundary input voltage can be defined when the input power $P_{in}(t)$ is equal to V_{o2} output power $P_{o2}(t)$. Letting the input power be equal to P_{o2_avg} in Eq. (6), the boundary input voltage can be expressed approximately as

$$V_{boundary} = \sqrt{\frac{P_{o2\_avg} \times (V_{in\_rms})^2}{P_{LED}}} \quad (7)$$

With the original circuit, when the input voltage is below the boundary input voltage, the input power will be lower than the output power P_{o2}. Therefore, with the modified circuit, V_{aux} should be set no less than the boundary input voltage. For example, when P_{o2_avg} is 0.85W, V_{in_rms} is 110V and P_{LED} is 8.5W, the boundary input voltage is calculated to be 35V. Therefore, V_{aux} should be no less than 35V in order to guarantee that the input power is always sufficient to maintain V_{o2}.

On the other hand, V_{aux} should be designed to its required minimum to reduce zero crossing distortion. As discussed, when V_{aux} provides the energy to the output, the input rectifier is reverse biased. The AC input current from the AC source is zero. A higher V_{aux} will result a larger AC current dead time, which increases the zero crossing distortion. Therefore, when the boundary voltage is calculated to be 35V, V_{aux} can be designed to be 40V to provide some margin and minimize zero crossing distortion.

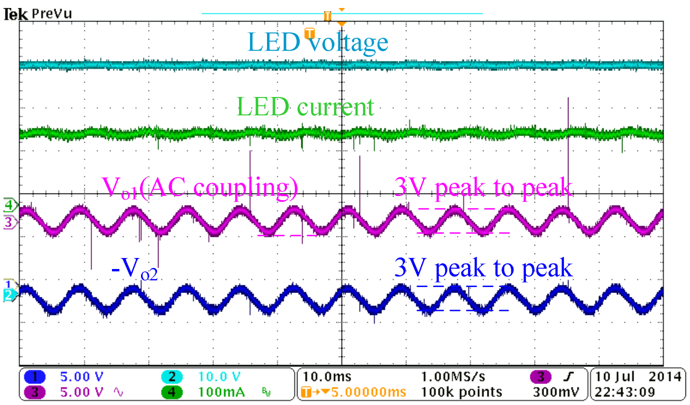
IV. EXPERIMENTAL RESULT

In order to verify the proposed energy channeling LED driving method, a 8.5W, 50V/0.17A Buck-Boost prototype has been built and tested. Table 1 shows the critical circuit parameters of the Buck-Boost prototype.

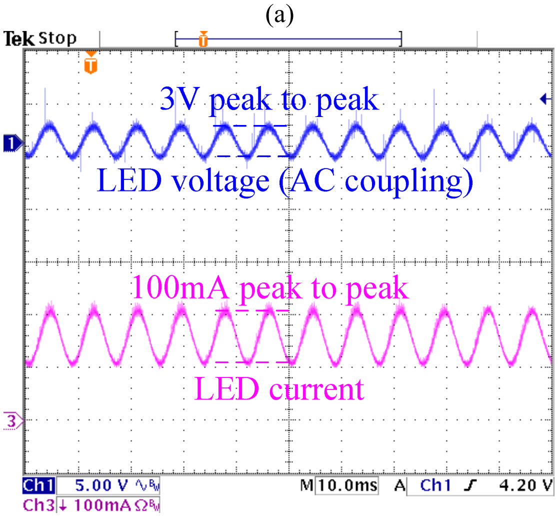
Table 1 Critical circuit parameters of the Buck-Boost Prototype

Input voltage range	Universal
LED voltage	~ 50V
LED current	0.17A
Output V_{o1}	~45VDC±1.5Vrip
Output capacitor (V_{o1})	130µF
Output V_{o2}	~5VDC±1.5Vrip
Output capacitor (V_{o2})	20µF
Switching frequency	20KHz
Winding N1 Inductance	800µH
Winding N2 Inductance	50µH
Main MOSFET Q1	STP3NK80Z
MOSFET Q2	NTD4906N

Fig. 8 shows the performance comparison between the proposed LED driver and a conventional Buck-Boost LED driver. Two LED drivers are connected with the same 130µF output capacitor and the LED load. Fig. 8(a) shows the waveforms of the proposed LED driver. The Ripple voltage cancellation is achieved between the main output V_{o1} and the low output V_{o2}. A DC LED voltage is obtained to applied to the LED load. The twice line frequency ripple LED current is measured to be 8mA peak to peak, which is 2.4% of the averaged LED current. Fig. 8(b) shows the key waveforms of the conventional Buck-Boost LED driver. The twice line frequency ripple LED current is measured to be 100mA, which is 29.5% of the averaged LED current. Therefore, the experimental LED driver prototype achieved a 12.3 times ripple current reduction.



(a)



(b)

Fig. 8 Ripple current comparison between the proposed LED driver and a conventional Buck-Boost LED driver under 50V/0.17A output (a) Proposed Buck-Boost LED driver, (b) Conventional Buck-Boost LED driver

Fig. 9 shows the switching waveforms of the proposed LED driver. When Q1 is turned on, the inductor is charged through winding N1. The inductor current rises from zero. Q2 is not immediately turned on after Q1 is turned off. Trailing edge modulation is used to generate the gate driving signal G_{Q2} for MOSFET Q2. After Q2 is turned on, the inductor current commutes from winding N1 to winding N2. The inductor current keep decreasing until it becomes zero. A new switching cycle doesn't start immediately after the inductor current become zero to ensure DCM operation.

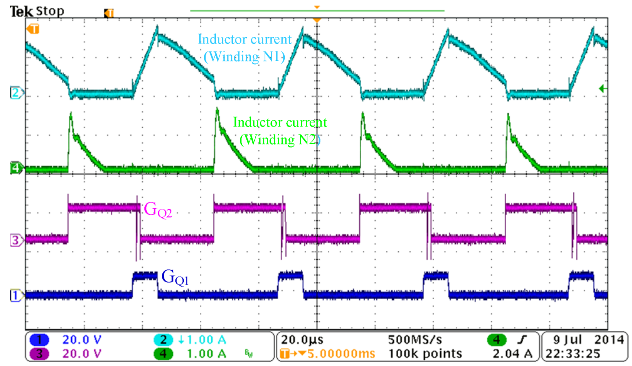


Fig. 9 Critical switching waveforms of the proposed energy channelling LED driver

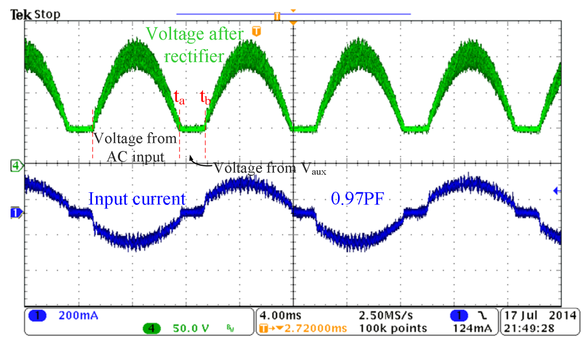


Fig. 10 Input voltage (after rectifier) and input current (before rectifier)

Fig. 10 shows the input voltage (after rectifier) and the AC input current (before rectifier). During time period [t_a-t_b], the voltage V_{aux} provides the energy to the output. Therefore, the after rectifier input voltage is flat during [t_a-t_b]. Because V_{aux} is higher than the AC input voltage, the input rectifier is reverse biased in the time period. As shown in Fig. 10, there is no AC input current during [t_a-t_b]. The power factor is measured to be 0.97 under this input current waveform.

Fig. 11 shows the efficiency of the proposed Buck-Boost experimental prototype. 85% efficiency has been achieved at full load.

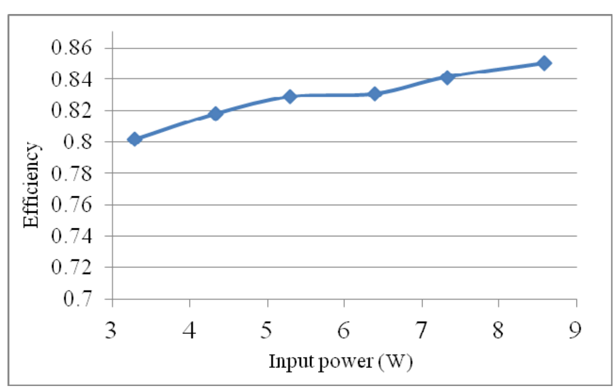


Fig. 11 Efficiency of the proposed LED driver at different load (Vin=110Vrms)

V. CONCLUSION

In this paper, an energy channelling single-stage LED driver has been proposed. The key concept of the proposed LED driver is splitting the input power and producing two output voltages, where each output voltage contains an opposite twice line frequency ripple voltage to another's. A DC LED voltage is obtained to applied to LEDs and achieves DC LED current driving.

The proposed LED driver achieves a true single stage power conversion because the inductor is only charged one time and discharged one time in one switching cycle while transferring the power from the input side to the output side. Trailing edge modulation is used to produce gate driving signal for Q2. As a results, the current stress of Q2 is minimized. An additional auxiliary voltage V_{aux} is produced to solve the problem of no sufficient input power when the input voltage is close to zero. Zero crossing distortion is introduced due to this arrangement. However, a high enough power factor can still be achieved.

A 8.5W, 50V/0.17A Buck-Boost experimental prototype has been built and tested to validate the proposed LED driver. A 0.97 power factor, 85% efficiency at full load and flicker free LED driving performance have been achieved.

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