Article

Development of Four-Channel Buck-Type LED Driver with Automatic Current Sharing

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Abstract: A buck-type light-emitting diode (LED) driver is proposed herein. The proposed LED driver automatically possesses current sharing and high step-down voltage gain. Without complex control, the proposed LED driver, with a single input and multiple outputs, can achieve automatic current sharing of four-channel LED strings, even under the different number of LEDs of each LED string. Furthermore, as compared with the traditional four-phase interleaved buck converter with a single input and a single output having current sharing required, the proposed circuit has the duty cycle up to 0.5, not 0.25, meaning that under the same input voltage the latter has a wider output voltage range than that of the former. Above all, if the proposed circuit with N outputs, then it still has the duty cycle up to 0.5, not one over N as shown traditionally. Moreover, as compared with the current sharing based on the differential-mode transformer, the proposed circuit has no magnetic resetting loop required. In this paper, the operating principles and design considerations of the proposed converter are discussed. Finally, the theoretical analyses and performances of the proposed LED driver are verified by simulation and experiment.

Keywords: capacitor; current sharing; differential-mode transformer; FPGA; LED driver; single input; multiple outputs

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1. Introduction

With the progress of science and technology and economic development, traditional incandescent lamps and power-saving bulbs have been gradually replaced by light-emitting diodes (LEDs), because LEDs have the advantages of small size, long life, high efficiency of electricity, low pollution and rapid reaction, more in line with the needs of today's market [1,2].

In general, LEDs will be connected first in series and then in parallel. This is because if only the series connection is used, the resulting voltage across the LED string will be relatively high, thereby causing the output capacitor to endure a high voltage stress. In addition, if one of the LEDs is burned out, then the whole series of LEDs cannot work. When connected in parallel, LEDs are necessary to have a function of current balance. This is because LEDs have negative temperature coefficient characteristics. If the currents in two LEDs paralleled are imbalanced, one of LED currents will gradually rise, resulting in shortening or even burning out this LED. Therefore, many studies have proposed LED current balance methods so as to make the currents evenly distributed among paralleled LED strings.

The LED current balance is mainly divided into active and passive. The active method [3–7] makes the load current evenly distributed among phases or modules strings. The work in [3] shows a dimming LED driver based on single-input multi-output (SIMO) buck DC-DC converter. The authors in [4] display the current balance for an interleaved LLC resonant converter based on a hybrid rectifier which is utilized to compensate the voltage gain of each phase so as to achieve current balance between the two phases. The study in [5] employs inner loop current sharing and average current processing to balance the current between two modules. The authors in [6] utilize inner current sharing plus a fixed duty cycle not only to balance the currents between the two phases but also to reduce the output ripple. The authors in [7] employ variable inductors to achieve current balance. From the studies in [3–7], it can be seen that more current sensors and controllers are required, thereby making the corresponding circuit more complex and costly. Accordingly, the passive method is presented based on circuit features to achieve the current balance. This method can be classified into two types: differential-mode transformer [8–10] and capacitor [11–15]. Both types generally belong to the SIMO structure, different from single-input single-output (SISO). The former type needs demagnetizing loop, making the required circuit more complicated than the latter type. As compared with active current balance, these two types generally have structure extension and no current balance controller, relatively small size, and low cost. The first type is based on transformer behavior and its operating principle is that when the currents in two LED strings are unbalanced, the differential-mode transformer has the voltages across both sides, and then the differential-mode transformer is activated, thereby forcing the current balance between the two LED strings. The authors in [8] employ the Zeta converter as driving LED strings and recycling magnetizing energy of the differential-mode transformer. The study in [9] couples the two input inductors of the boost converter into the differential-mode transformer to render the current evenly distributed between two LED strings. In [10], both types are used to achieve current balance.

Regarding capacitive current balance, it is based on ampere-second balance, that is, the average current is zero over one period in the steady state. The work in [11] presents an LED driver with galvanic isolation and capacitive current balance. The authors in [12] present an LED with regenerative snubber and capacitive current balance, but the corresponding circuit has no structure extension. The study in [13] displays the non-isolated resonant LED driver with capacitive current balance, but the number of resonant components is increased if the number of outputs increased. The authors in [14] use resonant current balance module to realize capacitive current balance as well as zero-voltage-switching (ZVS) turn-on, but the number of resonant components and switches is increased if the number of outputs increased. The study in [15] utilizes the inherent features of the half-bridge series resonant converter to achieve capacitive current balance and dimming, but the current sharing error percentage is large. In addition, the experimental step-up voltage gains used in the studies [11–13] and the experimental step-down voltage gains used in [14,15] are not so good.

On the other hand, ref. [16] presents the interleaved buck LED driver with automatic capacitive current balance and a suitable duty cycle needed to improve the step-down voltage gain. However, in this circuit, the input and output grounds are separated by the switches, thus restricting converter applications as well as complicating control design. Moreover, the phase counts cannot be changed. Consequently, [17] presents a circuit to conquer the demerits coming from the work in [16].

The authors in [17] propose a four-phase interleaved buck-type converter having high step-down voltage gain, as shown in Figure 1. Furthermore, such a circuit has automatic current balance under the condition that the maximum duty cycle is limited to 0.25. The corresponding voltage gain in the continuous conduction mode (CCM) is 0.25. In [17], the four switches have the same duty cycle with phases of 0° , 90° , 180° and 270° , respectively. The duty cycle is up to 0.25 if the current sharing is required. In the proposed circuit, the four switches also have the same duty cycle with phases of 0° , 180° , 0° and 180° ,

respectively, and the duty cycle is up to 0.5 if the current sharing is required. Accordingly, under the same input voltage the latter has a wider output voltage range than that of the former. Above all, under automatic current sharing, the maximum duty cycle of the latter is relatively large and not changed for structure extension whereas the maximum duty cycle of the former will be reduced as the number of outputs is increased. In addition, the former is of SISO and the latter is of SIMO.

Figure 1. Circuit proposed by the study in [17].

In the following, there are four sections left. The first section will describe the proposed LED driver. The second section will talk about design considerations. The third section will give experimental results. The final section will make a conclusion.

2. Proposed LED Driver

The proposed LED driver with automatic current balance is displayed in Figure 2, derived from the circuit shown in [17], and constructed by four switches S_1 , S_2 , S_3 and S_4 , four diodes D_1 , D_2 , D_3 and D_4 , three energy-transferring capacitors C_1 , C_2 and C_3 , four output inductors L_1 , L_2 , L_3 and L_4 , and four output capacitors C_{o1} , C_{o2} , C_{o3} and C_{o4} . As for the load, it is built up by four LED strings LS_1 , LS_2 , LS_3 and LS_4 . The four output inductors of the proposed LED driver, different from the circuit shown in [17], are connected to individual output capacitors and LED strings. Speaking lucidly, the proposed circuit is of SIMO, whereas the circuit shown in [17] is of SISO.

Figure 2. LED driver with automatic current sharing.

2.1. Converter Operating in CCM

Prior to proceeding with the circuit analysis, some related symbol definitions and required assumptions will be made as follows.

(1) The values of the energy-transferring capacitors C_1 , C_2 , C_3 , and the output capacitors C_{o1} , C_{o2} , C_{o3} and C_{o4} are large enough, so the voltages across them are regarded as fixed values.

(2) V_{in} is the input voltage, and V_{o1} , V_{o2} , V_{o3} and V_{o4} are the voltages across LS_1 , LS_2 , LS_3 , and LS_4 , respectively and they are identical, assuming that $V_{o1} = V_{o2} = V_{o3} = V_{o4} = V_o$.

(3) It is assumed that the values of the four inductors are equal, that is, $L_1 = L_2 = L_3 = L_4 = L$.

(4) i_{L1} , i_{L2} , i_{L3} and i_{L4} are the currents flowing through L_1 , L_2 , L_3 and L_4 , respectively, and i_{C1} , i_{C2} and i_{C3} are the currents flowing through C_1 , C_2 and C_3 , respectively.

(5) v_{L1} , v_{L2} , v_{L3} and v_{L4} are the voltages across L_1 , L_2 , L_3 and L_4 , respectively, v_{ds1} , v_{ds2} , v_{ds3} and v_{ds4} are the voltages across S_1 , S_2 , S_3 and S_4 , respectively, v_{D1} , v_{D2} , v_{D3} and v_{D4} are the voltages across D_1 , D_2 , D_3 and D_4 , respectively, and V_{C1} , V_{C2} and V_{C3} are the voltages across C_1 , C_2 and C_3 , respectively.

(6) T_s is the switching period and f_s is the switching frequency, where $T_s \times f_s = 1$.

(7) D_1T_s , D_2T_s , D_3T_s and D_4T_s are the conduction times of the switches S_1 , S_2 , S_3 and S_4 , respectively. Assume that the four duty cycles are identical, that is, $D_1 = D_2 = D_3 = D_4 = D$. Since the four-phase currents must flow evenly, the duty cycle D must be less than 0.5.

(8) All the switches, diodes, inductors and capacitors in the circuit are regarded as ideal components, but the body diode connected to the switch is still considered. That is, the turn-on resistance of the switch, the forward voltage and parasitic resistance of the diode, and the parasitic resistance of the capacitor are ignored.

(9) v_{gs1} , v_{gs2} , v_{gs3} and v_{gs4} are the gate driving signals for the four switches S_1 , S_2 , S_3 and S_4 , respectively.

(10) $v_{dsi,statey}$ means the voltage across the i -th switch during the y -th state, whereas $v_{Di,statey}$ means the voltage across the i -th diode during the y -th state.

(11) The circuit operates in the continuous conduction mode (CCM). There are four operating states over one switching cycle, as shown in Figure 3.

Figure 3. Waveforms relevant to the converter operating in CCM.

State 1 [$t_0 \leq t \leq t_1$]: As shown in Figure 4, the switches S_1 and S_3 are turned on but the switches S_2 and S_4 are turned off, whereas the diodes D_1 and D_3 are turned off but D_2 and D_4 are turned on. There are two loops in this state. The input voltage V_{in} , the energy-transferring capacitor C_1 , the inductor L_1 , and the LED string LS_1 form one loop, so that the energy-transferring capacitor C_1 is charged while the inductor L_1 is magnetized. The energy-transferring capacitors C_2 and C_3 , the inductor L_3 and the LED string LS_3 form the other loop, so that the energy-transferring capacitor C_3 is charged while the inductor L_3 is magnetized. The inductors L_2 and L_4 are demagnetized via the diodes D_2 and D_4 , respectively. In this state, the voltages on the non-conducting diodes and switches are:

$$v_{D1,state1} = V_{in} - V_{C1} \quad (1)$$

$$v_{D3,state1} = V_{C2} - V_{C3} \quad (2)$$

$$v_{ds,state1} = V_{in} - V_{C2} \quad (3)$$

$$v_{ds4,state1} = V_{C3} \quad (4)$$

Figure 4. Current flow in state 1 in CCM.

States 2 and 4 [$t_1 \leq t \leq t_2, t_3 \leq t \leq t_4$]: As shown in Figure 5, the switches S_1, S_2, S_3 and S_4 are turned off, whereas the diodes D_1, D_2, D_3 and D_4 are turned on. At this time, the inductors L_1, L_2, L_3 and L_4 are demagnetized via the diodes D_1, D_2, D_3 and D_4 , respectively. In this state, the voltages across the non-conducting switches are:

$$v_{ds1,state2} = V_{in} - V_{C1} \quad (5)$$

$$v_{ds2,state2} = V_{C1} - V_{C2} \quad (6)$$

$$v_{ds3,state2} = V_{C2} - V_{C3} \quad (7)$$

$$v_{ds4,state2} = V_{C3} \quad (8)$$

Figure 5. Current flow in states 2 and 4 in CCM.

State 3 [$t_2 \leq t \leq t_3$]: As shown in Figure 6, the switches S_2 and S_4 are turned on but S_1 and S_3 are turned off, whereas the diodes D_2 and D_4 are turned off but D_1 and D_3 are turned on. The energy-transferring capacitors C_1 and C_2 , the inductor L_2 , and the LED string LS_2 form a loop, so that the energy-transferring capacitor C_1 releases energy to the energy-transferring capacitor C_2 , the inductor L_2 , and the LED string LS_2 . In addition, the

energy-transferring capacitor C_3 , the inductor L_4 and the LED string LS_4 also form a loop, so that the energy-transferring capacitor C_3 releases energy to the inductor L_4 and the LED string LS_4 . At this time, the inductors L_1 and L_3 are demagnetized via diodes D_1 and D_3 , respectively. In this state, the voltages across the non-conducting switches and diodes are:

$$v_{D2,state3} = V_{C1} - V_{C2} \quad (9)$$

$$v_{D4,state3} = V_{C3} \quad (10)$$

$$v_{ds1,state3} = V_{in} - V_{C1} \quad (11)$$

$$v_{ds3,state3} = V_{C1} - V_{C3} \quad (12)$$

Figure 6. Current flow in state 3 in CCM.

2.2. Voltage Conversion Ratio M in CCM and Voltages on C_1 , C_2 and C_3

By applying the volt-second balance to the inductors L_1 , L_2 , L_3 and L_4 , the following equations can be obtained:

$$D(V_{in} - V_{C1} - V_{o1}) + (1 - D)(-V_{o1}) = 0 \quad (13)$$

$$D(V_{C1} - V_{C2} - V_{o2}) + (1 - D)(-V_{o2}) = 0 \quad (14)$$

$$D(V_{C2} - V_{C3} - V_{o3}) + (1 - D)(-V_{o3}) = 0 \quad (15)$$

$$D(V_{C3} - V_{o4}) + (1 - D)(-V_{o4}) = 0 \quad (16)$$

In the case of exactly the same load, V_{o1} , V_{o2} , V_{o3} , and V_{o4} in the above equations can be regarded as fixed values of V_o . Then, the voltage conversion ratio M and the voltages across the energy-transferring capacitors, called V_{C1} , V_{C2} and V_{C3} , can be obtained by rearranging the above equations:

$$M = \frac{V_o}{V_{in}} = \frac{D}{4} \quad (17)$$

$$V_{C1} = 3 \times \frac{V_o}{D} = \frac{3}{4} V_{in} \quad (18)$$

$$V_{C2} = 2 \times \frac{V_o}{D} = \frac{V_{in}}{2} \quad (19)$$

$$V_{C3} = \frac{V_o}{D} = \frac{V_{in}}{4} \quad (20)$$

2.3. Converter Operating in Discontinuous Conduction Mode (DCM)

Prior to the circuit operation analysis, some related symbol definitions and required assumptions in DCM are the same as those in CCM. The relevant waveforms in the circuit are shown in Figure 7. Furthermore, it is assumed that the demagnetization time of the four inductors in the circuit is the same, and its value is $D_2 T_s$ and the four output voltages are clamped by the four LED strings so $V_{o1} = V_{o2} = V_{o3} = V_{o4} = V_o$. In addition, $v_{dsi,statey}$

means the voltage across the i -th switch during the y -th state, whereas $v_{Di,statey}$ means the voltage across the i -th diode during the y -th state.

Figure 7. Waveforms relevant to the converter operating in DCM.

State 1 [$t_0 \leq t \leq t_1$]: The circuit operation in state 1 under DCM and the circuit operation in state 1 under CCM are the same, and this will not be redescrbed herein.

State 2 [$t_1 \leq t \leq t_2$]: The circuit operation in state 2 under DCM and the circuit operation in state 2 under CCM are the same, and this will not be redescrbed herein.

State 3 [$t_2 \leq t \leq t_3$]: As shown in Figure 8, the difference from state 2 is that the inductors L_2 and L_4 have been demagnetized entirely at this time and the currents in these

two inductors are both zero. In this state, the voltages across the non-conducting switches and diode are:

$$v_{D2,state3} = v_{D4,state3} = V_o \quad (21)$$

$$v_{ds1,state3} = V_{in} - V_{C1} \quad (22)$$

$$v_{ds2,state3} = V_{C1} - V_{C2} - V_o \quad (23)$$

$$v_{ds3,state3} = V_{C2} - V_{C3} - V_o \quad (24)$$

$$v_{ds4,state3} = V_{C3} - V_o \quad (25)$$

Figure 8. Current flow in state 3 in DCM.

State 4 $[t_3 \leq t \leq t_4]$: The circuit operation in state 4 under DCM and the circuit operation in state 3 under CCM are the same, and this will not be redescribed herein.

State 5 $[t_4 \leq t \leq t_5]$: The circuit operation in state 5 under DCM and the circuit operation in state 4 under CCM are the same, and this will not be redescribed herein.

State 6 $[t_5 \leq t \leq t_0 + T_s]$: As shown in Figure 9, the difference from state 2 is that the inductors L_1 and L_3 have been demagnetized entirely at this time, and the currents in these two inductors are both zero. In this state, the voltages across the non-conducting diode and switches are:

$$v_{D1,state6} = v_{D3,state6} = V_o \quad (26)$$

$$v_{ds1,state6} = V_{in} - V_{C1} - V_o \quad (27)$$

$$v_{ds2,state6} = V_{C1} - V_{C2} - V_o \quad (28)$$

$$v_{ds3,state6} = V_{C2} - V_{C3} - V_o \quad (29)$$

$$v_{ds4,state6} = V_{C3} - V_o \quad (30)$$

Figure 9. Current flow in state 6 in DCM.

2.4. Voltage Conversion Ratio M in DCM and Voltages on C_1 , C_2 and C_3

By applying the volt-second balance to the inductors L_1 , L_2 , L_3 and L_4 , the following equations can be obtained:

$$(V_{in} - V_{C1} - V_o) \times D_1 T_s + (-V_o) \times D_2 T_s = 0 \quad (31)$$

$$(V_{C1} - V_{C2} - V_o) \times D_1 T_s + (-V_o) \times D_2 T_s = 0 \quad (32)$$

$$(V_{C2} - V_{C3} - V_o) \times D_1 T_s + (-V_o) \times D_2 T_s = 0 \quad (33)$$

$$(V_{C3} - V_o) \times D_1 T_s + (-V_o) \times D_2 T_s = 0 \quad (34)$$

Then, after sorting (34) and substituting it back to (31), (32) and (33), the voltages across the energy-transferring capacitors, called V_{C1} , V_{C2} and V_{C3} , and the relationship between V_{in} and V_o can be obtained, respectively:

$$V_{C3} = \frac{D_1 + D_2}{D_1} \times V_o = \frac{1}{4} V_{in} \quad (35)$$

$$V_{C2} = \frac{D_1 + D_2}{D_1} \times 2V_o = \frac{1}{2} V_{in} \quad (36)$$

$$V_{C1} = \frac{D_1 + D_2}{D_1} \times 3V_o = \frac{3}{4} V_{in} \quad (37)$$

$$V_{in} = \frac{D_1 + D_2}{D_1} \times 4V_o \quad (38)$$

Since the output current will be equal to the average value of the inductor current, it can be obtained that the relational expression is as shown in (39), where R signifies the equivalent resistance of the LED, equal to the LED forward voltage divided by the LED forward current, and $i_{L1,max}$ is the maximum current in L_1 :

$$\frac{1}{2} \times i_{L1,max} \times (D_1 + D_2) = I_o = \frac{V_o}{R} \quad (39)$$

According to the inductor voltage and current formula, it can be known that:

$$D_2 T_s = \frac{L_1 \times i_{L1,max}}{V_o} = \frac{2L_1}{R(D_1 + D_2)} \quad (40)$$

Then, by substituting (41) into (40) and solving out, (42) can be obtained:

$$K \equiv \frac{2L_1}{RT_s} \quad (41)$$

$$\begin{aligned} D_2^2 + D_1 D_2 - K &= 0 \\ \Rightarrow D_2 &= D_1 \times \frac{\sqrt{1 + \frac{4K}{D_1^2}} - 1}{2} \end{aligned} \quad (42)$$

Eventually, by substituting (42) into (38) can obtain the voltage conversion ratio M :

$$M = \frac{V_o}{V_{in}} = \frac{1}{2 \left[1 + \sqrt{\frac{4K}{D_1^2} + 1} \right]} \quad (43)$$

From (43), the voltage conversion ratio M in DCM is related to the load.

2.5. Boundary Condition for Inductor L_1

Since the current ripple Δi_{L1} flowing through the inductor L_1 can be expressed as:

$$\Delta i_{L1} = \frac{v_{Li} \Delta t}{L_1} = \frac{(V_{in} - V_{C1} - V_o) \times D T_s}{L_1} \quad (44)$$

As $I_{L1} \geq \frac{\Delta i_{L1}}{2}$, the inductor L_1 will operate in CCM, namely:

$$\begin{aligned} I_{L1} &\geq \frac{\Delta i_{L1}}{2} \\ \Rightarrow I_{L1} &\geq \frac{(V_{in} - V_{C1} - V_o) \times DT_s}{2 \times L_1} \\ \Rightarrow L_1 &\geq \frac{(V_{in} - V_{C1} - V_o) \times DT_s}{2 \times I_{o1}} \end{aligned} \quad (45)$$

As (45) is satisfied, the inductor L_1 operates in CCM; otherwise, it operates in DCM.

2.6. Boundary Condition for Inductors L_2 , L_3 and L_4

As the boundary condition analyses for the inductors L_2 , L_3 , and L_4 are the same as that for the inductor L_1 , it will not be redescribed herein.

2.7. Analysis of Automatic Current Sharing Principle

The principle of automatic current sharing in DCM is the same as that in CCM. The average values of the charging and discharging currents of the energy-transferring capacitors C_1 , C_2 , and C_3 are the same, so automatic current sharing between four phases can be achieved.

2.8. Structure Extension

The proposed LED driver can be extended to six phases as shown in Figure 10. Its operating principle is the same as that of the proposed LED driver. As long as the adjacent switches are not turned on at the same time and the duty cycle is not larger than 0.5, the automatic current sharing of multiple strings of LEDs can be achieved. Although the extended circuit has more components, it has a lower voltage conversion ratio.

Figure 10. Six-phase LED driver.

3. Design Considerations

Figure 11 is the LED drive circuit with automatic current sharing proposed. The system includes the main power stage and the feedback control circuit. As for the feedback control circuit, it uses the current detection IC, named ACS712, to obtain the analog signal of the output current. Then, based on the sampling without ADC technique [18], the corresponding digital signal is obtained and sent to the FPGA for calculation to create the required digital control force. Such a digital control force is sent to a gate driver TLP250H with galvanic isolation function to control the switch.

Table 1 shows the system specifications. Because the current is small in the light load, the inductance will be too large if all this load is operated in CCM. So, 25% of the rated load is designed in DCM. Table 2 shows the LED specifications. The LEDs used are high-brightness LEDs manufactured EVERLIGHT Electronics Co., Ltd. Table 3 shows a summary of the components of the proposed circuit. Table 4 shows the FPGA specifications.

Figure 11. System configuration diagram.

Table 1. System specifications.

Parameter Name	Value
System operation mode	Rated loaded: CCM Half load: CCM Light load: DCM
Rated input voltage (V_{in})	400 V
Rated output voltage (V_o)	27.6 V ($=3.45 \times 8$)
Rated output current ($I_{o, rated}$)	1.4 A ($=0.35 \text{ A} \times 4$)
Rated output power ($P_{o, rated}$)	38.64 W
Minimum output current under CCM ($I_{o, min, CCM}$)/power ($P_{o, min, CCM}$)	0.175 A/17.36 W
Minimum output current under DCM ($I_{o, min, DCM}$)/power ($P_{o, min, DCM}$)	0.0875 A/8.12 W
Switching frequency (f_s)/period (T_s)	100 kHz/10 μ s

Table 2. LED specifications.

Parameter Name	Value
Forward voltage (V_F)	2.95 V~3.85 V with a typical value of 3.45 V
DC operating current ($I_{F, max}$)	400 mA
Pulsed forward current	500 mA
Junction temperature	125 °C
Operating temperature	−40 °C~ +85 °C
Typical light flux output	100 lm @ 350 mA

Table 3. Component specifications.

Component	Product Name
S_1, S_2, S_3, S_4	SPA07N60C3
D_1, D_2, D_3, D_4	DSEP8-02A
C_1, C_2, C_3	47 μ F/400 V LTEC electrolytic capacitor
$C_{o1}, C_{o2}, C_{o3}, C_{o4}$	220 μ F/50 V electrolytic capacitor
L_1, L_2, L_3, L_4	Core: PQ20/16 $L_1 = 603.5 \mu$ H, $L_2 = 601.7 \mu$ H, $L_3 = 598.2 \mu$ H, $L_4 = 599.1 \mu$ H
Gate driver	TLP250H
Current sensor	ACS712
LED	EHP-AX08EL/GT01H-P01/5670/Y/K42

Table 4. EP3C5E144C8N specifications.

Device	Logic Elements	Total RAM Bits	18 × 18 Multipliers	PLLs	User I/O Pins
EP3C5E144C8N	5136	423,936	23	2	94

3.1. Design of Inductors L_1 , L_2 , L_3 and L_4

The circuit operating principle in CCM has been described in Section 2.1, the behavior of the four inductors is the same if these inductors are identical, so the following only focuses on design of the first inductor L_1 . From (17), the duty cycle D can be expressed to be:

$$D = \frac{4V_o}{V_{in}} \quad (46)$$

Based on Table 2 and (46), the corresponding duty cycles for 100% and 50% loads are 0.276 and 0.248, respectively, to be calculated as follows:

$$D_{100\%Load} = \frac{4V_o}{V_{in}} = \frac{4 \times 8 \times 3.45}{400} = 0.276 \quad (47)$$

$$D_{50\%Load} = \frac{4V_o}{V_{in}} = \frac{4 \times 8 \times 3.1}{400} = 0.248 \quad (48)$$

If the inductor is to operate in inductive current continuous mode, the value of inductor L_1 must meet the inequalities (45):

$$\begin{aligned} L_1 &\geq \frac{(V_{in} - V_{C1} - V_o) \times DT_s}{2 \times I_{L1}} \\ &= \frac{(400 - 300 - 24.8) \times 0.248 \times 10^{-5}}{2 \times 0.175} \\ &\cong 532.85 \mu\text{H} \end{aligned} \quad (49)$$

By the same way for L_2 , L_3 and L_4 , the values for L_1 , L_2 , L_3 and L_4 are finally chosen to be 603.5 μH , 601.7 μH , 598.2 μH and 599.1 μH , respectively.

3.2. Design of Energy-Transferring Capacitors C_1 , C_2 and C_3

From Figure 12, it can be seen that during the period of DT_s , the constant current of C_1 , called $I_{C1,DTs}$, is the same as that of L_1 , called $I_{L1,DTs}$, that is,

$$I_{C1,DTs} = I_{L1,DTs} = I_{L1} \quad (50)$$

Substituting the rated input voltage V_{in} shown in Table 1 into (18), the DC voltage across C_1 is:

$$V_{C1} = 300 \text{ V} \quad (51)$$

Assume that the peak-to-peak voltage ripple Δv_C is within 0.1% of V_{C1} , and the corresponding inequality is:

$$\begin{aligned} C_1 &\geq \frac{I_{C1,DTs} \cdot DT_s}{\Delta v_{C1}} \\ &= \frac{I_{L1} \cdot DT_s}{0.1\% \times V_{C1}} \\ &= \frac{0.35 \times 0.276 \times 10}{0.001 \times 300} \mu\text{F} = 3.22 \mu\text{F} \end{aligned} \quad (52)$$

By the same way for C_2 and C_3 and considering the effect of switching frequency on electrolytic capacitance, the values for C_1 , C_2 , and C_3 are all set at 47 μF /400 V based on measurements from the LCR meter.

3.3. Design of Output Capacitors C_{o1} , C_{o2} and C_{o3}

Assume that the peak-to-peak voltage ripple $\Delta v_{C_{o1}}$ is within 0.1% of V_{o1} , and the corresponding inequality is:

$$\begin{aligned} C_{o1} &\geq \frac{\Delta i_{L1} \times T_s}{8 \times \Delta v_{C_{o1}}} \\ &= \frac{0.34 \times 10^{-5}}{8 \times 0.001 \times 27.6} = 15.4 \mu\text{F} \end{aligned} \quad (53)$$

By the same way for C_{o2} , C_{o3} and C_{o4} and considering the effect of switching frequency on electrolytic capacitance, the values for C_1 , C_2 , and C_3 are all set at $220 \mu\text{F}/50 \text{ V}$ based on measurements from the LCR meter.

Figure 12. Current waveforms relevant to energy-transferring capacitors C_1 , C_2 and C_3 .

4. Experimental Results

4.1. Measured Waveforms

Figure 13 is a photo of the proposed LED driver. Figures 14–21 are obtained at rated load in CCM whereas Figures 22–29 are obtained at 25% rated load in DCM. From Figures 15, 16, 23 and 24, under any two loads, except for the voltage across switch S_1 , v_{ds1} , the voltages across the switches S_2 , S_3 and S_4 are all in a three-stage form, and the voltage stress on switch S_1 is the smallest, which is 100 V (0.25 times of the input voltage), and the voltage stresses on switches S_2 , S_3 , and S_4 are all 200 V (0.5 times the input voltage).

Figure 13. Photo of experimental setup.

Figure 14. Measured waveforms relevant to rated load: (1) v_{gs1} ; (2) v_{gs2} ; (3) v_{gs3} ; (4) v_{gs4} .

Figure 15. Measured waveforms relevant to rated load: (1) v_{gs1} ; (2) v_{gs2} ; (3) v_{ds1} ; (4) v_{ds2} .

Figure 16. Measured waveforms relevant to rated load: (1) v_{gs3} ; (2) v_{gs4} ; (3) v_{ds3} ; (4) v_{ds4} .

Figure 17. Measured waveforms relevant to rated load: (1) V_{in} ; (2) V_{C1} ; (3) V_{C2} ; (4) V_{C3} .

Figure 18. Measured waveforms relevant to rated load: (1) v_{L1} ; (2) i_{L1} ; (3) v_{L2} ; (4) i_{L2} .

Figure 19. Measured waveforms relevant to rated load: (1) v_{L3} ; (2) i_{L3} ; (3) v_{L4} ; (4) i_{L4} .

Figure 20. Measured waveforms relevant to rated load: (1) v_{D1} ; (2) v_{D2} ; (3) v_{D3} ; (4) v_{D4} .

Figure 21. Measured waveforms relevant to rated load: (1) I_{o1} ; (2) I_{o2} ; (3) I_{o3} ; (4) I_{o4} .

Figure 22. Measured waveforms relevant to 25% of the rated load: (1) v_{gs1} ; (2) v_{gs2} ; (3) v_{gs3} ; (4) v_{gs4} .

Figure 23. Measured waveforms relevant to 25% of the rated load: (1) v_{gs1} ; (2) v_{gs2} ; (3) v_{ds1} ; (4) v_{ds2} .

Figure 24. Measured waveforms relevant to 25% of the rated load: (1) v_{gs3} ; (2) v_{gs4} ; (3) v_{ds3} ; (4) v_{ds4} .

Figure 25. Measured waveforms relevant to 25% of the rated load: (1) V_{in} ; (2) V_{C1} ; (3) V_{C2} ; (4) V_{C3} .

Figure 26. Measured waveforms relevant to 25% of the rated load: (1) v_{L1} ; (2) i_{L1} ; (3) v_{L2} ; (4) i_{L2} .

Figure 27. Measured waveforms relevant to 25% of the rated load: (1) v_{L3} ; (2) i_{L3} ; (3) v_{L4} ; (4) i_{L4} .

Figure 28. Measured waveforms relevant to 25% of the rated load: (1) v_{D1} ; (2) v_{D2} ; (3) v_{D3} ; (4) v_{D4} .

Figure 29. Measured waveforms relevant to 25% of the rated load: (1) I_{o1} ; (2) I_{o2} ; (3) I_{o3} ; (4) I_{o4} .

From Figures 17 and 25, the voltages on three energy-transferring capacitors C_1 , C_2 , and C_3 , called V_{C1} , V_{C2} and V_{C3} , are all stabilized at different certain voltage values, which will not vary with the load. Also, it can be known that from Figures 18, 19, 26 and 27, the DC values of the four inductor currents i_{L1} , i_{L2} , i_{L3} and i_{L4} are approximately the same under each load, implying that the load current can be evenly distributed among four phases. From Figures 20 and 28, under any load, the voltages on the four diodes D_1 , D_2 , D_3 and D_4 are all 100 V (0.25 times the input voltage). Furthermore, it can be known that from Figures 21 and 29, the output current can be evenly distributed among four phases.

In addition, in Figure 28, the voltage on the diode has a small resonance waveform before the cut-off. This is caused by the resonance of the parasitic capacitance on the diode and the inductance in the circuit, and will affect the voltages on the other components in the circuit, as shown in Figures 23, 24, 26 and 27.

4.2. LED Current Error Percentage

According to the experimental data measured above, the current error percentage of the LED string under different loads is calculated, as shown in Table 5. The current error percentage calculation formula used is:

$$e_y = \frac{\left(I_{oy} - \frac{1}{n} \sum_{x=1}^n I_{ox} \right)}{\frac{1}{n} \sum_{x=1}^n I_{ox}} \times 100\% \quad (54)$$

where e_y is the current error percentage of the y -th LED string, n is the number of LED strings, and I_{oy} is the current flowing through the y -th LED string, $\sum_{x=1}^n I_{ox}$ is the sum of all LED string currents. The values of e_y can be obtained by using (54) and the current measurement values under different loads.

Table 5. Current error percentage under different loads.

Load Percentage		LS_1	LS_2	LS_3	LS_4
100%	I_{on} (mA)	348	349	351	350
	e_n (%)	0.43	0.14	−0.43	−0.14
75%	I_{on} (mA)	259	257	260	261
	e_n (%)	0.096	0.87	−0.29	−0.67
50%	I_{on} (mA)	173	176	173	177
	e_n (%)	1.00	−0.72	1.00	−1.29
25%	I_{on} (mA)	86.1	88.0	85.6	90.2
	e_n (%)	1.57	−0.6	2.14	−3.12

Therefore, it can be known that from Table 5, the absolute error percentage is below 1.3% except for 25% load. As for 25% load, the maximum absolute error percentage is 3.12% and this is because the LED driver operates in DCM.

4.3. Discussion on Unequal Numbers of LEDs for Four Outputs

Figures 30–33 are simulations of the unequal LED loads of the four outputs. Only the number of LEDs on LS_2 and LS_4 is reduced by two, and the component specifications in the circuit remain unchanged. By the way, the simulation software is called PSIM, made by Powersim, Inc., Rockville, MD, USA.

Figure 30. Simulated waveforms relevant to different numbers of LEDs for LED strings: (1) V_{in} ; (2) V_{C1} ; (3) V_{C2} ; (4) V_{C3} .

Figure 31. Simulated waveforms relevant to different numbers of LEDs for LED strings: (1) v_{gs1} ; (2) v_{gs2} ; (3) v_{ds1} ; (4) v_{ds2} .

Figure 32. Simulated waveforms relevant to different numbers of LEDs for LED strings: (1) i_{L1} ; (2) i_{L2} ; (3) i_{L3} ; (4) i_{L4} .

Figure 33. Simulated waveforms relevant to different numbers of LEDs for LED strings: (1) I_{o1} ; (2) I_{o2} ; (3) I_{o3} ; (4) I_{o4} .

It can be seen that from Figure 30, in the case of unequal loads, the difference in voltage between different loads will appear on the energy-transferring capacitor, which in turn affects the voltages across the switches, as shown in Figure 31. Because the voltage of the energy-transferring capacitor and the output voltage change, the slopes of magnetization and demagnetization on the inductor currents will also change, as shown in Figure 32. It can be seen from Figure 33 that the different numbers of LEDs on the loads do not affect the characteristics of the capacitor current sharing, and the currents between the four-phase outputs can still achieve automatic current sharing.

4.4. Efficiency Measurement

The method for efficiency measurement is shown in Figure 34. First, use a digital meter (Fluke 179) to measure the input voltage and current, and output voltage and current, so that the input power and output power can be obtained. Eventually, the efficiency can be figured out with the obtained input and output power, as shown in Figure 35. From Figure 35, the efficiency is above 85% all over the load current range and the maximum efficiency is about 90.8%.

Figure 34. Block diagram of efficiency measurement.

Figure 35. Curve of efficiency versus load current.

4.5. Comparison

The circuits in [11–15], with capacitive current balance, are chosen for comparison. In Table 6, the comparison items contain converter type, output count, component count, structure extension, regenerative snubber, switching frequency, resonance, isolation, input voltage, output voltage, and maximum efficiency. From this table, it can be seen that the proposed LED has high step-down voltage gain from 400 V to 27.6 V.

Table 6. Circuit comparison between the existing [11–15] and the proposed.

Item	[11]	[12]	[13]	[14]	[15]	Proposed
Converter type	Step-up	Step-up	Step-up	Step-down	Step-down	Step-down
Output count	2	2	3	2	6	4
Component count	Switch×1 Diode×2 Capacitor×4 Transformer×1	Switch×1 Diode×3 Capacitor4 Coupled Inductor×1	Switch×1 Diode×3 Capacitor×5 Inductor×2	Switch×4 Diode×2 Capacitor×5 Inductor×1	Switch×2 Diode×6 Capacitor×11 Inductor×3	Switch×4 Diode×4 Capacitor×7 Inductor×4
Structure extension	Yes	No	Yes	Yes	Yes	Yes
Regenerative snubber	No	Yes	No	No	No	No
Switching frequency	100 kHz	100 kHz	300 kHz	65 kHz	77 kHz	100 kHz
Resonance	No	No	Yes	Yes	Yes	No
Isolation	Yes	No	No	No	No	No
Input voltage	12 V	3.3 V	19 V	100 V	100 V	400 V
Output voltage	17.3 V	17.5 V	29.7 V	25 V	49.2 V	27.6 V
Maximum efficiency	98.8%	92.8%	91.9%	92.6%	93.5%	90.8%

4.6. Loss Breakdown Analysis

The loss analysis will be performed at room temperature and the proposed circuit is operated at rated load, without considering the copper loss under the skin effect, the line loss of the circuit layout, and the additional switching loss caused by the voltage spike on the switch. At the same time, the power loss of each component is estimated by its specifications. For analysis convenience, it is assumed that the four channels are identical except for four inductor resistances, and hence only the components in the first channel will be considered. Prior to estimating component losses, some currents such as $I_{L1,max}$, $I_{L1,rms}$, $I_{ds1,max}$, $I_{ds1,rms}$, $I_{C1,rms}$, $I_{Co1,rms}$, to be shown below:

$$I_{L1,max} = I_{L1} + 0.5 \times \Delta i_{L1} \Rightarrow I_{L1,max} = 0.35 + 0.5 \times 0.34 = 0.52 \text{ A} \quad (55)$$

$$I_{L1,rms} = \left[I_{L1}^2 + (0.5 \times \Delta i_{L1})^2 / 3 \right]^{1/2} \Rightarrow I_{L1,rms} = \left[0.35^2 + (0.5 \times 0.34)^2 / 3 \right]^{1/2} = 0.364 \text{ A} \quad (56)$$

$$I_{ds1,max} = I_{L1,max} \Rightarrow I_{ds1,max} = 0.52 \text{ A} \quad (57)$$

$$I_{ds1,rms} = \sqrt{D} \times I_{L1,rms} \Rightarrow I_{ds1,rms} = \sqrt{0.276} \times 0.364 = 0.191 \text{ A} \quad (58)$$

$$I_{C1,rms} = \sqrt{2} \times I_{ds1,rms} = \sqrt{2} \times 0.191 = 0.27 \text{ A} \quad (59)$$

$$I_{Co1,rms} = \Delta i_{L1,rms} = 0.5 \times \Delta i_{L1} / \sqrt{3} = 0.5 \times 0.34 / \sqrt{3} = 0.098 \text{ A} \quad (60)$$

where $I_{L1,max}$ and $I_{ds1,max}$ are the maximum currents in L_1 and S_1 , respectively, $I_{L1,rms}$, $I_{ds1,rms}$, $I_{C1,rms}$ and $I_{Co1,rms}$ are the rms currents of L_1 , S_1 , C_1 and C_o , respectively, and $\Delta i_{L1,rms}$ is the rms value of the current ripple of L_1 .

Based on (60), the current ripple factor, obtained by $\Delta i_{L1,rms}$ divided by I_{L1} , is $0.098/0.52$, equal to 0.28.

4.6.1. Loss of Switches S_1 , S_2 , S_3 , S_4

$P_{on,cond}$ is the switch conduction loss, $P_{turn-on}$ is the switch conduction switching loss, $P_{turn-off}$ is the turn-off switching loss, $P_{discharge}$ is the output charge loss, $P_{g,dri}$ is the drive

loss, and V_{ds} is the voltage on the switch when the switch is off. $I_{ds,max}$ is the maximum current when the switch is on, C_{oss} is the output capacitance of the switch, $Q_{g-total}$ is the total charge of the gate, $R_{ds,on}$ is the resistance when the switch is fully on, t_r is the rise time when the switch is on, and t_f is the fall time when the switch is off. According to the datasheet of SPA20N60C3, it can be known that $R_{ds,on}$ is 600 mΩ, C_{oss} is 260 pF, $Q_{g-total}$ is 87 nC, t_r is 5 ns, and t_f is 4.5 ns.

$$\begin{aligned} P_{S,loss} &= P_{on,cond} + P_{turn-on} + P_{turn-off} + P_{discharge} + P_{g,dri} \\ &= I_{ds,rms}^2 \times R_{ds,on} + \left(\frac{V_{ds} \times I_{ds,max} \times t_r \times f_s}{6} \right) \\ &\quad + \left(\frac{V_{ds} \times I_{ds,max} \times t_f \times f_s}{6} \right) + \frac{4}{3} (C_{oss} \times V_{ds}^2 \times f_s) \\ &\quad + (Q_{g-total} \times V_{gs} \times f_s) \end{aligned} \quad (61)$$

$$\begin{aligned} P_{S1,Loss} &= 0.191^2 \times 0.6 + \frac{100 \times 0.52 \times 5 \text{ n} \times 100 \text{ k}}{6} + \frac{100 \times 0.52 \times 4.5 \text{ n} \times 100 \text{ k}}{6} \\ &\quad + \frac{4}{3} (260 \text{ p} \times 100^2 \times 100 \text{ k}) + (87 \text{ n} \times 15 \times 100 \text{ k}) \\ &= 0.506 \text{ W} \end{aligned} \quad (62)$$

$$\begin{aligned} P_{S,Loss,Total} &= P_{S1,Loss} + P_{S2,Loss} + P_{S3,Loss} + P_{S4,Loss} \\ &= 0.506 + 0.506 + 0.506 + 0.506 = 2.024 \text{ W} \end{aligned} \quad (63)$$

4.6.2. Loss of Diodes D₁, D₂, D₃, D₄

I_D is the rated-load current flowing through the diode, V_F is the forward voltage of the diode, V_D is the voltage across the diode when the diode is off, I_R is the maximum value of the reverse recovery current, and t_{rr} is the time for the reverse recovery current. f_s is the switching frequency. According to the datasheet of DSEP8-02A, t_{rr} is 25 ns, and V_F is measured to be 0.45 V.

$$P_{D,Loss} = I_D \times V_F + \frac{1}{2} \times V_D \times I_R \times T_{rr} \times f_s \quad (64)$$

$$\begin{aligned} P_{D1,Loss} &= 0.35 \times 0.45 + \frac{1}{2} \times 100 \times 50 \mu \times 25 \text{ n} \times 100 \text{ k} \\ &= 0.158 \text{ W} \end{aligned} \quad (65)$$

$$\begin{aligned} P_{D,Loss,Total} &= P_{D1,Loss} + P_{D2,Loss} + P_{D3,Loss} + P_{D4,Loss} \\ &= 0.158 + 0.158 + 0.158 + 0.158 = 0.632 \text{ W} \end{aligned} \quad (66)$$

4.6.3. Loss of Inductances L₁, L₂, L₃ and L₄

Copper Loss

$$P_{L1,Loss} = I_{L1,rms}^2 \times R_{L1} = 0.364^2 \times 45.13 \text{ m} = 6.41 \text{ mW} \quad (67)$$

$$P_{L2,Loss} = I_{L2,rms}^2 \times R_{L2} = 0.364^2 \times 39.99 \text{ m} = 5.68 \text{ mW} \quad (68)$$

$$P_{L3,Loss} = I_{L3,rms}^2 \times R_{L3} = 0.364^2 \times 37.12 \text{ m} = 5.28 \text{ mW} \quad (69)$$

$$P_{L4,Loss} = I_{L4,rms}^2 \times R_{L4} = 0.364^2 \times 43.61 \text{ m} = 6.20 \text{ mW} \quad (70)$$

Iron Loss

$$\begin{aligned} B_{m1} &= \frac{\Delta B_1}{2} = \frac{L_1 \times I_{L1,max}}{2 \times N \times A_e} \\ &= \frac{600 \times 10^{-6} \times 0.52}{2 \times 17 \times 0.62} \times 10^8 \\ &= 148 \text{ mT} \end{aligned} \quad (71)$$

$$\begin{aligned}
 P_{fe1} &= 0.15 \text{ W/cm}^3 \times V_e \\
 &= 0.15 \text{ W/cm}^3 \times 2.79 \text{ cm}^3 \\
 &= 0.419 \text{ W}
 \end{aligned} \tag{72}$$

where R_{L1} , R_{L2} , R_{L3} and R_{L4} are the resistance values of inductors L_1 , L_2 , L_3 and L_4 , respectively. According to the core specification table of Ferrite MB4, the core of this material will have a loss of 0.15 W/cm^3 at $B_{m1} = 148 \text{ mT}$. Multiplying this loss by the effective core volume V_e yields the estimated core loss.

Eventually, adding the results of (67), (68), (69), (70), and (72) yields the power loss of the inductors:

$$\begin{aligned}
 P_{L, \text{Loss}, \text{Total}} &= P_{L1, \text{Loss}} + P_{L2, \text{Loss}} + P_{L3, \text{Loss}} + P_{L4, \text{Loss}} + P_{fe1} \times 4 \\
 &= 6.41 \text{ m} + 5.68 \text{ m} + 5.28 \text{ m} + 6.20 \text{ m} + 0.419 \times 4 \\
 &= 1.7 \text{ W}
 \end{aligned} \tag{73}$$

4.6.4. Loss of Capacitors C_1 , C_2 , C_3 and C_0

The resistors R_{C1} , R_{C2} , R_{C3} and R_{C0} are the equivalent series resistance (ESR) measured by the capacitors C_1 , C_2 , C_3 , C_4 , C_5 and C_0 at 100 kHz , respectively.

$$P_{C1, \text{Loss}} = I_{C1, \text{rms}}^2 \times R_{C1} = 0.27^2 \times 0.5 = 30.5 \text{ mW} \tag{74}$$

$$P_{C2, \text{Loss}} = I_{C2, \text{rms}}^2 \times R_{C2} = 0.27^2 \times 0.5 = 30.5 \text{ mW} \tag{75}$$

$$P_{C3, \text{Loss}} = I_{C3, \text{rms}}^2 \times R_{C3} = 0.27^2 \times 0.5 = 30.5 \text{ mW} \tag{76}$$

$$P_{C01, \text{Loss}} = I_{C01, \text{rms}}^2 \times R_{C01} = 0.098^2 \times 45 \text{ m} = 0.87 \text{ mW} \tag{77}$$

$$\begin{aligned}
 P_{C, \text{Loss}, \text{Total}} &= P_{C1, \text{Loss}} + P_{C2, \text{Loss}} + P_{C3, \text{Loss}} + P_{C01, \text{Loss}} \times 4 \\
 &= 30.5 \text{ m} + 30.5 \text{ m} + 30.5 \text{ m} + 0.87 \text{ m} \times 4 \\
 &= 94.98 \text{ mW}
 \end{aligned} \tag{78}$$

After adding the results of (63), (66), (73), and (78), the estimated total loss is

$$\begin{aligned}
 P_{\text{Loss}, \text{Total}} &= P_{S, \text{Loss}, \text{Total}} + P_{D, \text{Loss}, \text{Total}} + P_{L, \text{Loss}, \text{Total}} + P_{C, \text{Loss}, \text{Total}} \\
 &= 2.024 + 0.632 + 1.7 + 0.095 \\
 &= 4.44 \text{ W}
 \end{aligned} \tag{79}$$

According to the calculation result of (79), it can be known that the conversion efficiency at rated load is:

$$\begin{aligned}
 \eta &= \frac{P_{o, \text{rated}}}{P_{o, \text{rated}} + P_{\text{Loss}, \text{Total}}} \\
 &= \frac{38.64}{38.64 + 4.44} \times 100\% = 89.69\%
 \end{aligned} \tag{80}$$

From the result of (80), the estimated efficiency of the proposed circuit at rated load is 89.69%, which is about 1.1% different from the actual measured value of 90.8%. The reason is that the measurement environment of the relevant parameters on the datasheet of each component is different from that due to the mentioned LED driver. For example, the measurement environment for switches to measure t_r and t_f is under the condition that $v_{ds, \text{off}} = 380 \text{ V}$ and $i_{ds, \text{avg}} = 7.3 \text{ A}$, but the actual circuit parameters are $v_{ds, \text{off}} = 200 \text{ V}$, $i_{ds, \text{avg}} = 0.35 \text{ A}$, so the estimated efficiency is lower than the actual efficiency. Figure 36 shows the estimated power loss distribution percentage. It can be seen from this figure that the power loss caused by the inductance and switches has the greatest impact on the overall efficiency. If a wire diameter with a lower resistance value or an active switch with better features can be selected, then the efficiency of the circuit can be improved further.

Figure 36. Loss breakdown analysis.

5. Conclusions

The proposed LED driver has several merits as following:

- (1) This LED driver inherently possesses automatic current sharing and high step-down voltage gain.
- (2) This circuit is of SIMO and structure extension, and under automatic current sharing, the number of LED strings for each output can be different.
- (3) As compared with the traditional four-phase interleaved buck converter under automatic current sharing, the maximum duty cycle of the proposed circuit is relatively large and not changed for structure extension whereas the maximum duty cycle of the traditional circuit will be reduced as the number of outputs is increased.
- (4) As compared with automatic current sharing by using the differential-mode transformer, the proposed circuit has no magnetic resetting loop required.
- (5) The efficiency is above 85% over all the load current and the maximum efficiency is about 90.8%.

By the way, in the future, the soft switching technique will be applied to this circuit to improve the efficiency further.

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References

1. Yu, W.; Lai, J.S.; Ma, H.; Zheng, C. High-efficiency DC-DC converter with twin bus for dimmable LED lighting. *IEEE Trans. Power Electron.* **2011**, *26*, 2095–2100. [\[CrossRef\]](#)
2. Jiang, W.-Z.; Hwu, K.-I.; Shieh, J.-J. LLC LED driver with current-sharing capacitor having low voltage stress. *Energies*. **2021**, *14*, 112. [\[CrossRef\]](#)
3. Zhang, Y.; Rong, G.; Qu, S.; Song, Q.; Tang, X.; Zhang, Y. A high-power LED driver based on single inductor-multiple output DC-DC converter with high dimming frequency and wide dimming range. *IEEE Trans. Power Electron.* **2020**, *35*, 8501–8511. [\[CrossRef\]](#)

4. Sun, J.; Tang, X.; Xing, Y.; Chen, B.; Wu, H.; Sun, K. Current sharing control of interleaved LLC resonant converter with hybrid rectifier. In Proceedings of the Applied Power Electronics Conference and Exposition, Anaheim, CA, USA, 17–21 March 2019; pp. 2223–2227.
5. Mazelan, F.-N.; Kannan, R.; Hasan, K.N.-M.; Ali, A. Multi-input power converter for renewable energy sources using active current sharing schemes. In Proceedings of the IEEE Student Conference on Research and Development, Bandar Seri Iskandar, Malaysia, 15–17 October 2019; pp. 275–279.
6. Yau, Y.T.; Hwu, K.I.; Shieh, J.J. Minimization of output voltage ripple of two-phase interleaved buck converter with active clamp. *Energies* **2021**, *14*, 5215. [[CrossRef](#)]
7. Martins, M.; Perdigao, M.S.; Mendes, A.M.S.; Pinto, R.A.; Alonso, J.M. Analysis, design, and experimentation of a dimmable resonant-switched-capacitor LED driver with variable inductor control. *IEEE Trans. Power Electron.* **2017**, *32*, 3051–3062. [[CrossRef](#)]
8. Hwu, K.-I.; Tu, W.-C.; Hong, M.-J. A dimmable LED driver based on current balancing transformer with magnetizing energy recycling considered. *IEEE J. Disp. Technol.* **2014**, *10*, 388–395. [[CrossRef](#)]
9. Lin, Y.L.; Chiu, H.J.; Lo, Y.K.; Leng, C.M. Light-emitting diode driver with a combined energy transfer inductor for current balancing control. *IET Power Electron.* **2015**, *8*, 1834–1843. [[CrossRef](#)]
10. Jiang, W.-Z.; Hwu, K.-I.; Chen, H.-H. Applying hybrid passive current-sharing components to non-isolated LED driver. In Proceedings of the International Symposium on Computer, Consumer and Control, Taichung City, Taiwan, 13–16 November 2020; pp. 259–262.
11. Hwu, K.-I.; Jiang, W.-Z. Expendable two-channel LED driver with galvanic isolation and automatic current balance. *IET Power Electron.* **2018**, *11*, 825–833. [[CrossRef](#)]
12. Hwu, K.-I.; Jiang, W.-Z. Single-switch coupled-inductor-based two-channel LED driver with a passive regenerative snubber. *IEEE Trans. Power Electron.* **2017**, *32*, 4482–4490. [[CrossRef](#)]
13. Liu, X.; Zhou, Q.; Xu, J.; Lei, Y.; Wang, P.; Zhu, Y. High-efficiency resonant LED backlight driver with passive current balancing and dimming. *IEEE Trans. Ind. Electron.* **2018**, *65*, 5476–5486. [[CrossRef](#)]
14. Zhang, X.; Cai, H.; Guan, Y.; Han, S.; Wang, Y.; Dalla Costa, M.A.; Alonso, J.M.; Xu, D. A soft-switching transformer-less step-down converter based on resonant current balance module. *IEEE Trans. Power Electron.* **2021**, *36*, 8206–8218. [[CrossRef](#)]
15. Gucin, T.N.; Fincan, B.; Biberoglu, M. A series resonant converter-based multichannel LED inherent current balancing and dimming capability. *IEEE Trans. Ind. Electron.* **2019**, *34*, 2693–2703. [[CrossRef](#)]
16. Chuang, C.F.; Pan, C.T.; Cheng, H.C. A novel transformer-less interleaved four-phase step-down DC converter with low switch voltage stress and automatic uniform current-sharing characteristics. *IEEE Trans. Power Electron.* **2016**, *31*, 406–441. [[CrossRef](#)]
17. Hwu, K.-I.; Jiang, W.-Z.; Wu, P.-Y. An expandable four-phase interleaved high step-down converter with low switch voltage stress and automatic uniform current sharing. *IEEE Trans. Ind. Electron.* **2016**, *63*, 6064–6072. [[CrossRef](#)]
18. Hwu, K.-I.; Yau, Y.-T. Performance enhancement of boost converter based on PID controller plus linear-to-nonlinear translator. *IEEE Trans. Power Electron.* **2010**, *25*, 1351–1361. [[CrossRef](#)]