

Article

A Soft-Switching Proportional Dimming LED Driver Based on Switched Capacitor

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Abstract: This paper proposes a switched capacitor-based soft-switching proportional dimming LED driver to address color-mixing LED applications. The driver utilizes a half-bridge switch structure and incorporates a switched capacitor to regulate the dual resonant network. This configuration facilitates current sharing across each resonant network and enables fixed proportion dimming between them. The common inductance leveraged in this design results in fewer switching devices, thereby enhancing efficiency. Furthermore, all switches and diodes achieve soft switching, contributing to high efficiency. Finally, this paper built a 19 W experimental prototype, and the maximum efficiency reached 95.56%, which verified the correctness and feasibility of the driver.

Keywords: soft switching; current sharing; dimming

1. Introduction

In recent years, light-emitting diodes (LEDs) have rapidly advanced due to their energy-saving, environmental-friendly, long-lasting, and flexible attributes, making them extensively utilized in commercial, residential, and industrial lighting [1,2]. LEDs offer versatility in terms of their form factor and are often employed in series and parallel connections. For applications in plant factory lighting systems, LED lamps are preferred over natural lighting for plant growth, due to their cost effectiveness, high power efficiency, and specialized spectrum [3,4]. The predominant red and blue light ratio utilized in plant growth LED lamps necessitates a specific current-dimming proportion for multi-channel LED drivers, making it essential to investigate the current control of multi-channel output LED drivers. The volt–ampere characteristics of LEDs link the luminous efficiency to the forward current, leading to the prevalent use of a constant current drive to ensure balanced luminous intensity across branches. Current-controllable multi-channel LED drivers are primarily designed for multi-color lighting and color-mixing applications, necessitating the capability to output current across multiple channels while preserving an adjustable current accuracy.

References [5,6] discuss the implementation of single-inductor multiple-output (Simo) LED drivers for achieving current control and dimming operation. Reference [5] introduces a Simo LED driver structure with time reuse control but is hampered by a limitation on the maximum controllable current and a reduction in the number of LED strings. Reference [6] presents a current source mode Simo LED driver with independent dimming, yet its hard-switched switch tube leads to switching losses. Reference [6] introduces a dual-output LED driver based on coupled inductors, though it poses challenges in determining the load rating, since one of the loads is powered by leakage inductance, despite achieving soft switching. A noteworthy characteristic of these LED drivers is the exclusive use of inductors and memory [5,6].

Literature [7] proposes a dual-output LED driver based on coupled inductance. Although soft switching can be realized, the load rating is not well determined because one of the loads is powered by the leakage sense. These LED drives are characterized



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by the use of inductors, hard switch, dimming control, and other problems. LED drivers commonly employ half-bridge and full-bridge resonant converters, due to their simple design, soft-switching capabilities, and inherent constant current (CC) characteristics. For instance, reference [8] introduces a high-gain circuit topology that integrates buck-boost and half-bridge series resonant converters, enabling the adjustment of output current by changing the frequency of an additional dimming switch. However, the challenge arises when the LED load branch increases, presenting difficulties in dimming control. In contrast, reference [9] presents a variable mode LED driver that combines buck-boost and full-bridge LLCs, achieving three distinct voltage gains by modifying the switch combination on the inverter side but with reduced overall efficiency due to the cascade structure. Additionally, reference [10] suggests a continuous conduction mode (CCM) boost and LLC resonant network integration, but it is limited by its single-stage converter that can only drive one load. Similarly, reference [11] proposes a dimmable LED driver based on an LCLC network, in which the auxiliary switch's buffer capacitor affects output current accuracy while providing soft switching. Moreover, references [12,13] introduce an LED multi-channel current-sharing topology, employing controllable switched capacitors to enable a wide dimming range, albeit at the cost of increasing the number of switches and passive components. Literature [14] deals with a design and experimental verification of a 400 W class light-emitting diode (LED) driver with the cooperative control method for two parallel, connected DC/DC converters. In the cooperative control method, one DC/DC converter is selected to supply the output current for the LED, based on the reference value of the LED current. Thus, the proposed cooperative control strategy achieves wide dimming range operation, but with reduced overall efficiency due to the cascade structure. Literature [15] puts forth a three-switch, double half-bridge, double-output LED driver, utilizing asymmetric duty cycle control to adjust output current but encountering the issue of diode inability to achieve soft switching, leading to significant diode conduction losses.

In order to solve the problems of LED multi-channel current-sharing topology dimming and low efficiency, this paper proposes a four-channel constant proportion dimming LED driver based on an integrated half-bridge resonant switching network. In this network, switches S_1 and S_2 in the half-bridge display complementary conduction, while the controllable switched capacitor unit's switch S_3 and switch S_2 share the same drive signal. The constant frequency pulse width modulation (PWM) controls four current outputs. The converter's working mode and dimming strategy are discussed in detail. Through the utilization of only one common inductor, the driver enables the simultaneous soft switching of switching tubes and diodes, allowing the upper and lower groups of LED current-sharing output units to proportionally adjust light and current.

2. Analysis of Circuit Structure and Working Principle

2.1. Topology

In Figure 1, the integrated half-bridge resonant switching unit comprises switching tubes S_1 and S_2 , common inductance L_s , and resonant capacitors C_{r1} and C_{r2} , as well as controllable switching capacitor units, which include switching tube S_3 and switching capacitor C_{S1} . This unit serves as the foundation for the soft-switching proportional dimming LED driver based on switched capacitors. Moreover, two LED current-sharing output units are depicted in the figure. The first unit consists of freewheeling diodes D_1 – D_4 , filter capacitors C_{O1} – C_{O2} , and output loads LED₁–LED₂, while the second unit encompasses freewheeling diodes D_5 – D_8 , filter capacitors C_{O3} – C_{O4} , and output loads LED₃–LED₄. Together, these components establish the topology of the LED driver, facilitating the implementation of soft-switching proportional dimming for efficient LED control.

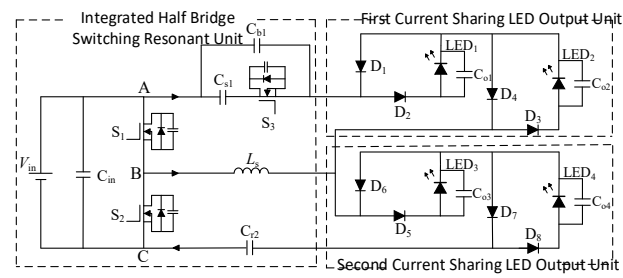


Figure 1. Proposed LED driver circuit diagram.

2.2. Operational Modal Analysis

To simplify the analysis, we assume that all switches, diodes, capacitors, and inductors in the system are ideal components and that the values of resonant capacitors C_{r1} and C_{r2} are lower than those of filter capacitors C_{o1} , C_{o2} , C_{o3} , and C_{o4} , resulting in constant values for $V_{LED1} \sim V_{LED4}$. Figure 2 presents the key waveforms of the proposed converter. The intermediate voltage values of resonant capacitors C_{r1} and C_{r2} are denoted by V_{Cr1} and V_{Cr2} , respectively. Throughout the entire cycle, the converter operates in six modes, each corresponding to an equivalent circuit diagram depicted in Figure 3a through Figure 3f.

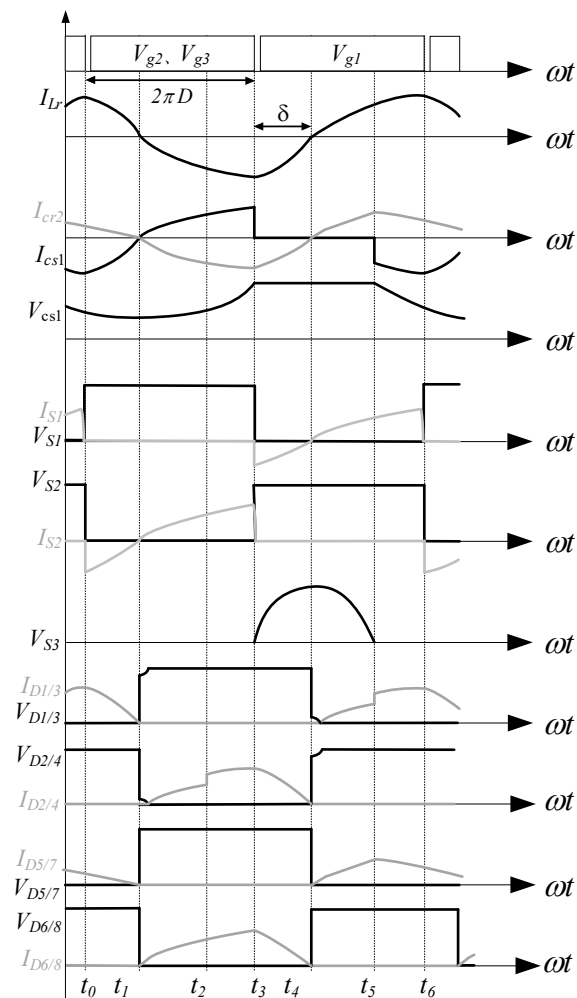


Figure 2. Proposed converter's key waveforms.

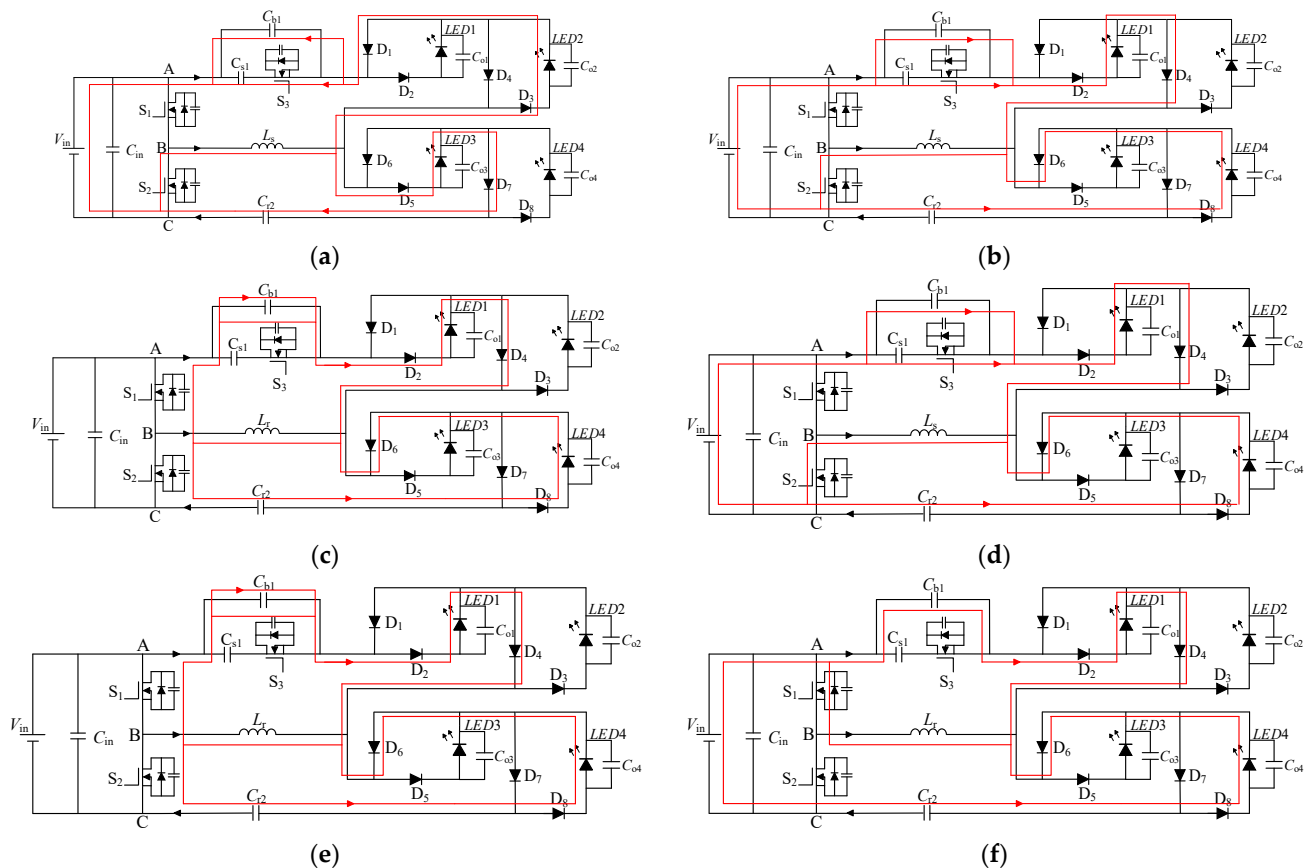


Figure 3. Proposed working modes of the converter: (a) Mode 1; (b) Mode 2; (c) Mode 3; (d) Mode 4; (e) Mode 5; (f) Mode 6.

Mode 1 [$t_0 \sim t_1$]: At time t_0 , when switch tubes S_2 and S_3 are simultaneously turned on and switch tube S_1 remains off, diodes D_1 , D_3 , D_5 , and D_7 conduct in the forward direction, while diodes D_2 , D_4 , D_6 , and D_8 are cut off in the reverse direction. The common inductor current I_{L_S} drops to zero under back voltage, facilitating zero-voltage turn-on conditions by freewheeling through the body diodes of switches S_2 and S_3 . Simultaneously, the adjustable switched capacitor unit and the common inductance L_S undergo series resonant discharging, while the resonant capacitor C_{r2} and the common inductance L_S undergo series resonant charging. When the resonance reaches zero, diodes D_1 , D_3 , D_5 , and D_7 realize zero-current shutdown, terminating the mode.

Mode 2 [$t_1 \sim t_2$]: During this interval represented in Figure 3b, switch tubes S_2 and S_3 remain open, while diodes D_2 , D_4 , D_6 , and D_8 conduct in the forward direction, and diodes D_1 , D_3 , D_5 , and D_7 are cut off in the reverse direction. The common inductor L_S current decreases linearly. The L_S is charged in series with the controllable switched capacitor unit and discharged in series with the resonant capacitor C_{r2} until the sum of the voltages at both ends of C_{r2} and the output load LED_4 matches the sum of the voltages of the controllable switched capacitor unit and the output load LED_1 , signaling the transition to the next mode.

Mode 3 [$t_2 \sim t_3$]: At time t_3 , the instantaneous current of the common inductance L_S drops to the peak, and diodes D_2, D_4, D_6 , and D_8 conduct in the forward direction, while diodes D_1, D_3, D_5 , and D_7 are cut off in the reverse direction. The L_S continues to charge the parasitic capacitance of the controllable switched capacitor unit and the switches S_2 and S_3 and discharges the resonant capacitance C_{r2} and the parasitic capacitance of switch S_1 . When the drain-source voltage of switch S_1 drops to zero, the bulk diode of switch S_1 opens and ushers in mode 4.

Mode 4 [$t_3 \sim t_4$]: Figure 3d illustrates this mode where switch tube S_1 is open, and switch tubes S_2 and S_3 remain off. Diodes D_1 , D_3 , D_5 , and D_7 turn off in reverse, while diodes D_2 , D_4 , D_6 , and D_8 turn on in the forward direction. The common inductor current I_{LS} rises to zero under positive voltage, providing zero-voltage opening conditions through the body diode of switch tube S_1 . The switching capacitor C_{s1} is clamped, while the switching capacitor C_{b1} is mainly charged in series resonance with the common inductance L_S , and C_{r2} is discharged in series resonance with L_S until the resonance reaches zero, leading to zero-current shutdown and the end of the mode.

Mode 5 [$t_4 \sim t_5$]: As depicted in Figure 3e, during this section, switch tube S_1 remains open, diodes D_1 , D_3 , D_5 , and D_7 conduct in the forward direction, and diodes D_2 , D_4 , D_6 , and D_8 are cut off in the reverse direction. The common inductor current L_S increases linearly and starts to discharge C_{r2} and charge C_{b1} , with the switched capacitor voltage V_{CS1} remaining unchanged. When the drain-source voltage of switch S_3 drops to zero, the body diode of switch S_3 opens, transitioning to the next mode.

Mode 6 [$t_5 \sim t_6$]: At time t_6 , the instantaneous current of the common inductance L_S rises to the peak, charging the adjustable switched capacitor unit, the parasitic capacitance of switch S_1 , and the parasitic capacitances of switches S_2 and S_3 . Finally, as diodes D_1 , D_3 , D_5 , and D_7 conduct in the forward direction and diodes D_2 , D_4 , D_6 , and D_8 cut off in the reverse direction, the mode concludes as the drain-source voltage of switch S_2 drops to zero, prompting the body diode to open and re-enter the next switching cycle.

3. Gain Analysis and Dimming Control Strategy

3.1. Circuit Gain Analysis

The resonant network is composed of upper and lower half-bridge switches, accompanied by resonant capacitors sharing a common inductance, denoted as L_S . As a result, both resonant networks are subjected to square-wave input voltage. As depicted in Figure 4, the resonant inductance current, referred to as I_{LS1} , flows from point A to point B, while another resonant inductance current, denoted as I_{LS2} , flows from point C to point B.

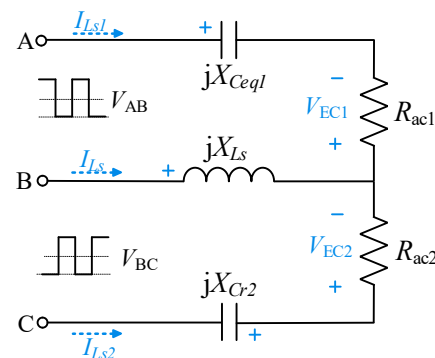


Figure 4. Equivalent circuit of the double-resonant network.

By utilizing fixed-frequency PWM modulation, switches S_1 and S_2 demonstrate complementary conduction with respect to their duty cycles. Based on Fourier series analysis [16], the equivalent input voltage for each phase can be calculated as follows:

$$\begin{cases} V_{AB} = V_{in}/2 + 2V_{in} \sin(Dt)/\pi \\ V_{BC} = V_{in}/2 + 2V_{in} \sin[(1-D)t]/\pi \end{cases} \quad (1)$$

where D represents the duty cycle of switch tube S_1 , while V_{AB} denotes the square-wave voltages generated by the upper bridge arms, and V_{BC} denotes the square-wave voltages generated by the lower bridge arms.

The resonant frequencies of the two resonant networks are different, due to the varying capacitance values of the two resonant capacitors.

$$\begin{cases} f_{r1} = \frac{1}{2\pi\sqrt{L_s C_{eq1}}} \\ f_{r2} = \frac{1}{2\pi\sqrt{L_s C_{r2}}} \end{cases} \quad (2)$$

Now, we will delve into a detailed analysis of the upper half-bridge resonant network. Upon application of the square wave V_{AB} of the half-bridge inverter network to the resonant network comprised of L_s and C_{r1} , it is observed that the basic components of the AC-equivalent output load voltage v_{ec1} and current exhibit sinusoidal characteristics. Following a thorough classic AC analysis, the voltage gain of the AC equivalent circuit of the upper half-bridge resonant network can be expressed as:

$$\frac{V_{EC1}}{V_{AB}} = \frac{R_{ac1} \sin[\pi(1-D)]}{R_{ac1} + j(X_{Ls} - X_{Ceq1})} \quad (3)$$

where X_{Ls} represents the common inductance reactance, X_{Cr1} signifies the capacitance reactance of the upper half-bridge resonant network, and R_{ac1} denotes the AC-equivalent resistance.

$$\begin{cases} X_{Ls} = 2\pi f_s L_s \\ X_{Ceq1} = 1/2\pi f_s C_{eq1} \\ R_{ac1} = 8R_{eq1}/\pi^2 \end{cases} \quad (4)$$

The LED parallel impedance of the first group of two current-sharing output units, R_{eq1} , is given by R_{LED1}/R_{LED2} . According to Equations (3) and (4), the static gain G_1 of the upper half-bridge resonant network is:

$$G_1 = \frac{V_{LED1} + V_{LED2}}{V_{in}} = \frac{(\pi + 4 \sin \pi D) \sin(\pi D)}{2\pi \sqrt{1 + \left(\frac{X_{Ls} - X_{Cr1}}{R_{ac1}}\right)^2}} \quad (5)$$

The static gain G_2 of the lower half-bridge resonant network is similarly defined and can be expressed as follows:

$$\begin{aligned} G_2 &= \frac{V_{LED3} + V_{LED4}}{V_{in}} \\ &= \frac{[(\pi - 4 \sin \pi(1-D)) \sin \pi(1-D)]}{2\pi \sqrt{1 + \left(\frac{X_{Ls} - X_{Cr2}}{R_{ac2}}\right)^2}} \end{aligned} \quad (6)$$

where the capacitance reactance X_{Cr2} and the AC-equivalent resistance R_{ac2} of the lower half-bridge resonant network are determined by the following equations:

$$\begin{cases} X_{Cr2} = 1/2\pi f_s C_{r2} \\ R_{ac2} = 8R_{eq2}/\pi^2 \end{cases} \quad (7)$$

Additionally, R_{eq2} represents the LED parallel impedance of the second group of two current-sharing output units, with a specific value denoted as R_{LED3}/R_{LED4} .

3.2. Constant-Frequency PWM Dimming

The integrated half-bridge switched resonant unit comprises switches S_1 and S_2 , the common inductance L_s , resonant capacitors C_{r1} and C_{r2} , and the controllable switched capacitor unit, which includes switch S_3 and switched capacitor C_{S1} . Utilizing the complementary conduction of switches S_1 and S_2 and a fixed switching frequency, this study implements constant-frequency PWM to regulate the output currents of the four LEDs. Equation (2) indicates that the LED driver possesses two distinct resonant frequencies. Upon meticulous analysis, the switching frequency f_s of the driver is selected to be near the

maximum resonant frequency, and the output current is adjusted by modifying the duty cycle D of switch S_1 . The relationship between load currents $I_{LED1/2}$ and $I_{LED3/4}$ and duty cycle D is depicted in Figure 5, based on the PSIM 9.1.4 simulation. The maximum load current occurs at $D = 0.4$. Specifically, the load current $I_{LED1/2}$ can continuously vary from 219 mA to 392 mA, while the load current $I_{LED3/4}$ can continuously shift from 118 mA to 219 mA. The discussion proceeds to explore the LED load current of the upper and lower current-sharing output units. By referencing the equivalent circuit diagram in Figure 4, the branch current ratio K is taken into account.

$$K = \frac{|I_{LED1/2}|}{|I_{LED3/4}|} = \frac{R_{ac1} + j(X_{Ls} - X_{Ceq1})}{R_{ac2} + j(X_{Ls} - X_{Cr2})} \quad (8)$$

$$= \sqrt{\frac{R_{ac1}^2 + (X_{Ls} - X_{Ceq1})^2}{R_{ac2}^2 + (X_{Ls} - X_{Cr2})^2}}$$

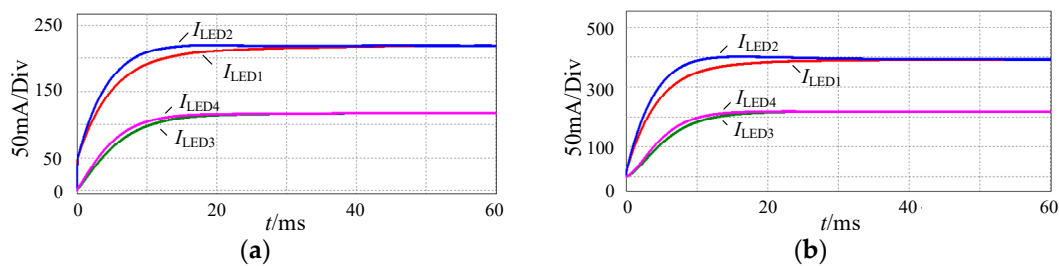


Figure 5. Four load output current simulation waveforms: (a) $D = 0.15$; (b) $D = 0.4$.

In this study, the branch current ratio K can be adjusted by choosing different capacitance values, subsequently affecting the average current in any branch of the first two-way current-sharing output unit ($I_{LED1/2}$) and the second two-way current-sharing output unit ($I_{LED3/4}$). The current values of the two LED current-sharing output units change proportionally by modifying the equivalent capacitance value of the controllable switched capacitor unit. It is essential to highlight that the amount of charge in a switching cycle is directly proportional to the capacitance. As a result, the capacitance change value, C_{sv1} , when adjusting the controllable switched capacitor unit plays a critical role in determining the overall current distribution.

$$C_{sv1} = \frac{\Delta Q_{cs1}}{Q_{cs1}} = \frac{1 - \cos[2\pi(D - \delta/180)]}{2} C_{s1} \quad (9)$$

The equivalent capacitance C_{eq1} of the controllable switched capacitor unit can be ascertained using the following formula, where q_{cs1} symbolizes the charge amount corresponding to capacitor C_{s1} , and D signifies the duty cycle of switch tube S_3 within the unit. Assuming C_{b1} maintains a constant value, the angle of the voltage V_{AB} phase, denoted by θ , precedes the output current I_{LS} angle in this context.

$$C_{eq1} = C_{b1} + C_{sv1} = C_{b1} + \frac{1 - \cos[2\pi(D - \delta/180)]}{2} C_{s1} \quad (10)$$

The aforementioned formula demonstrates that the controllable switched capacitor unit possesses an equivalent capacitance range of $[C_{b1}, C_{s1} + C_{b1}]$, with the maximum duty cycle D not surpassing 0.9. In the steady state, the current is filtered by the rectifier diode and output capacitor according to the driver's operating mode. Further analysis establishes the correlation between the output current and the absolute value I_m of the resonant inductance peak, as follows:

$$I_{LED1} + I_{LED3} = \frac{2|I_{m1} + I_{m2}|}{\pi} \quad (11)$$

The switching loss is calculated as follows:

$$P_{sw} = f_b (\sum E_{on}(I_{on}, T) + \sum E_{off}(I_{off}, T)) \quad (12)$$

Among them, $E_{on}(I_{on}, T)$ is the turn-on loss energy corresponding to the current of a single turn-on process in the positive half-cycle of the chopping of the semiconductor switching device, which is obtained by quadratic interpolation of the junction temperature T and the conduction current I_{on} ; $E_{off}(I_{off}, T)$ is the turn-off loss energy corresponding to the current of a single turn-off process in the positive half-cycle of the chopping of the semiconductor switching device, which is obtained by quadratic interpolation of the junction temperature T and the turn-off current I_{off} ; $E_{on}(I_{on}, T)$ and $E_{off}(I_{off}, T)$ are added, respectively, to obtain the switching loss in a single fundamental period, and switching loss is multiplied in a single fundamental period by the fundamental frequency f_b to obtain the chopper switching loss P_{sw} .

4. Performance Analysis of LED Driver

4.1. Current-Sharing Characteristics among Output Branches

Based on the modal analysis, modes 2 and 6 show that the unidirectional conductivity of the diodes leads to the controllable switched capacitor unit discharging through load LED₁ and freewheeling diodes D₂ and D₄. Within each switching cycle, the quantity of charge stored by the controllable switched capacitor unit, ΔQ_{1ch} , is equivalent to the charge flowing through load LED₁. In a similar manner, modes 1 and 4 reveal that the controllable switched capacitor unit discharges through load LED₂ and freewheeling diodes D₁ and D₃. In this scenario, the amount of charge discharged by the controllable switched capacitor unit during a switching cycle, ΔQ_{1dis} , equals the charge flowing through load LED₂. Owing to the ample filter capacitance, the output current can be considered constant, resulting in the following relationship:

$$\begin{cases} I_{LED1} = \Delta Q_{1ch} / T \\ I_{LED2} = \Delta Q_{1dis} / T \end{cases} \quad (13)$$

The switching cycle, denoted as t , involves the charge amount, ΔQ_{1ch} , ΔQ_{1dis} , of the controllable switched capacitor unit throughout the entire cycle. Additionally, Equation (13) results from the modal analysis of modes 2 and 6, and modes 1 and 4 will not be reiterated here.

$$\begin{cases} I_{LED3} = \Delta Q_{2ch} / T \\ I_{LED4} = \Delta Q_{2dis} / T \end{cases} \quad (14)$$

The amount of charge flowing through the capacitors during the entire switching cycle can be determined using the average currents of LED₃ and LED₄, denoted as I_{LED3} and I_{LED4} , respectively, along with the charge amount of the resonant capacitor C_{r2} (ΔQ_{2ch} , ΔQ_{2dis}). Figure 2 illustrates the charging and discharging processes of the controllable switched capacitor unit and resonant capacitor C_{r2} under different modes. Based on the charge balance principle of the capacitor, the relationship between the charge amounts during the whole switching cycle is established.

$$\begin{cases} \Delta Q_{1dis} = -\int_{t1}^{t4} i_{Ceq1} dt \\ \quad \quad = \Delta Q_{1ch} = \int_{t0}^{t1} i_{Ceq1} dt + \int_{t4}^{t6} i_{Ceq1} dt \\ \Delta Q_{2dis} = \int_{t1}^{t4} i_{Cr2} dt \\ \quad \quad = \Delta Q_{2ch} = -\left(\int_{t0}^{t1} i_{Cr2} dt + \int_{t4}^{t6} i_{Cr2} dt \right) \end{cases} \quad (15)$$

From Equations (13)~(15), it is evident that:

$$\begin{cases} I_{LED1} = I_{LED2} \\ I_{LED3} = I_{LED4} \end{cases} \quad (16)$$

Therefore, the circuit topology can use the charge balance principle of the capacitor to realize the current sharing of two groups of current-sharing output units with four output currents.

4.2. Realization of Soft-Switching Conditions

To analyze the conditions for achieving zero-voltage switching with two switches, we first consider the current flowing through S_1 and S_2 .

$$i_{s1} = \begin{cases} i_{Ls}(t), 0 \leq t < DT \\ 0, DT \leq t < T \end{cases} \quad (17)$$

$$i_{s2} = \begin{cases} 0, 0 \leq t < DT \\ -i_{Ls}(t), DT \leq t < T \end{cases} \quad (18)$$

The condition for soft switching depends on the current flowing through the switch tube. After analysis, the instantaneous value of the resonant inductance current through the upper half-bridge resonant network at any time is:

$$\begin{cases} i_{Ls1}(t) = I_{m1} \sin(w_{s1}t - \alpha_1 - \beta_1) \\ I_{m1} = \frac{[\pi + 4 \sin \pi(1-D)] V_{in}}{Z_{in1}} \end{cases} \quad (19)$$

where Z_{in1} is the input impedance of the first group of current-sharing units.

$$\begin{cases} \alpha_1 = \tan^{-1} \left(\frac{X_{Ls} - X_{Ceq1}}{R_{ac1}} \right) \\ \beta_1 = \tan^{-1} \left[\frac{\sin 2\pi(1-D)}{1 - \cos 2\pi(1-D)} \right] \\ Z_{in1} = R_{ac1} + j(X_{Ls} - X_{ceq1}) \end{cases} \quad (20)$$

Similarly, the instantaneous value of the resonant inductance current flowing through the lower half-bridge resonant network at any time is:

$$\begin{cases} i_{Ls2}(t) = I_{m2} \sin(W_{s2}t - \alpha_2 - \beta_2) \\ I_{m2} = \frac{(\pi - 4 \sin \pi D) V_{in}}{Z_{in2}} \end{cases} \quad (21)$$

where Z_{in2} is the input impedance of the first group of current-sharing units.

$$\begin{cases} \alpha_2 = \tan^{-1} \left(\frac{X_{Ls} - X_{cr2}}{R_{ac2}} \right) \\ \beta_2 = \tan^{-1} \left[\frac{\sin 2\pi D}{1 - \cos 2\pi D} \right] \\ Z_{in2} = R_{ac2} + j(X_{Ls} - X_{cr2}) \end{cases} \quad (22)$$

In the steady-state analysis of the converter, the appropriate choice of the switching frequency f_s for the half-bridge inverter network is vital to ensure that the resonant inductance current I_{Ls} operates in continuous conduction mode (CCM) and achieves zero-voltage switching (ZVS) for both switches [17]. As depicted in Figure 2, the resonant inductor current I_{Ls} peaks at t_3 and t_6 . At this point, the energy stored in the common inductor must be adequate to completely charge and discharge the junction capacitors of switches S_1 , S_2 , and S_3 . Consequently, the ZVS opening conditions for all three switches are established.

$$\begin{cases} (C_{oss1} + C_{oss2} + C_{oss3}) V_{in}^2 < \frac{1}{2} L_s I_{Ls}^2(t_3) \\ (C_{oss1} + C_{oss2} + C_{oss3}) V_{in}^2 < \frac{1}{2} L_s I_{Ls}^2(t_6) \end{cases} \quad (23)$$

C_{oss1} , C_{oss2} , and C_{oss3} represent the junction capacitances of switches S_1 , S_2 , and S_3 .

5. Simulation Analysis and Experimental Verification

5.1. Simulation Analysis

To validate the dual-channel current-sharing features and dimming control strategy of the proposed circuit topology, this study employed PSIM to build a simulation model. The simulation incorporated various impedance values for LED models to determine the current-sharing characteristics under extreme asymmetric loads. The circuit parameters used in the simulation are presented in Table 1. Figures 5 and 6 display the corresponding steady-state waveforms for $D = 0.15$ and $D = 0.4$, offering a visual representation of the simulation outcomes.

Table 1. Parameters and model of the LED driver.

Name	Value or Model
Input voltage (V_{in})	48 V
Common inductance (L_s)	10 μ H
Resonant capacitance ($C_{s1}/C_{r2}/C_{b1}$)	100 nF/68 nF/68 nF
Switching frequency (f_s)	112 kHz
Filter capacitor ($C_{o1} \sim C_{o4}$)	10 μ F
Load	10 Ω
Switching (S_1, S_2, S_3)	NCE0224K
Diode ($D_1 \sim D_8$)	ES5DB

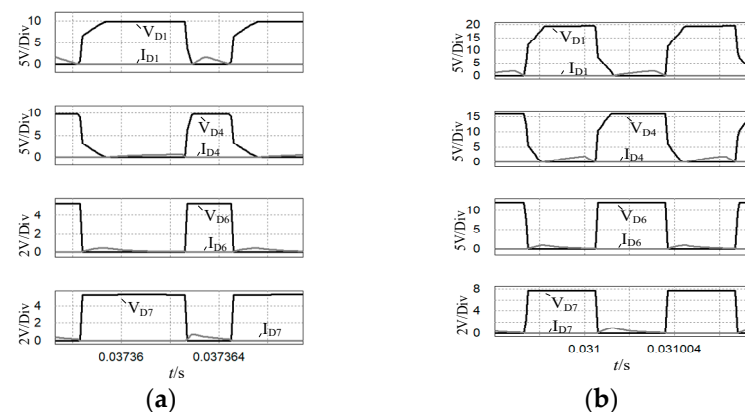


Figure 6. Simulation waveform of diode soft switching. (a) $D = 0.15$; (b) $D = 0.4$.

Figure 5 demonstrates that the amplitude of the steady-state load output current remains consistent for both the upper and lower current-sharing units across various impedance values. Figure 6 illustrates that the diode achieves zero-current switching (ZCS) shutdown when the current at both ends crosses zero and begins to bear the reverse voltage.

5.2. Experimental Verification

To evaluate the feasibility and performance of the proposed work mode, a 19 W experimental prototype was built. The experimental setup employed 3 W LED and included two load cases. In load case 1, the driver operated under rated load, with each of the output loads LED₁, LED₂, LED₃, and LED₄ consisting of 5, 5, 3, and 3 lamp beads, respectively. Furthermore, load case 2 was designed to assess extreme unbalanced load, with the number of lamp beads for each output load adjusted to 5, 4, 4, and 2 for LED₁, LED₂, LED₃, and LED₄, respectively. Figure 7 is one picture with the experimental test.

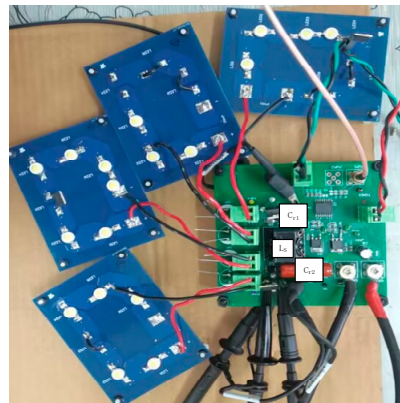


Figure 7. One picture with the experimental test.

The starting process of output current under load condition 2 when $D = 0.4$ is illustrated in Figure 8. It is noteworthy that the starting time is less than 15 ms, and there is no overshoot. Additionally, Figure 9 displays the ripple waveform of the output current under asymmetric load with a duty cycle D of 0.4. Importantly, the current amplitudes of load currents I_{LED1} and I_{LED2} remain stable at 384 mA and 380 mA, respectively. Similarly, the current amplitude of load current I_{LED3} is stable at 210 mA, and that of load current I_{LED4} is stable at 214 mA.

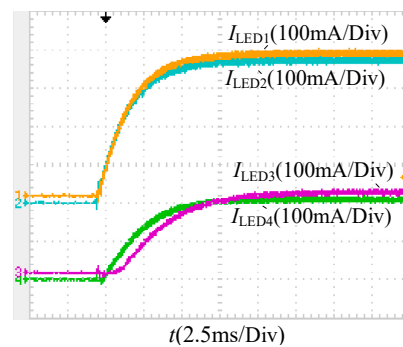


Figure 8. $D = 0.4$, output current startup waveform.

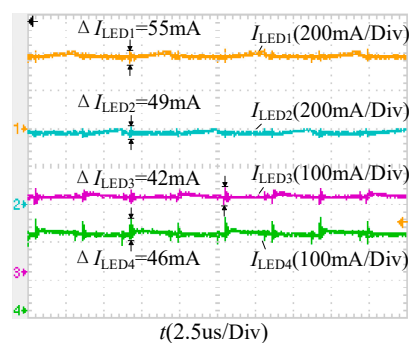


Figure 9. $D = 0.4$, output current ripple at load 2.

At a duty cycle of 0.15, Figure 10 demonstrates the ripple waveform of the four output currents under rated load conditions. The stable current amplitudes for load currents I_{LED1} and I_{LED2} are 188 mA and 198 mA, respectively. In addition, load currents I_{LED3} and I_{LED4} exhibit stable current amplitudes of 104 mA and 108 mA, respectively.

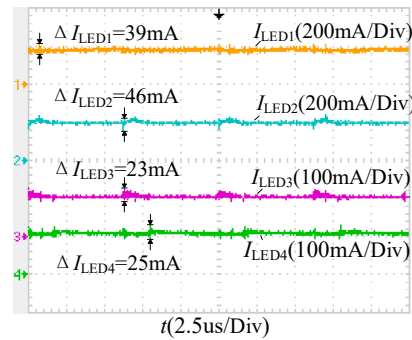


Figure 10. $D = 0.15$; output current ripple at load 1.

The experimental results indicate that the fixed dimming ratio of the two current-sharing output units is approximately 1.9:1, and despite some fluctuation, the output current maintains a high quality. In the context of luminous flux and LED service life, it is crucial to ensure that the LED's current ripple is less than 30% to prevent flickering [18]. Figure 10 illustrates that the maximum output current ripple of the four loads is 23.2%, thus meeting the specified ripple requirements. Furthermore, the amplitudes of the four load currents under two different load conditions align closely with the simulation results shown in Figure 5, as depicted in Figures 9 and 10.

The voltage and current waveforms of two groups of current-sharing units under asymmetric load are displayed in Figure 11. Under different voltage conditions, the output load currents $I_{LED1/2}$ and $I_{LED3/4}$ measure 380 mA and 200 mA, respectively, as shown in the figure. The discrepancies between V_{LED1} and V_{LED2} in Figure 11a and V_{LED3} and V_{LED4} in Figure 11b are attributed to the varying number of LED loads. This observation serves as evidence that the upper and lower current-sharing output units of this driver exhibit favorable current-sharing characteristics.

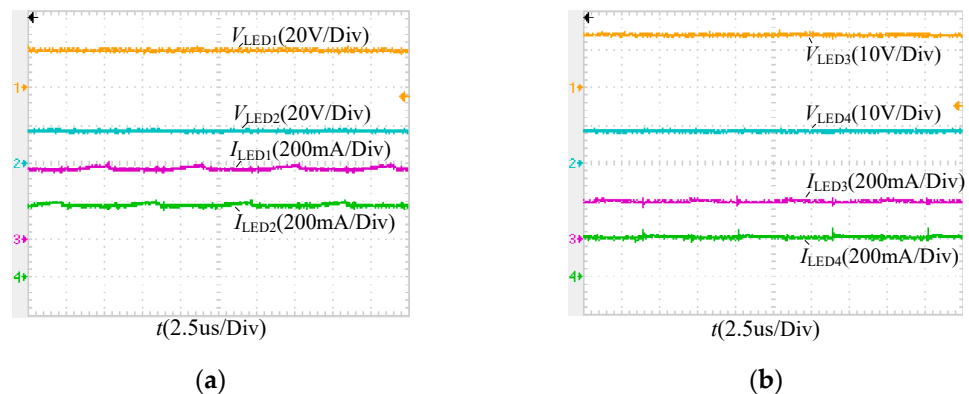


Figure 11. $D = 0.4$, output waveform at load 2. (a) The current-sharing unit in Group I. (b) The second set of units sharing currents.

Additionally, Figure 12 presents the soft-switching waveform of the switch tube when the duty cycle $D = 0.4$, demonstrating that all the switch tubes can achieve ZVS opening. Furthermore, the voltage and current waveforms of diodes are detailed in Figures 13 and 14 under two different load conditions, revealing that the voltage stresses of diodes D_1 and D_4 in Figure 13a are equivalent to the output load voltages v_{led1} and v_{led2} in Figure 11a, while the voltage stresses of diodes D_6 and D_7 in Figure 13b correspond to the output load voltages v_{led3} and v_{led4} in Figure 11b. This equality results in lower voltage stress for the diodes. Additionally, the constant-frequency PWM modulation process indicates that ZCS can be turned off throughout, thereby reducing the reverse recovery loss of the diode.

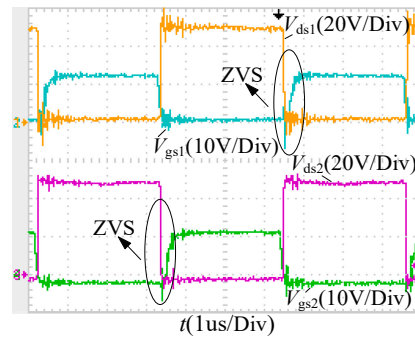


Figure 12. Switch tubes S_1 and S_2 produce a soft-switching waveform.

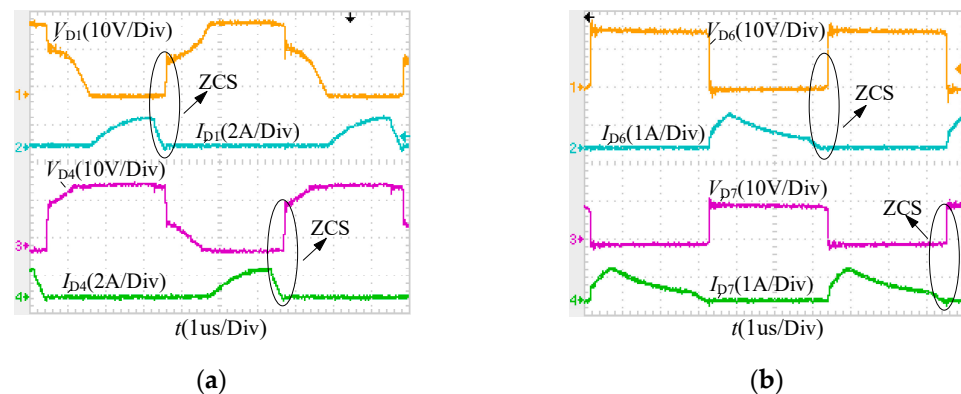


Figure 13. $D = 0.4$, diode soft-switching waveform under load 2. (a) Voltage and current waveforms of diodes D_1 and D_4 . (b) Voltage and current waveforms of diodes D_6 and D_7 .

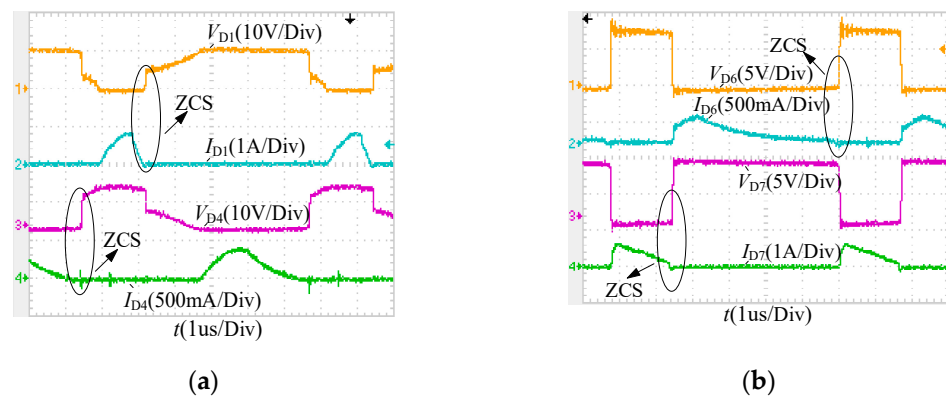


Figure 14. $D = 0.15$, diode soft-switching waveform under load 1. (a) Voltage and current waveforms of diodes D_1 and D_4 . (b) Voltage and current waveforms of diodes D_6 and D_7 .

6. Discussion

In recent years, extensive research has been conducted on multi-channel dimming LED drivers. This study provides a comparison of the number of components, soft-switching capability, output load, and efficiency of these drivers, both domestically and internationally, as shown in Table 2. Notably, the input voltage of references [6,12,15] and the LED driver in this study is 48 V. Upon comparing these parameters, it becomes evident that the LED driver proposed in this paper has fewer switching devices and can accommodate more output loads. As a result, the entire system boasts a higher power density and has the capability to adjust the output current in proportion to the upper and lower current-sharing output units, offering superior soft-switching conditions and overall high efficiency. The efficiency curve obtained from experimental tests conducted on the proposed LED driver is presented in Figure 15, specifically when the current of the first two-way current-sharing output unit falls within the range of 188 mA to 400 mA. Notably, the efficiency is observed

to peak at 380 mA, reaching an impressive 96.56%. This finding underscores the high efficiency and robust performance of the LED driver under varying loads, as demonstrated in the experimental results.

Table 2. LED driver comparison.

	Number of Switch Tubes	Number of Inductance	Number of Capacitance	Number of Load	ZVS	ZCS	Maximum Efficiency
Literature [4]	3	1	2	2	no	no	88%
Literature [5]	2	1	3	2	yes	no	95.5%
Literature [7]	4	2	3	1	yes	no	94.19%
Literature [9]	3	2	3	1	yes	yes	97.8%
Literature [10]	3	1	10	4	yes	no	94.1%
Literature [12]	3	2	6	2	yes	no	91.5%
this paper	3	1	7	4	yes	yes	96.56%

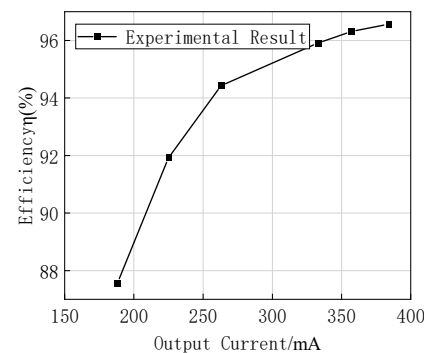


Figure 15. Output current efficiency curve.

7. Conclusions

This paper presents a resonant soft-switching four-way LED output driver designed for multi-color lighting applications. The driver's capability for proportional dimming of two sets of current-sharing output units using only three switches and one inductor is demonstrated through theoretical analysis and experimental prototype verification. The amplitude of the output current can be proportionally adjusted under constant-frequency PWM control by controlling the switched capacitor. Additionally, both sets of current-sharing output units exhibit effective current sharing. Throughout the operation, all switches and diodes achieve soft switching, thereby reducing switching losses and ensuring high efficiency.

In a global context, the benefits of the resonant soft-switching four-way LED output driver designed for multi-color lighting applications could include the following:

1. **Energy efficiency:** The driver's ability to achieve soft switching reduces switching losses, leading to higher efficiency. This is especially valuable in a global context, where energy conservation is a major concern.
2. **Versatile lighting options:** The multi-color lighting application of this driver allows for greater flexibility in creating diverse lighting effects and moods, catering to different global design and aesthetic preferences.
3. **Cost savings:** The use of only three switches and one inductor in the design simplifies the circuit and potentially reduces manufacturing costs. This can have a positive impact on the global market, making the technology more accessible and cost-effective.

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