A Low-Cost Digital PWM-Controlled LED Driver with PFC and Low Light Flicker

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Abstract – This paper proposes an LED driving circuit with a digital controller, power factor correct (PFC) function, and low light flicker. The key topology of the proposed circuit is a conventional Flyback combined with a pre-stage. As a result, there will be less light flicker than with other one-stage PFC circuits. A digital controller, implemented using a low-cost microcontroller, dsPIC30F2020, will meet PFC and low light flicker. The experimental results validate the functionality of the proposed circuit.

Keywords: Digital control, LED driver, PFC, Light flicker

1. Introduction

Digital controllers are currently popular for high-frequency, low-power switching mode power supplies because of their advantages in terms of programmability, different implementation and low sensitivity to variations [1]-[4]. An LED driver controlled by a digital chip allows the control strategy to be user-defined, and therefore the circuit design becomes more flexible and better able to fulfill our target requirement. In the proposed circuit, we program a low-cost microcontroller to achieve closed loop control and Pulse-Width-Modulation (PWM) functions. Compared with the complicated analog circuit that utilizes a comparator and optical coupler to close the voltage and current-control loop, the proposed digital circuit only requires the user to define the voltage and current control by code, which is not only simpler but also reduces the material cost. LED drive circuits often contain power factor correction (PFC) functions. A single-stage topology such as Flyback or boost which is switched by the PWM signal can have PFC function with good power factor. PWM signal could control the switch to make the input current in phase with the input voltage. LED driving using a single-stage PFC circuit often has a huge light flicker, which can be harmful to eyes [5-7]. This flicker is triggered by the twice-line-frequency driving current. According to [8] and [9], another switch added to the secondary side to control the power flow will reduce the capacitor's value. However, this is not suitable for a condition that allows only one switch. Our circuit implements a pre-stage, which helps ensure that the driving-current ripple is low, even with a small-output capacitor. The benefit here, in comparison to the conventional two-stage PFC, is that it only uses one MOSFET. Meanwhile,

Fig. 1. Circuit topology of the proposed single-stage

the output current ripple is much smaller than that of the conventional Flyback with the same output capacitor. Therefore the proposed circuit will lead to a smaller light flicker.

A new LED driver circuit, as shown in Fig. 1, is proposed in this paper. The output current has a low ripple because of the energy-storage capacitor on the primary side. The proposed circuit is compatible with a digital controller. The operating principle of the proposed circuit was analyzed, a theoretical analysis of the output-current ripple was conducted, a design procedure for the proposed LED driver circuit was provided, and a prototype was built to verify performance.

2. Circuit Configuration and Analysis

Fig. 2 demonstrates the operating principles of the proposed circuit. From t_0 to t_1 , the switch Q_1 is on, the coupled inductor is charged, and the primary-side capacitor is discharged. The secondary side diode is biased, as is the current flow from the output capacitor to the LED load. From t_1 to t_2 , on the primary side, the coupled inductor is discharged, and the primary-side capacitor is charged. Current flows through the secondary-side capacitor and charges the output capacitor. From t_2 to t_3 , no current flows

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Fig. 2. Operating principles of the proposed circuit

Fig. 3. System diagram of the proposed circuit

through the primary side. The situation on the secondary side is the same as that of t_1 to t_2 . In the last period, the secondary-side diode is reverse biased. The load current comes only from the output capacitor.

3. Digital Control of the Proposed Driver Circuit

Digital control of the proposed circuit is implemented using the dsPIC30F2020 chip [10]. The system diagram is

shown in Fig.3.

The circuit works with a dual-loop control, with both the voltage regulation and current regulation. Here the dual loop control is different with the common series connected one. The traditional one is taking the voltage error as the reference of the current loop. We use a parallel structure, as shown in Fig.3. The advantage is that implementation is simple. The voltage and current loop will not have interference with each other. The algorithm of control is presented in Fig. 4.

Fig. 4. PI control algorithm flow chart

The PI controller for the voltage loop can be represented as

$$G(S) = 2.2 + \frac{1000}{S} \quad (1)$$

in the continuous domain. Through the bilinear z-transformer [11-17], it can be represented as

$$V_o = V_o Z^{-1} + \left(\frac{4.4 + 1000T}{2} \right) V_i + \left(\frac{1000T - 4.4}{2} \right) V_i Z^{-1} \quad (2)$$

Since $T=10^{-5}$ s, we can get

$$V_o = V_o Z^{-1} + 2.205V_i - 2.195V_i Z^{-1} \quad (3)$$

Therefore, the control algorithm of voltage loop can be written as

$$V_o[n] = V_o[n-1] + 2.205V_i[n] - 2.195V_i[n-1] \quad (4)$$

The current-control loop can also be represented in this way. A similar equation can be derived, such as

$$I_o[n] = I_o[n-1] + 2.205I_i[n] - 2.195I_i[n-1] \quad (5)$$

Then the outputs of the voltage loop and the current loop are compared. The one with the larger value has a larger error, and it should therefore be chosen as the input for the PWM.

4. Analysis of the Current Ripple and Flicker

Methods for analyzing the twice-line-frequency ripple for a PFC circuit are introduced in [18-20]. In this paper,

we found and solved the relationship between the twice-line-frequency ripple and the circuit parameters [18]. We first analyzed the output-voltage ripple. We know the current ripple of the LED is related to the voltage ripple of the LED; therefore, as long as we know the voltage ripple, we can analyze the current ripple based on the I-V curve of the LED. The averaged state-space model over T_s can be built based on the voltage-time balance of the coupled inductor and the Flyback transformer. Eqs. (6) and (7) are derived from the averaged state-space model:

$$C_1 \dot{v}_1(t) = \langle i_{C_1} \rangle = \langle i_{C_1_{in}} \rangle - \langle i_{C_1_{out}} \rangle$$

$$= \frac{1}{2} \frac{v_g^2(t) D^2 T_s}{L_B [v_1(t) - v_g(t)]} - \frac{1}{2} \frac{v_1(t)}{L} D^2 T_s \quad (6)$$

$$C_1 \dot{v}_2(t) = \langle i_{C_2} \rangle = \langle i_{C_2_{in}} \rangle - \langle i_{C_2_{out}} \rangle$$

$$= \frac{1}{2} \frac{v_1^2(t) D^2 T_s}{L v_2(t)} - \frac{v_o(t)}{R} \quad (7)$$

Furthermore, we decomposed each state and input into DC and ripple components. Then, we gathered the first order ripple terms in the differential equations. We did not consider the perturbation of the duty ratio because the circuit was designed to operate with a constant duty ratio and a constant switching frequency. Finally, we get Eq. (8), which describes the perturbations in the primary-side capacitor's voltage and in the output voltage:

$$\begin{pmatrix} \Delta \dot{v}_1(t) \\ \Delta \dot{v}_o(t) \end{pmatrix} = \begin{pmatrix} \frac{-(2V_1 - V_g) D^2 T_s}{2C_1 L (V_1 - V_g)} & 0 \\ \frac{2V_1 D^2 T_s}{2C_2 V_o L} & \frac{1}{-C_2 V_o R} \end{pmatrix} \begin{pmatrix} \Delta v_1(t) \\ \Delta v_o(t) \end{pmatrix} + \begin{pmatrix} \frac{1}{V_1 - V_g} \left(\frac{D^2 T_s V_g}{C_1 L_B} + \frac{D^2 T_s V_g}{C_1 L_B} \right) \\ 0 \end{pmatrix} \Delta v_g(t) \quad (8)$$

The twice-line-frequency ripple components of the output voltage can be solved according to the following steps [18]. First, the equations above can be represented as

$$\Delta \dot{x}_i(t) + \sum_{j=1}^M a_{ij} \Delta x_j(t) = b_{gi} \Delta v_g(t) \quad (i=1, \dots, M) \quad (9)$$

Therefore, the ripple dynamic equations of a certain

ripple can be written with a superposition of sinusoidal functions on the right-hand side:

$$\Delta \dot{x}_i(t) + \sum_{j=1}^M a_{ij} \Delta x_j(t) = b_{Ci}(\omega) \cos \omega t + b_{Si}(\omega) \sin \omega t \quad (10)$$

where $(\omega \in \{\omega_d, 2n\omega_L; n=1,2,\dots\}, i=1,\dots,M)$

$$\begin{pmatrix} b_{Ci} \\ b_{Si} \end{pmatrix} (\omega = \omega_d) = \begin{cases} b_{gi} \begin{pmatrix} v_g^{(m)} \\ 0 \end{pmatrix} & \text{if } \omega = \omega_d = 2m\omega_L \\ 0 & \text{if } \omega = \omega_d \neq 2m\omega_L \end{cases} \quad (11)$$

The solution to Eq. (10) is assumed to have the form

$$\Delta x_i(\omega, t) = x_{Ci}(\omega) \cos \omega t + x_{Si}(\omega) \sin \omega t \quad (12)$$

and can be solved as

$$\begin{pmatrix} x_{C1} \\ \vdots \\ x_{CM} \\ x_{S1} \\ \vdots \\ x_{SM} \end{pmatrix} (\omega) = \begin{pmatrix} A & -\omega I \\ \omega I & A \end{pmatrix}^{-1} \begin{pmatrix} b_{C1} \\ \vdots \\ b_{CM} \\ b_{S1} \\ \vdots \\ b_{SM} \end{pmatrix} \quad (13)$$

where A is

$$\begin{pmatrix} a_{11} & \cdots & a_{1M} \\ \vdots & \ddots & \vdots \\ a_{M1} & \cdots & a_{MM} \end{pmatrix}$$

The inverse of the matrix $\begin{pmatrix} A & -\omega I \\ \omega I & A \end{pmatrix}$ can be found using MATLAB. We neglect the trivial terms of the parameters and retain the dominant ones. Finally, we get the approximation equation for

$$x_{C2} = \frac{V_1 D^2 T_s}{C_2 V_o \omega^2 L} \left(\frac{1}{V_1 - V_g} \right) \left(\frac{D^2 T_s V_g}{C_1 L_B} + \frac{D^2 T_s V_1}{2 C_1 L} \right) v_g^{(m)} \quad (14)$$

Because x_{S2} is much smaller than x_{C2} , the twice-line-frequency ripple's waveform can be approximated as

$$\Delta v_o(t) \approx x_{C2} \cos \omega t \quad (15)$$

Where ω is 2π times the twice-line frequency. According to the energy balance of the Flyback transformer, we can derive the equation

$$P_o = \frac{V_1^2 D^2 T_s}{2L} \quad (16)$$

Based on this equation, Eq. (14) can be re-written as

$$x_{C2} = \frac{2P_o}{C_2 V_1 V_o \omega^2} \left(\frac{1}{V_1 - V_g} \right) \left(\frac{2P_o L V_g}{V_1^2 C_1 L_B} + \frac{P_o}{V_1 C_1} \right) v_g^{(m)} \quad (17)$$

According to the I-V curve for the LED, as shown in Fig.5, we can derive the relationship between the current ripple and the voltage ripple as

$$r_c = \frac{2\Delta i}{I_o} = \frac{2 \left(\frac{X_{C2}}{9} \right)}{0.11V_o - 2.67} = \frac{V_o r_v}{9(0.11V_o - 2.67)} \quad (18)$$

Where r_c is the current ripple rate and r_v is the voltage ripple rate.

Once the current ripple has been estimated from the parameters of the circuit, the intensity of the light flicker can also be estimated. The following is the procedure for deriving the relationship between the current ripple and the two measurements of the light flicker, which are the flicker percentage and the flicker index. Fig. 6 shows the luminous flux waveform of the LED. The flicker index is defined as

$$\text{flicker index} = \frac{\text{area } A}{\text{area } A + \text{area } B} \quad (19)$$

Because the output luminous flux of the LED has an almost sinusoidal shape,

Fig. 5. LED V-I Characteristics

Fig. 6. Waveform of the luminous flux from the LED

$$\frac{y}{\text{area } A} = \frac{2\pi}{2\Delta i} \quad (20)$$

So, Eq. (19) can be expressed as

$$\text{flicker index} = \frac{\text{area } A}{\text{area } A + \text{area } B} = \frac{\text{area } A}{x \left(\frac{\pi \text{area } A}{\Delta i} \right)} = \frac{\Delta i}{x\pi} \quad (21)$$

Because

$$\text{current ripple rate} = \frac{2\Delta i}{x} \quad (22)$$

we get the relationship

$$\text{flicker index} = \frac{\text{percent flicker}}{\pi} = \frac{\text{current ripple rate}}{2\pi} \quad (23)$$

5. Design Procedures of the Proposed Circuit

Fig. 7 demonstrates the design procedures for the proposed circuit. In this paper, we mainly discuss the guidelines for choosing the coupled-inductor value, the transformer's magnetizing inductance value, and the primary-side capacitor's value. We also analyze the voltage and the current stress for switching devices to acquire the knowledge needed for device selection. First, we calculated the Flyback magnetizing inductance. In order to achieve a good power factor, we needed to guarantee that the coupled inductor operates in a discontinuous conduction mode (DCM), as shown in Fig. 8.

Therefore,

$$D + D_f \leq 1 \quad (24)$$

On the other hand, we can get

Fig. 7. Design procedures for this proposed circuit

Fig. 8. Current waveform for the coupled inductor

$$D = \frac{V_o}{V_1(t)} \sqrt{\frac{2L}{T_s R_{load}}} \quad (25)$$

$$D_f = \frac{|V_{in}(t)|}{V_1(t) - |V_{in}(t)|} \quad (26)$$

where R_{load} is the equivalent resistance of the LED, $V_1(t)$ is the primary-side capacitor's voltage, and T_s is the period of the switching cycle.

Substituting Eqs. (24) and (25) into Eq. (26), we get the Flyback magnetizing inductance:

$$L \leq \frac{T_s}{2P_o} [V_1(t) - V_{in}(t)]^2 \quad (27)$$

where P_o is the output power. To simplify the calculation, we approximate $V_1 = V_1(t)$. Therefore, to guarantee that L_B operates in the DCM for a universal input range, we demand that the constraint

$$L \leq \left\{ \frac{T_s}{2P_o} [V_1(t) - V_{in}(t)]^2 \right\}_{\min} \quad (28)$$

be satisfied. Finally, we get

$$L \leq \frac{T_s}{2P_o} [V_1 - V_{in_peak@88V}]^2 \quad (29)$$

where $V_{in_peak@88V}$ is the maximum value of V_{in} when its Root-Mean-Square (RMS) value is 88 V. In our prototype design, we found that $L \leq 0.77\text{mH}$. Here, we choose 0.3 mH. Next, we calculated the coupled inductance. The maximum voltage of the primary capacitor should be set to a certain value. Here, we chose a maximum voltage of 450 V. Through derivation, the primary side capacitor's voltage, V_1 , can be expressed as

$$V_1 = \frac{V_{in_peak}}{2} \left(0.92 + \sqrt{1.92 \left(\frac{L}{L_B} \right) + 0.8464} \right) \quad (30)$$

where V_{in_peak} is the maximum value of V_{in} . Therefore, we got $L/L_B=0.75$. Third, we calculated the maximum duty ratio. For a Flyback transformer, the maximum duty ratio cannot exceed 0.5. For a universal input range, the maximum duty ratio occurs at 88 V. Because

$$D = \frac{V_o}{V_1} \sqrt{\frac{2L}{T_s R_{load}}} \quad (31)$$

we found that $D=0.25$. Fourth, we calculated the turns ratio of the Flyback transformer. To guarantee that L operates in the DCM, we can use

$$L \leq \frac{T_s R_{load} n^2}{2} \left(1 - \frac{nV_o}{nV_o + V_1} \right)^2 \quad (32)$$

Therefore, we calculated the turns ratio as $n = 3$. Fifth, we decided on the capacitance of the primary side capacitor. We fixed the output capacitor at a certain capacitance. In our prototype, considering the package size, we chose the value to be 1880 μF , and we aimed to achieve a current ripple of less than 15%. According to Eq. (18), we can estimate that a current ripple under 15% means a voltage ripple under 3.93%. On the other hand, with averaged state-space modeling, we derive

$$X_{C_2} = \frac{2P_o}{C_2 V_1 V_o \omega^2} \left(\frac{1}{V_1 - V_2} \right) \left(\frac{2P_o L V_g}{V_1^2 C_1 L_B} + \frac{P_o}{V_1 C_1} \right) v_g^{(m)} \quad (33)$$

Where ω is two times the twice-line frequency. Based on Eq. (33), we established that the primary-side capacitor should have a capacitance of at least 10 μF . As to the existence of a high switching-frequency ripple, the value of C_1 should be chosen to be larger than the same critical value. Sixth, we found the voltage and current stress of the switching device. Within the analysis, we obtain

$$V_{sw_max} = nV_o + V_1@264V \quad (34)$$

$$V_{d1_max} = nV_o + V_1@264V \quad (35)$$

$$V_{d2_max} = V_1@264V \quad (36)$$

$$V_{d3_max} = V_o + \frac{V_1@264V}{n} \quad (37)$$

$$i_{sw_max} = \left(\frac{V_{in_peak@88V}}{L_B} + \frac{V_1}{L} \right) DT_s \quad (38)$$

$$i_{d1_max} = \left(\frac{V_{in_peak@88V}}{L_B} \right) DT_s \quad (39)$$

$$i_{d2_max} = i_{d1_max} \quad (40)$$

$$i_{d3_max} = \left(\frac{V_1@88V}{nV_o} \right) i_{sw_max} \quad (41)$$

where V_{sw_max} , V_{d1_max} , V_{d2_max} , and V_{d3_max} are the maximum voltages of the switch Q_1 and of the diodes D_1 , D_2 , and D_3 ; i_{sw_max} , i_{d1_max} , i_{d2_max} , and i_{d3_max} are the maximum currents of the switch Q_1 and of the diodes D_1 , D_2 , and D_3 ; and $V_1@264V$ shows the voltage of V_1 when the input voltage is 264 V.

6. Simulation Verification

Our simulation verified that the current-ripple rate in the simulation matched our value calculated on the derived equations very well. The simulation results are shown in

Fig. 9. Simulation result of output current ripple versus input AC voltage

Fig. 9. Our simulation was designed for various input AC voltages. The simulation parameters were as follows: output voltage $V_o=33$ V, output power $P_o=33$ w, $C_1=10$ uF, $C_2=1880$ uF, $L=0.3$ mH, and $L/L_B=0.75$.

7. Experimental Verification

Fig. 10 shows a picture of the hardware prototype we built based on the design guidelines. Fig. 11 shows details

Fig. 10. Hardware prototype of the proposed circuit

Fig. 11. Details of the proposed circuit board

Fig. 12. Experimental results of the conventional Flyback with 110V input voltage

Fig. 13. Experimental results of the proposed circuit with 110V input voltage

Fig. 14. Experimental results of the conventional Flyback with 220V input voltage

Fig. 15. Experimental results of the proposed circuit with 220V input voltage

8. Conclusion

We provide a design guideline for the digital-controlled LED driver. An experimental prototype was built based on the design guideline. The experimental results verified that the proposed circuit had a lower current-ripple rate than the conventional Flyback circuit and that it had a power factor above 0.90.

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