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, 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 t₀ to t₁, the switch Q₁ is on, the coupled inductor is charged, and the primary-side capacitor is discharged. Current flows through the secondary-side capacitor and charges the output capacitor. From t₁ to t₂, no current flows...
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.
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Fig. 4. PI control algorithm flow chart

The PI controller for the voltage loop can be represented as

\[ G(S) = 2.2 + \frac{1000}{S} \] (1)

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

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

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

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

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

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

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

\[ I_o[n] = I_o[n-1] + 2.205I_i[n] - 2.195I_i[n-1] \] (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 \( T_i \) 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:

\[ C_1 v_1(t) = \frac{1}{2} \frac{v_i^2(t)}{L} - \frac{1}{2} \frac{v_i(t)}{L} D^2 T_s \] (6)

\[ C_2 v_2(t) = \frac{1}{2} \frac{v_i^2(t)}{L} - \frac{v_i(t)}{R} \] (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:

\[ \begin{pmatrix} \Delta V_1(t) \\ \Delta V_o(t) \end{pmatrix} = \begin{pmatrix} \frac{-2V_i - V_o}{2C_i L (V_i - V_o)} & 0 \\ \frac{2V_i D^2 T_s}{2C_i V_o L} & -\frac{2V_i D^2 T_s}{C_i V_o L} \end{pmatrix} \begin{pmatrix} \Delta V_1(t) \\ \Delta V_o(t) \end{pmatrix} \] (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

\[ \Delta x_i(t) + \sum_{j=1}^{M} a_{ij} \Delta x_j(t) = b_{ig} \Delta v_g(t) \quad (i=1,...,M) \] (9)

Therefore, the ripple dynamic equations of a certain
ripple can be written with a superposition of sinusoidal functions on the right-hand side:

\[
\Delta x_j(t) = \sum_{j=1}^{M} a_j \Delta x_j(t) = b_{C_i}(\omega) \cos \omega t + b_{S_i}(\omega) \sin \omega t
\]  
(10)

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

\[
\begin{align*}
&b_{C_i}(\omega) = \omega \left( \frac{V_g}{b} \right) & \text{if } \omega = \omega_d = 2m\omega_d \\
&0 & \text{if } \omega = \omega_d \neq 2m\omega_d
\end{align*}
\]  
(11)

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

\[
\Delta x_i(\omega, t) = x_{C_i}(\omega) \cos \omega t + x_{S_i}(\omega) \sin \omega t
\]  
(12)

and can be solved as

\[
\begin{pmatrix}
x_{C_1} \\
\vdots \\
x_{C_M} \\
x_{S_1} \\
\vdots \\
x_{S_M}
\end{pmatrix}
(\omega) =
\begin{pmatrix}
A & -\omega I^3 \\
\omega I & A
\end{pmatrix}
^{-1}
\begin{pmatrix}
b_{C_1} \\
\vdots \\
b_{C_M} \\
b_{S_1} \\
\vdots \\
b_{S_M}
\end{pmatrix}
\]  
(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

\[
x_{C_2} = \frac{V_iD^2TS}{C_2V_iV_g} \left( \frac{1}{V_i - V_g} \right) \left( \frac{D^2T_gV_g}{C_iL_g} + \frac{D^2T_gV_g}{2C_iL_g} \right) v^{(m)}_g
\]  
(14)

Where \( \omega \) is \( 2\pi \) times the twice-line frequency. According to the energy balance of the Flyback transformer, we can derive the equation

\[
P_o = \frac{V_o^2D^2T}{2L}
\]  
(16)

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

\[
x_{C_2} = \frac{2P_o}{C_2V_iV_g} \left( \frac{\frac{1}{V_i - V_g}}{V_i^2C_iL_g} + \frac{P_o}{V_iC_i} \right) v^{(m)}_g
\]  
(17)

According to the I-V curve for the LED, as shown in Fig.5, we can derive the relationship between the twice-line frequency ripple's waveform can be approximated as

\[
\Delta V_o(t) \approx x_{C_2} \cos \omega t
\]  
(15)
Fig. 6. Waveform of the luminous flux from the LED

\[
\frac{y}{\text{area } A} = \frac{2\pi}{2\Delta i} \tag{20}
\]

So, Eq. (19) can be expressed as

\[
\text{flicker index} = \frac{\text{area } A}{\text{area } A + \text{area } B} = \frac{\text{area } A}{\pi \text{area } A} = \frac{\Delta i}{\pi} \tag{21}
\]

Because

\[
\text{current ripple rate} = \frac{2\Delta i}{x} \tag{22}
\]

we get the relationship

\[
\text{flicker index} = \frac{\text{percent flicker}}{\pi} = \frac{\text{current ripple rate}}{2\pi} \tag{23}
\]

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,

\[
D + D_f \leq 1 \tag{24}
\]

On the other hand, we can get

\[
D = \frac{V_o}{V_1(t)} \sqrt{\frac{2L}{T_s R_{load}}} \tag{25}
\]

\[
D_f = \left| \frac{V_{in}(t)}{V_1(t) - V_{in}(t)} \right| \tag{26}
\]

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:

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

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

\[
\text{calculate the magnetizing inductance of Flyback transformer and the coupled inductance}
\]

\[
\text{calculate the maximum duty ratio of the switch}
\]

\[
\text{calculate the turn ratio of the Flyback transformer}
\]

\[
\text{select the value of primary side capacitor } C_1 \text{ and output capacitor } C_2
\]

\[
\text{design the coupled inductor and Flyback transformer}
\]

\[
\text{calculate the maximum voltage and current of the switches for selecting devices}
\]
\[
L \leq \frac{T_c}{2P_o} \left[ V_{i}^2(t) - V_{io}(t) \right] \quad (28)
\]
be satisfied. Finally, we get
\[
L \leq \frac{T_c}{2P_o} \left[ V_{i}^2 - V_{i_{\text{peak}}(88\text{v})}^2 \right] \quad (29)
\]
where \( V_{i_{\text{peak}}(88\text{v})} \) is the maximum value of \( V_i \) 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, \( V_i \), can be expressed as
\[
V_i = \frac{V_{i_{\text{peak}}}}{2} \left[ 0.92 + \sqrt{1.92 \left( \frac{L}{L_p} \right) + 0.8464} \right] \quad (30)
\]
where \( V_{i_{\text{peak}}} \) is the maximum value of \( V_i \). Therefore, we got \( L/L_p=0.75 \). 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
\[
D = \frac{V_o}{V_i} \sqrt{\frac{2L}{T_sR_{\text{load}}}} \quad (31)
\]
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
\[
L \leq \frac{T_sR_{\text{load}}}{2} \left[ 1 \cdot \frac{nV_o}{nV_o + V_i} \right] \quad (32)
\]
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 \( \mu \text{F} \), 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
\[
X_{C_2} = \frac{2P_o}{C_2V_iV_o} \left[ \frac{1}{V_i - V_2} \left( \frac{2P_o L_p}{V_i V_2} + \frac{P_o}{V_i C_1} \right) \right] \quad (33)
\]
where \( \omega \) 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 \( \mu \text{F} \). 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
\[
V_{\text{sw}_{\text{max}}} = nV_o + V_i @ 264\text{V} \quad (34)
\]
\[
V_{d1_{\text{max}}} = nV_o + V_i @ 264\text{V} \quad (35)
\]
\[
V_{d2_{\text{max}}} = V_i @ 264\text{V} \quad (36)
\]
\[
V_{d3_{\text{max}}} = \frac{V_o + V_i @ 264\text{V}}{n} \quad (37)
\]
\[
i_{\text{sw}_{\text{max}}} = \frac{V_{i_{\text{peak}}(88\text{v})} + V_i}{L_p} DT_s \quad (38)
\]
\[
i_{d1_{\text{max}}} = \frac{V_{i_{\text{peak}}(88\text{v})}}{L_p} DT_s \quad (39)
\]
\[
i_{d2_{\text{max}}} = i_{d1_{\text{max}}} \quad (40)
\]
\[
i_{d3_{\text{max}}} = \left( \frac{V_i @ 88\text{V}}{nV_o} \right) i_{\text{sw}_{\text{max}}} \quad (41)
\]
where \( V_{\text{sw}_{\text{max}}}, V_{d1_{\text{max}}}, V_{d2_{\text{max}}}, \) and \( V_{d3_{\text{max}}} \) are the maximum voltages of the switch \( Q_1 \) and of the diodes \( D_1, D_2, \) and \( D_3 \); \( i_{\text{sw}_{\text{max}}}, i_{d1_{\text{max}}}, i_{d2_{\text{max}}}, \) and \( i_{d3_{\text{max}}} \) are the maximum currents of the switch \( Q_1 \) and of the diodes \( D_1, D_2, \) and \( D_3 \); and \( V_i @ 264\text{V} \) shows the voltage of \( V_i \) 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](image-url)
7. Experimental Verification

Fig. 10 shows a picture of the hardware prototype we built based on the design guidelines. Fig. 11 shows details of the circuit board. Fig. 12 and Fig. 13 present the experimental results of the conventional Flyback LED driver and proposed LED driver with 110V input voltage. Fig. 14 and Fig. 15 present the experimental results of the conventional Flyback LED driver and proposed LED driver with 220V input voltage. From the results, we can conclude that the proposed LED driver circuit has a much lower current ripple than that of the conventional circuit. Meanwhile, the power factor of the proposed circuit is above 0.90, which is considered to be good.
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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References


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